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A DIRECT SEQUENCE SPREAD SPECTRUM

TIMING BEACON

by

William R. Seed, B. Eng.

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

Master of Engineering.

Faculty of Engineering
Department of Electronics
Carleton University

May 1986

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ISBN 0-315-29852-9

~~The undersigned recommend to the faculty of Graduate~~
Studies and Research acceptance of the thesis :

"A DIRECT SEQUENCE SPREAD SPECTRUM
TIMING BEACON"

Submitted by William Seed in partial
fulfillment of the requirements for the
degree of Master of Engineering

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May 1986

ABSTRACT

A satellite downlink spread spectrum timing beacon system has been proposed, built and tested. The timing beacon is required to provide the users of a multiple access system with a timing reference. A 300 bps data signal is spread with a 1 MHz direct sequence pseudorandom code to produce a 1 μ sec timing mark at the receiver. Code acquisition is achieved using serial search synchronization techniques. The received signal is despread using a delay-locked loop. A squaring loop with frequency searching capabilities is used for data demodulation. The recovered data and timing mark essentially enable the users to obtain the satellite time. The ability to achieve both a 1 μ sec timing accuracy and a high system reliability for effective input SNR's below -20 dB is demonstrated.

ACKNOWLEDGEMENTS

The author would like to express his gratitude to his supervisor, Professor J.S. Wight, for his continual encouragement and friendly guidance. His drive and dedication have provided the necessary inspiration to allow me to complete the thesis in a relatively short time frame.

I would also like to acknowledge the many thoughtful comments of Dr. LeRoy Pearce. Invaluable discussions with Dr. Pearce help to guide much of the system strategies and configurations. Without the facilities and equipment available at the Communications Research Center of the Department of Communications, this work would not have been possible.

Mr. Doug Lambert and Mr. Al McEwen have been extremely helpful, providing numerous practical suggestions. Their willingness to contribute to the overall design is gratefully acknowledged.

I would like to thank the Space Systems office staff - Mary Fulton, Jill Weitzel and Wanda Dolan - for the expediency in which equipment was procured and the thesis typed. Their efforts are truly appreciated.

Financial support provided by the Department of National Defence Research Branch under Project No. 32A59 has made possible the procurement of vital system components, for which the author is grateful.

Finally, I would like to thank my wife, Wendy, for her encouragement, patience and understanding during the completion of this endeavour.

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GLOSSARY OF TERMS

A signal amplitude
 A_r, A_t receive and transmit antenna areas
 $b(\cdot)$ noise discriminator characteristic
 B bandwidth
 C carrier power
 $d(t)$ data signal
 $d_f(t)$ filtered data signal
 D dither offset
 D_m data modulation distortion
 D_o directivity
 $e(t)$ error signal
 E signal energy
 E_b bit energy
 E_{sig} normalized signal voltage
 $E_{rms/sig}$ root-mean-square of the signal voltage
 $E_{rms/noise}$ rms noise voltage
 $E[\cdot]$ expected value of
 f_D doppler offset
 f_o VCO central frequency
 $f(t)$ impulse response of loop filter
 $F(p)$ loop filter transfer function
 $g(\cdot)$ tracking loop discriminator characteristic
 $g_n(\cdot)$ normalized tracking loop discriminator characteristic
 G dc tracking loop gain
 G_n normalized tracking loop gain
 $h_1(t)$ impulse response of equivalent lowpass filter
 $H_1(p)$ equivalent lowpass filter transfer function
 k loop gain
 k_D detector gain, data modulation/filter degradation
 k_L filter degradation
 k_i ($i=1,2,3, VCO, VCC$) sensitivity or component gain
 L_n noise intermodulation loss
 $n(t)$ white gaussian noise
 n_c, n_s filtered noise cross receiver generated signal processes
 $n'(t)$ equivalent VCO modulation due to noise
 N noise power
 N_o noise spectral density
 $N_c(t), N_s(t)$ statistically independent, lowpass filtered AWG noise processes
 P code period, also signifies d/dt
 $p(\cdot)$ probability
 P power
 $PN(t)$ pseudonoise sequence
 P_{fa} false alarm probability
 P_d detection probability
 Q_i state i of shift register, latch, etc...
 R autocorrelation function
 R_b bit rate
 R_c chip rate
 R_s symbol rate

$s(t)$ transmitted or receiver generated signal
 $S_d(f)$ data modulation spectrum
 S_L squaring loss
 $S_{PN}(w)$ PN spectrum
 SNR signal-to-noise ratio
 T period of data symbol, chip, false alarm, etc...
 TW time-bandwidth product
 u_1, u_2 error signals
 v_{out} detector output voltage
 v_T threshold voltage
 ω angular frequency
 ω_n natural angular frequency
 x_E, x_L, x_P crosscorrelated early, late and punctual signals
 x_E sum of crosscorrelated early and late signals
 x_A difference between crosscorrelated early and late signals
 $y(t)$ output signal
 $\delta()$ Dirac delta function
 ϵ PN code timing error
 $\phi()$ random variable
 ξ damping constant
 $\theta(t)$ signal phase
 ψ VCO output phase
 ψ_0 mean-square of the noise voltage
 σ standard deviation
 ρ effective SNR in a given bandwidth
 Ω angular frequency offset or range
 λ free-space wavelength
 Λ triangular function
 τ_d propagation delay
 $\hat{\tau}_d$ receiver estimate of propagation delay

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CHAPTER 1

INTRODUCTION

1.1 INTRODUCTION

The many favorable attributes of spread spectrum techniques have made them an integral part of modern military satellite communication systems [1]-[3]. In this thesis a spread spectrum timing beacon designed to provide a users uplink timing reference for a multiple access satellite communications system is described. A low data rate signal is spread with a pseudorandom direct sequence code to produce a 1 μ sec timing mark for the users.

The initial application of spread spectrum techniques was to minimize the effects of jamming in military communication systems. Not only has the use of spread spectrum techniques increased in modern military communication systems, but there is also a growing demand for civilian applications. Global positioning systems (GPS), mobile radio and NASA's tracking and data relay satellite system (TDRSS) are examples of current non-military uses of spread spectrum techniques. The benefits achieved, often simultaneously, using such techniques include

- interference rejection
- secure communications
- low probability of intercept
- low spectral density
- multiple access capabilities

- high resolution range measurements
- high resolution timing reference

It is the high resolution timing reference capability of the spread spectrum technique that is of particular interest in this thesis.

1.2 MULTIPLE ACCESS SYSTEMS

There are three main forms of multiple access systems. These are time division multiple access (TDMA), frequency division multiple access (FDMA) and code division multiple access (CDMA), also referred to as spread spectrum multiple access (SSMA). The users of TDMA systems have access to the entire bandwidth but not all the transmitting time. FDMA system users can usually transmit continuously but are restricted to a bandwidth that is much less than that of the transponder. CDMA system users can occupy all the transponder bandwidth and transmit continuously due to the low crosscorrelation between the users. While TDMA, FDMA and CDMA systems all offer the same theoretical communications capacity, only the CDMA system offers an interference rejection capability.

Timing synchronization for multiple access systems is often a prime concern. Since bursts of information are transmitted in TDMA systems, timing synchronization among the users is required. Timing synchronization among the users of an FDMA system can also be an important specification. The requirement for an accurate timing reference for CDMA systems also exists, particularly where the user uplink signals are to be processed onboard the satellite.

1.3 THESIS OBJECTIVES

For the reception or demodulation of the user CDMA signals onboard, a very accurate users' uplink timing reference is required. A spread spectrum timing beacon that could provide the required 1 μ sec timing accuracy and satisfy the transponder power limitations was to be investigated. A low data rate signal (300 bps) supplies the earth stations with time of day and orbital information. By spreading the data with a 1 MHz direct sequence pseudonoise code, a timing accuracy of better than 1 μ sec can be achieved for the tracked code. Yet, the required transponder transmit power need only be marginally higher than would normally be used for a 300 bps signal. The time of day information, also referred to as the "satellite time", will be stored at the receiver in some form of buffer. The receiver clock is then set accordingly and waits for a timing mark (a pseudonoise code vector) to start the clock. Hence, the earth station will essentially have the satellite time upon reception of the code timing mark. This method is analogous to the "time at the sound of the beep" concept. The timing beacon design was thus to address the problems of low SNR pseudorandom code synchronization and data demodulation under a relatively wide doppler uncertainty.

1.4 THESIS OUTLINE

Included in Chapter 2 is an introduction to the various spread spectrum techniques, with emphasis on direct sequence modulation. Analyses of pseudonoise code acquisition and tracking schemes are presented. Relevant means for data demodulation of the despread signal are also introduced. Rather than presenting only the background material necessary for the various subsystem configurations that are actually used in the timing beacon design, an attempt is made to provide a description of the competing technologies and techniques.

In Chapter 3 the system requirements and proposed solutions are described in detail. Areas of analysis include the code length, code repetition rate, code acquisition time and associated statistics, code tracking jitter, carrier reference jitter and bit error rate of the recovered data.

In Chapter 4 the experimental results obtained with the spread spectrum timing beacon are presented. The performance of the code acquisition loop, code tracking loop, carrier synchronization loop and the system bit error rate are all observed and compared with the theoretical predictions.

In Chapter 5 a synopsis of the experimental results is presented. Suggestions for improving the system performance are discussed. Also included are suggestions for future work related to this project.

CHAPTER 2

SPREAD SPECTRUM SYSTEM CONCEPTS

2.1 INTRODUCTION

There are many aspects that must be addressed when contemplating receiver configurations. An understanding of the various subsystem performances and their interrelations is essential for system design. Indeed, the choice or necessity of particular schemes often impose bounds on other system performances. For example, a pseudorandom code period of T_{cr} seconds bounds the minimum code tracking loop filter bandwidth to be greater than $1/T_{cr}$ Hertz. Furthermore, the time to acquire the PN code and techniques applied are also often dependent on T_{cr} . The purpose of this chapter is to present a number of analyses of such system considerations as direct sequence generation and spectral characteristics, pseudorandom code acquisition and tracking, and coherent demodulation.

2.2 SPREAD SPECTRUM TECHNIQUES

The essential idea of spread spectrum (SS) is to modulate a data signal with a pseudorandom or pseudonoise (PN) code such that the

resulting bandwidth is much greater than that of the information alone. The spectrum spreading can be realized by a variety of techniques (direct sequence, frequency hopping, time hopping, chirp and hybrids of the preceding) for a variety of reasons (secure communications, low probability of intercept, interference rejection, low spectral density, high resolution range measurements, and multiple access capabilities). The receiver crosscorrelates the spread signal with a local replica of the code. This effectively collapses the signal energy back into its original information bandwidth. Consequently, the receiver must be more complex - it must generate a replica of the code and keep it time-aligned with the aid of a tracking loop.

2.2.1 Direct Sequence

Direct sequence ss is achieved by digitally mixing a data stream with a PN sequence (see Figure 2.1) whose "chip" rate R_C is much greater than the symbol rate R_S . The data $(\sin x/x)^2$ spectrum is spread by what is referred to as the processing gain,

$$PG = R_C/R_S$$

(2.2-1)

The processing gain gives the theoretical upper limit on the SNR improvement from the input spread signal to the demodulated data stream.

Figure 2.1 : Direct sequence modulated data.

The direct sequence waveform is obtained using shift registers and the appropriate feedback taps. A maximal length (ML) code has a period length of $2^n - 1$ where n is the number of stages in the shift register. A list of ML sequences is given by Holmes [2] and Dixon [1]. A more thorough presentation on other types of PN generators, both linear and nonlinear, is found in Volume 1 of [3]. Multiple access systems make use of the superior crosscorrelation properties of Gold codes which are formed from 2 or more ML sequences.

PN Signal Power Spectrum

The power spectral density of the PN signal $S_{PN}(w)$ can be derived by first noting that the power spectrum is the Fourier transform of the autocorrelation function $R_{PN}(\tau)$.

$$S_{PN}(w) = \int_{-\infty}^{\infty} R_{PN}(\tau) e^{-jw\tau} d\tau \quad (2.2-2)$$

where τ is the time displacement of the code correlated with itself.

Consider first the autocorrelation function of the rectangular pulse, the familiar triangular function

$$\Lambda_{T_C}(\tau) = \begin{cases} 1 - \frac{|\tau|}{T_C} & , \quad |\tau| \leq T_C \\ 0 & , \quad \text{elsewhere} \end{cases} \quad (2.2-3)$$

Strictly speaking, the autocorrelation of the PN sequence is not zero for $|\tau|$ greater than a chip; in fact, the autocorrelation function for a ML PN sequence of period p is given as [2]

$$R_{PN}(\tau) = -\frac{1}{p} + \frac{p+1}{p} \Lambda_{T_C}(\tau) * \sum_{m=-\infty}^{\infty} \delta(\tau + pmT_C) \quad (2.2-4)$$

where $*$ represents the convolution operation. Taking the Fourier transform,

$$S_{PN}(w) = \frac{1}{p^2} \delta(f) + \frac{p+1}{p} \sum_{\substack{m=-\infty \\ m \neq 0}}^{\infty} \text{sinc}^2 \left(\frac{m\pi}{p} \right) \delta\left(f + \frac{m}{pT_C}\right) \quad (2.2-5)$$

where $\delta(f)$ is the Dirac delta function. The autocorrelation function closely approximates the triangular function for a reasonably long sequence.

Figure 2.2 : (a) Autocorrelation of a PN sequence and (b) its power spectral density.

It should be observed that the spectrum of the PN sequence alone is composed of discrete spectral lines separated by $1/(pT_c)$. If the separation is minimal, an envelope of the spectra is observed. But the power spectrum of the PN sequence exclusive-OR'ed with random data will not consist of discrete lines; instead, each PN spectral line effectively appears as a carrier for the data $(\sin x/x)^2$ spectrum. Since the line separation must be much less than the symbol rate R_s the resulting power spectrum will indeed be the envelope of the spread $(\sin x/x)^2$ spectrum.

2.2.2 Other Spreading Techniques

While the focus of this thesis is on direct sequence ss communications, a few words on the remaining ss techniques is warranted.

Frequency Hopping

A frequency hopping (FH) system employs the use of a variable frequency generator as the local oscillator for the data stream. Extremely wide bandwidths are possible with FH systems; state of the art frequency synthesizers have bandwidths of the order of 4 GHz. The carrier frequency is chosen by a PN sequence, resulting in a "hopping" throughout the band, as shown in Figure 2.3. Fast hoppers correspond to many hops per symbol while a slow hopper corresponds to several symbols per hop.

Maintaining phase coherence with an FH system is particularly difficult. Consequently, the trend has been to use binary or M-ary FSK with noncoherent detection. Note that the fast hopper will also suffer a noncoherent combining loss since the data energy is summed from a number of hops.

Figure 2.3 : A basic FH system

Chirp

Chirp spreading, shown in Figure 2.4, is achieved by modulating the data signal onto a carrier that is swept (not necessarily linearly) across a band. The chirp system makes use of a steadily increasing or decreasing carrier as opposed to FH systems which use discrete carriers. Chirp waveforms are more commonly found in radar pulse compression techniques, although their use for communications does exist. The direction of the chirp (i.e. up or down) usually indicates the data polarity.

Figure 2.4 : Chirp spread spectrum.

Time Hopping

Time hopping (TH) does not increase the bandwidth of the information signal as do the previous techniques. TH systems have bursts of data followed by silence as determined by a PN code. The main advantage of TH systems is the reduced duty cycle. The coded information symbols would also use interleaving in a pulse jamming environment. Because of the burst nature of the signal both the transmitter and receiver require some form of buffer storage. TH is often combined with an FH system, an example being the RACEP systems developed by Martin Marietta Corporation [1].

2.3 PN CODE ACQUISITION

There are numerous techniques and technologies available for direct sequence code acquisition. Current technologies used for PN matched filters include SAW correlators, SAW convolvers, digital, binary analog, acousto-optic convolvers/correlators, CCD correlators and SAW chirp transform filters. Colvin [5] and Grieco and Rappaport [6] summarize current technologies, trends and present state of the art performances and future expectations from these devices. Despite the impressive results of CCD correlators [6] RCA, Reticon, and Fairchild, the industry leaders, are no longer actively pursuing CCD correlators. As an alternative to the matched filter (MF) technique, serial synchronization can provide a less expensive approach to acquisition, although the time to acquire lock is usually an order of magnitude more than in the MF approach. The serial acquisition method is achieved by clocking the receiver PN code faster than the transmitter clock until the crosscorrelated result indicates that the codes are aligned. Another search method is to mix the received spread signal with the receiver generated code at the same clock frequency, integrate the crosscorrelated result for a given dwell period and then make a decision on whether to continue the integration or advance the receiver code by one-half chip.

The methods considered in this section are for direct sequence PN code acquisition; however, frequency hopping acquisition is analogous in many respects to direct sequence techniques. The essential difference is that phase acquisition is required for direct sequence while frequency acquisition is required for FH.

Irrespective of the particular technique chosen the common element is the large time-bandwidth (TB) product required. A large TB product is necessary to make a positive identification of the low SNR input signal. The matched filter accomplishes this by correlating a stored code segment with the incoming noise corrupted signal. The MF output yields a high correlation when the incoming PN sequence is identical to the stored code segment. The result is an effective integration of the despread signal equal to the MF length, such that the MF output SNR will be increased to an acceptable level for detection. The MF makes a chip by chip correlation and must therefore accept the noise in the spread bandwidth. The serial approach differs in this respect; the incoming and receiver generated signals are first crosscorrelated and then narrowband filtered. When the codes are more than a chip apart the narrowband filter output is virtually that of noise; however, when the two codes "pass" each other the signal energy is collapsed back into the information bandwidth and is passed by the narrowband filter. The narrowband filtering results in a large decrease in the noise power and, consequently, a large increase in SNR.

The choice of acquisition scheme is usually determined by the system requirements or expected environment. In addition to the maximum TB product or device processing gain, the dynamic range, programmability, size, power consumption and cost are also important considerations.

2.3.1 The Matched Filter

The normalized impulse of a matched filter given a signal $s(t)$, $0 \leq t \leq T$ is defined as [7]

$$h(\tau) = \begin{cases} s(T-\tau) & 0 \leq \tau \leq T \\ 0 & \text{elsewhere} \end{cases} \quad (2.3-1)$$

It follows that for the coherent channel the MF output of the signal plus noise is given by

$$y(t) = \int_0^T s(T-\tau)s(t-\tau)d\tau - \int_0^T s(T-\tau)n(t-\tau) d\tau \quad (2.3-2)$$

The first integral is the autocorrelation function of the signal and, hence, peaks at $t=T$. The MF output signal-to-noise ratio SNR_0 can be expressed in terms of (2.3-2) as

$$SNR_0 = \frac{\{E_S[y(T)] - E_{NS}[y(T)]\}^2}{\text{var } [y(T)]} \quad (2.3-3)$$

where $E[]$ denotes the statistical expectation, var is the variance and the subscripts S and NS represent signal present and signal absent, respectively.

Assuming the noise to be a stationary process and, furthermore, that the noise is additive, white and gaussian (AWG) with a double-sided spectral density of $N_0/2$, equation (2.3-3) can be reduced to

$$\begin{aligned} \text{SNR}_0 &= \frac{E}{\int_{-\infty}^{\infty} N_0/2 |S(f)|^2 df} \\ &= \frac{2E}{N_0} = 2TW(\text{SNR}_i) \end{aligned} \quad (2.3-4)$$

where

$$E = \int_0^T S^2(t) dt$$

is the signal energy, $S(f)$ is the Fourier transform of $s(t)$ and the effective input signal to noise ratio is $\text{SNR}_i = P_i/N_i$. The input signal power is $P_i = E/T$ and the noise power in a W Hz bandwidth is $N_i = N_0W$. If the direct sequence modulated signal uses 180° PSK then only one correlator is required, as shown in Figure 2.5. The linear increase

in SNR by a factor of 2TW provides the necessary motivation for MF techniques. Unfortunately, the likelihood of the coherent channel is remote, particularly when considering ss communications. In light of this observation, consideration of the noncoherent channel is necessary.

Figure 2.5: A correlation detector for PSK signals.

A baseband noncoherent matched filter for binary signals requires correlation on both the I and Q channels due to the unknown phase of the incoming signal. To implement a digital matched filter (DMF), shown in Figure 2.6, a sampler operating at least at the Nyquist rate is used to feed the digitizer. The level of quantization seldom exceeds four and is more commonly the one-bit digitizer shown in Figure 2.6. Because of the 180° phase ambiguity of the carrier, the I and Q channel correlator outputs are separately squared and then summed. For the case of $SNR_i \ll 1$ and $SNR_o \gg 1$, the DMF output SNR for one-bit quantization and AWG noise is

$$SNR_O = \frac{2}{\pi} \frac{TW}{2} SNR_i \quad (2.3-5)$$

The output for a noncoherent analog matched filter (AMF) for the same SNR_i and SNR_O case is given as

$$SNR_O = \frac{TW}{2} SNR_i \quad (2.3-6)$$

These results agree with Turin's results given in [7]. The noncoherent AMF suffers a 6 dB SNR_O degradation over the coherent channel while the DMF has an additional quantization loss. The result is well known and deduced in [7] as approximately 2 dB (i.e. $10\log(\pi/2)=1.96$ dB).

A higher level of quantization will naturally lower this degradation; however, each additional level requires another complete set of multipliers, summers, A/D converters, and D/A converters for correlation. The growth of the LSI, VLSI and VHSIC technologies have made possible complete DMF realizations on only a few IC's. TRW has a 64-bit digital correlator with a clocking speed of up to 20 MHz*. While current digital correlators perform only one-bit quantization (making it necessary to have parallel correlators for higher level quantization), a DMF with multibit quantization on a single chip is expected in the near future [5].

* See the TRW LSI Products Catalog for the TDC1023J monolithic digital correlator.

same as for the coherent channel case. For example, a SAW TDL correlator with a bandwidth of 60 MHz and a 60 μ sec processing capability offers a maximum processing gain of $40 \log[(60 \times 10^6)(60 \times 10^6)] = 35.5$ dB. A two-beam SAW acousto-optic time integrating correlator has recently been developed for wideband signals [8]. This new architecture for performing real-time correlation has achieved TB products exceeding 10^6 (34 MHz x 30 ms) and has several attributes which make it particularly well suited for use as a ss signal processor.

Figure 2.6 : A noncoherent DMF.

Figure 2.7 : Serial search technique.

2.3.2 Serial Search

The basic configuration for the serial search approach is illustrated in Figure 2.7. The receiver generated code is correlated with the incoming spread signal, narrowband filtered and then compared with some predetermined threshold voltage. Once the threshold has been exceeded a tracking loop is employed to control the receiver clock for the PN code.

Of paramount importance for acquisition is the interrelationship between the false alarm probability (P_{fa}), probability of detection (P_d) and the filter output SNR. Consider first the case of a single pulse radar detection receiver as shown in Figure 2.8. A pulse of duration T_p is transmitted. A portion of this pulse is reradiated from

a target back to the receiver with a carrier power of C . The received signal is passed through a narrowband filter whose bandwidth is B_{IF} and then envelope detected.

Figure 2.8 : Single pulse radar detection.

The noise entering the IF filter is assumed to be gaussian with a probability density function (pdf) given by

$$p(v) = \frac{1}{\sqrt{2\pi\psi_0}} \exp\left(-\frac{v^2}{2\psi_0}\right) \quad (2.3-7)$$

where v is the noise voltage and ψ_0 is the mean square value of the noise voltage. The pdf following the envelope detector can be shown [9] to obey

$$p(R) = \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) \quad (2.3-8)$$

where R is the envelope amplitude.

The false alarm probability is defined as the likelihood of the noise voltage exceeding the threshold voltage V_T ,

$$P_{fa} = \int_{V_T}^{\infty} \frac{R}{\psi_0} \exp\left(-\frac{R^2}{2\psi_0}\right) dR$$

$$= \exp\left(-\frac{V_T^2}{2\psi_0}\right) \quad (2.3-9)$$

Equation (2.3-9) may be manipulated to give the required threshold voltage in terms of the mean square noise voltage and false alarm probability,

$$V_T = \sqrt{2\psi_0 \ln(1/P_{fa})} \quad (2.3-10)$$

The average time between false alarms is given by Skolnik [9] as

$$T_{fa} = \frac{1}{P_{fa} B_{IF}} = \frac{1}{B_{IF}} \exp\left(\frac{V_T^2}{2\psi_0}\right) \quad (2.3-11)$$

When a sinewave signal of amplitude A is present with the noise at the BPF, the output from the envelope detector has a pdf given by

$$P_S(R) = \frac{R}{\psi_0} \exp\left(-\frac{R^2 + A^2}{2\psi_0}\right) I_0\left(\frac{RA}{\psi_0}\right) \quad (2.3-12)$$

where $I_0(z)$ is the modified Bessel function of zero order with argument z . The probability of detection is

$$P_d = \int_{-\infty}^{\infty} \frac{R}{\psi_0} \exp\left(-\frac{R^2 + A^2}{2\psi_0}\right) I_0\left(\frac{RA}{\psi_0}\right) dR \quad (2.3-13)$$

which cannot be evaluated by simple means, although approximations do exist.

The signal voltage to rms noise voltage ratio can be related to the more familiar signal power to noise power ratio by noting that

$$\frac{A}{\psi_0^{1/2}} = \frac{\text{signal amplitude}}{\text{rms noise amplitude}} = \frac{\sqrt{2} (\text{rms signal voltage})}{\text{rms noise voltage}} = \left(\frac{2S}{N}\right)^{1/2}$$

Hence, one obtains

$$S/N = A^2/(2\psi_0) = C/(\psi_0 B_{IF})$$

where C is the power in the carrier over the duration of the pulse. The P_{fa} , P_d and SNR relationships are reproduced from Skolnik in Figure 2.9 for a sinewave in noise.

Now consider the PN synchronization scheme in light of the preceding radar detection process. For synchronization, the autocorrelation pulse to be detected is described by $E_{sig}(t) = R_{PN}(\tau) \sqrt{2C} \cos \omega t$ where τ is the timing offset between the incoming and receiver generated codes. The maximum signal output of a filter matched to the input pulse occurs at $\tau=0$ and can be shown [10] to be given by $E_{sig}/\max = E/\sqrt{C}$ where E is the

Figure 2.9 : P_{fa} , P_d , SNR relationship for a sinewave in noise
(copied from [9])

Figure 2.10 : Envelope detector output for (a) radar pulse and
(b) triangular autocorrelation function.

energy in the incoming pulse. This represents the peak signal voltage. The rms carrier at $\tau=0$ is written as $E_{rms}/sig = E/(\sqrt{2C})$. The energy in the autocorrelation pulse is given by

$$E = \int_{-T}^T [C^{1/2}(1-|\tau|/T_p)]^2 d\tau = \frac{2}{3} CT_p \quad (2.3-14)$$

where T_p is the time to search one chip length. Consequently $E_{rms}/sig = \sqrt{2CT_p}/3$. The rms noise voltage at the MF output can be shown [10] to be given by $E_{rms}/noise = \sqrt{\psi_0 T_p/3}$. The MF output is envelope detected. The SNR at this point is described by

$$SNR_O = \frac{(E_{rms}/sig)^2}{(E_{rms}/noise)^2} = \frac{2}{3} \frac{T_s C}{\psi_0}$$

Assuming the $B_{IF} = 1/T_s$, a 1.75 dB decrease (i.e., $10 \log [2/3]$) in SNR is observed at the peak of the autocorrelation pulse when compared to that of the radar pulse. Since the noise statistics are the same for either case, the curves generated by Skolnik in Figure 2.9 may be used for PN serial synchronization applications as long as it is remembered that the PN synchronization pulse suffers a 1.75 dB degradation in SNR. For example, a 10 dB SNR in Figure 2.9 with a corresponding combination of P_{fa} and P_d can be realized for PN synchronization if the SNR of the despread signal is 11.75 dB following the filter. Sage [10] has analyzed serial synchronization of PN sequences and developed P_{fa} , P_d , SNR curves which agree with Skolnik's results when the degradation factor of 1.75 dB is incorporated.

Figure 2.11 (a) Single dwell PN code acquisition and (b) double dwell synchronization.

An alternative means of achieving serial synchronization is realized with a single or double dwell search scheme, shown in Figure 2.11. An integration of τ_D seconds is performed on the crosscorrelated signal. If the integration falls below the threshold, the PN code generator is advanced by, typically, one-half chip. If the threshold is exceeded, integration continues. The search is terminated when the PN codes are aligned. A false alarm, however, results in a longer integration time before the code generator is updated. Holmes [2] presents a thorough

analysis of the P_{fa} , P_d statistics for single and double dwell serial acquisition.

2.4 PN CODE TRACKING LOOPS

Once the acquisition circuits have detected the presence of the spread signal with the location of the PN code known to within a chip, a supporting circuit must be invoked to accurately track the PN code. The need to collapse all the signal energy back to the information bandwidth necessitates a close scrutiny of the PN code tracking loop performance.

J. J. Spilker, using results of W. J. Gill, first proposed a tracking device for binary PN sequences in 1963 [11]. Since that time numerous papers have appeared, supplying a more detailed analysis of various aspects of the loop. A handful of different code tracking implementations, most notably the less hardware intensive time-sharing loop or tau-dither loop, have also emerged. While the implementations may differ in the amount of hardware required and the manner in which they are used, they are all based on the same philosophy. In section 2.2.1 it was observed that the autocorrelation function of the PN sequence can be closely approximated by the triangular function. The crosscorrelation of the incoming signal with the receiver generated advanced and delayed versions of the PN sequence can be differenced to form a discriminator characteristic curve as shown in Figure 2.12. Note that $\tilde{x}_E^2$ and $\tilde{x}_L^2$ represent the narrowband filtered, envelope detected crosscorrelation of the early and late signals, respectively. While the example used in Figure 2.12 is for the "one-chip" delay-locked loop

(DLL); the discriminator characteristic for other implementations is easily construed. Each configuration has its own advantages and disadvantages in terms of tracking jitter, sensitivity to gain/phase imbalances, search rate, pull-in range and hardware simplicity.

The negative SNR associated with a received spread spectrum signal usually dictates that carrier synchronization be performed on the despread signal (i.e. following the acquisition and tracking circuits). With this in mind, the analysis is directed toward noncoherent PN code tracking loops.

Figure 2.12 : Discriminator characteristic for a code tracking loop.

2.4.1 The Delay-Locked Loop

Consider the noncoherent DLL given in Figure 2.13. The receiver PN generator generates three signals: a punctual or on-time signal and an early and late signal that are displaced in time from the punctual signal by equal amounts. Only the early and late signals are used by the DLL for tracking. The incoming signal is crosscorrelated with the early and late signals, with the outputs denoted as x_E and x_L , respectively. Since the signal energy has been essentially collapsed back to the information bandwidth at this point, the correlated signals can be passed through a narrowband filter - the filter bandwidth need only be wide enough to pass the data plus any doppler. The filtered crosscorrelation signals are then envelope detected and differenced to provide a discriminator signal. The discriminator signal is given a gain of k_2 to achieve an overall loop gain of k and then filtered with the loop filter. The output of the loop filter controls the voltage controlled clock.

Figure 2.13 : The delay-locked loop.

If the early and late signals are displaced from the punctual signal by one-half chip, the DLL is referred to as a "one-chip" DLL. While this may be the most common delay, shorter or longer delays can also be used. Varying the delay from the punctual signal affects the DLL's pull-in range and tracking sensitivity.

The received signal $r(t)$ is the sum of the signal $s(t)$ plus additive bandpass noise $n(t)$,

$$s(t) = \sqrt{2P_s} PN(t-\tau_d) d(t-\tau_d) \cos[w_0 t + \theta(t)] \quad (2.4-1)$$

$$n(t) = \sqrt{2} \{ N_C(t) \cos[w_0 t + \theta(t)] - N_S(t) \sin[w_0 t + \theta(t)] \} \quad (2.4-2)$$

The average signal power is denoted as P_s , $PN(t - \tau_d)$ is the received PN sequence with transmission delay τ_d , $d(t - \tau_d)$ is the baseband data for the same delay, w_0 is the carrier angular frequency and $\theta(t) = \theta_0 + \Omega_0 t$ is the unknown carrier phase. The noise processes $N_C(t)$ and $N_S(t)$ are assumed to be statistically independent, stationary, lowpass filtered white gaussian noise processes with a single-sided spectral density of N_0 W/Hz and a one-sided bandwidth of B_H with $B_H \ll w_0/2\pi$.

The signals generated at the receiver include versions of the early signal $s_E(t)$, the late signal $s_L(t)$ and the punctual signal $s_P(t)$ where

$$s_E(t) = \sqrt{2} PN(t - \hat{\tau}_d + T_c/2) \cos[(\omega_0 + \omega_1)t + \phi] \quad (2.4-3)$$

$$s_L(t) = \sqrt{2} PN(t - \hat{\tau}_d - T_c/2) \cos[(\omega_0 + \omega_1)t + \phi] \quad (2.4-4)$$

$$s_P(t) = \sqrt{2} PN(t - \hat{\tau}_d) \cos[(\omega_0 + \omega_1)t + \phi] \quad (2.4-5)$$

The DLL's estimate of the transmission delay is denoted as $\hat{\tau}_d$.

Considering only the difference frequency terms the crosscorrelator signal $x_E(t)$ becomes

$$\begin{aligned} x_E(t) = & \sqrt{2P_S} k_1 d(t - \tau_d) \overline{PN(t - \tau_d) PN(t - \hat{\tau}_d + T_c/2)} \cos[\omega_1 t + \psi(t)] \\ & + \sqrt{2P_S} k_1 d(t - \tau_d) \cdot [PN(t - \tau_d) PN(t - \hat{\tau}_d + T_c/2) \\ & - \overline{PN(t - \tau_d) PN(t - \hat{\tau}_d + T_c/2)}] \cos[\omega_1 t + \psi(t)] + n_E(t) \end{aligned} \quad (2.4-6)$$

The phase detector gain is k_1 , the overhead bar represents the statistical expectation, $\psi(t) = \phi(t) - \phi$ and the noise term is given as

$$n_E(t) = k_1 PN(t - \hat{\tau}_d + T_c/2) n(t) \quad (2.4-7)$$

The correlator output signal consists of the desired signal, a correlation self-noise term and a noise term.

For a large PN code period p , the statistical expectation becomes the autocorrelation triangle as discussed in Section 2.2.1.

$$R_{PN/E}(\epsilon/T_C) = PN(t-\tau_d)PN(t-\hat{\tau}_d+T_C/2)$$

$$= \begin{cases} 0 & \frac{\epsilon}{T_C} < -\frac{3}{2} \\ \frac{3}{2} + \frac{\epsilon}{T_C} & -\frac{3}{2} \leq \frac{\epsilon}{T_C} \leq -\frac{1}{2} \\ 1 - \frac{\epsilon}{T_C} & -\frac{1}{2} \leq \frac{\epsilon}{T_C} \leq \frac{1}{2} \\ 0 & \frac{\epsilon}{T_C} > \frac{1}{2} \end{cases} \quad (2.4-8)$$

The transmission delay error is defined as $\epsilon = \tau_d - \hat{\tau}_d$.

If we choose the loop bandwidth to be much smaller than the chip rate (this is certainly the case since B_L must be much less than the data rate!) then we can neglect any self-noise effects [12]. The resulting crosscorrelated early signal is thus given to a good approximation by

$$x_E(t) = \sqrt{2P_S} k_1 d(t-\tau_d) R_{PN/E}(\epsilon/T_C) \cos[w_1 t + \psi(t)]$$

$$+ \sqrt{2} \{ N_{CE}(t) \cos[w_1 t + \psi(t)] - N_{SE}(t) \sin[w_1 t + \psi(t)] \}$$

(2.4-9)

The signal is then passed through a narrowband filter. Denoting the impulse response of the bandpass filter as $h(t)$ and its lowpass equivalent as $h_1(t)$, the filtered output is written as [12]

$$Y_E(t) = \sqrt{2P_s} k_1 d_f(t - \tau_d) R_{PN/E}(\epsilon/T_C).$$

$$\cos [w_1 t + \psi(t)] + n_{Ef}(t) \quad (2.4-10)$$

where

$$d_f(t) = \int_{-\infty}^{\infty} d(u) h_1(t-u) du \quad (2.4-11)$$

$$n_{Ef}(t) = \sqrt{2} k_1 n_{cE}(t) \cos[w_1 t + \psi(t)] - \sqrt{2} k_1 n_{sL}(t) \sin[w_1 t + \psi(t)] \quad (2.4-12)$$

$$n_{cE}(t) = \int_{-\infty}^{\infty} N_{cE}(u) PN(u - \hat{\tau}_d + T_C/2) h_1(t-u) du \quad (2.4-13)$$

$$n_{sE}(t) = \int_{-\infty}^{\infty} N_{sE}(u) PN(u - \hat{\tau}_d - T_C/2) h_1(t-u) du \quad (2.4-14)$$

The filtered crosscorrelation of the received signal and the late signal follows a similar analysis and is given by.

$$y_L(t) = \sqrt{2P_s} k_1 d_f(t - \tau_d) R_{PN/L}(\epsilon/T_C) \cos [w_1 t + \psi(t)] + n_{Lf}(t) \quad (2.4-15)$$

where

$$n_{Lf}(t) = \sqrt{2} k_1 n_{cL}(t) \cos[w_1 t + \psi(t)] - \sqrt{2} k_1 n_{sL}(t) \sin[w_1 t + \psi(t)] \quad (2.4-16)$$

$$n_{cL}(t) = \int_{-\infty}^{\infty} N_{cL}(u) PN(u - \hat{\tau}_d - T_C/2) h_1(t-u) du \quad (2.4-17)$$

$$n_{SL}(t) = \int_{-\infty}^{\infty} N_{SL}(u) PN(u - \hat{T}_d - T_C/2) h_1(t-u) du \quad (2.4-18)$$

$$R_{PN/L} \left(\frac{\epsilon}{T_C} \right) = R_{PN/E} \left(\frac{\epsilon}{T_C} - 1 \right) \quad (2.4-19)$$

Note also that due to the periodicity of the PN code,

$$R_{PN/E/L} \left(\frac{\epsilon}{T_C} \right) = R_{PN/E/L} \left(\frac{\epsilon}{T_C} + np \right), \quad n = \pm 1, \pm 2, \pm 3, \dots \quad (2.4-20)$$

Ignoring the double frequency terms produced by the squaring, the error signal may be written as

$$\begin{aligned} e(t) &= y_L^2(\epsilon/T_C) - y_E^2(\epsilon/T_C) \\ &= P_s d_f^2(t - \hat{T}_d) g(\epsilon/T_C) + n_e(t, \epsilon/T_C) \end{aligned} \quad (2.4-21)$$

where

$$g\left(\frac{\epsilon}{T_C}\right) = R_{PN/L}^2\left(\frac{\epsilon}{T_C}\right) - R_{PN/E}^2\left(\frac{\epsilon}{T_C}\right) \quad (2.4-22)$$

is the discriminator characteristic curve given by

$$g\left(\frac{\epsilon}{T_C}\right) = \begin{cases} 0 & , \quad \epsilon/T_C \leq -3/2 \\ -(3/2 + \epsilon/T_C)^2 & , \quad -3/2 \leq \epsilon/T_C \leq -1/2 \\ 2(\epsilon/T_C) & , \quad |\epsilon/T_C| \leq 1/2 \\ (3/2 - \epsilon/T_C)^2 & , \quad 1/2 \leq \epsilon/T_C \leq 3/2 \\ 0 & , \quad \epsilon/T_C \geq 3/2 \end{cases} \quad (2.4-23)$$

Also, it is again noted that

$$g\left(\frac{\epsilon}{T_C}\right) = g\left(\frac{\epsilon}{T_C} + np\right), \quad n = \pm 1, \pm 2, \pm 3, \dots \quad (2.4-24)$$

The normalized discriminator characteristic must have a slope of one for small timing errors. It is given as

$$g_n\left(\frac{\epsilon}{T_c}\right) = 1/2 g\left(\frac{\epsilon}{T_c}\right) \quad (2.4-25)$$

The equivalent additive noise $n_e(t, \epsilon/T_c)$ is given by Simon [12] as

$$\begin{aligned} n_e\left(t, \frac{\epsilon}{T_c}\right) = & N_{CL}^2(t) - N_{CE}^2(t) + N_{SL}^2(t) - N_{SE}^2(t) / \\ & + 2/P_S d_f(t - \tau_d) [R_{PN}/L(\epsilon/T_c) N_{CL}(t) \\ & - R_{PN}/E(\epsilon/T_c) N_{CL}(t)] \end{aligned} \quad (2.4-26)$$

The receiver's instantaneous estimate of the propagation delay τ_d can be written in terms of the error signal as

$$\frac{\hat{\tau}_d}{T_c} = k_{VCC} k_2 \frac{F(p)}{P} e(t) \quad (2.4-27)$$

where k_{VCC} is the gain of the voltage controlled clock, k_2 is a voltage gain, $F(p)$ is the loop filter transfer function and p denotes d/dt .

Letting $k = k_1 k_2 k_{VCC}$ and combining the previous equations,

$$\frac{\hat{\tau}_d}{T_c} = k \frac{F(p)}{P} [P_S d_f^2(t - \tau_d) g(\epsilon/T_c) + n_e(t, \epsilon/T_c)] \quad (2.4-28)$$

or,

$$\frac{\epsilon}{T_c} = \frac{\tau_d}{T_c} - \frac{kF(p)}{P} [P_S d_f^2(t - \tau_d) g(\epsilon/T_c) + n_e(t, \epsilon/T_c)] \quad (2.4-29)$$

The degradation in signal power due to the squaring of the filtered data is now considered. The term $d_f^2(t - \tau_d)$ can be separated into its mean value and a modulation self-noise term as follows,

$$d_f^2(t-\tau_d)' = \overline{d_f^2(t-\tau_d)} + [d_f^2(t-\tau_d) - \overline{d_f^2(t-\tau_d)}] \quad (2.4-30)$$

where $\langle \rangle$ signifies the time average. If the loop filter is chosen much less than the data symbol rate then the modulation self-noise may be neglected [12]. The reduction in signal power due to the filter distortion on the data modulation can hence be expressed as

$$D_m = \overline{d_f^2(t-\tau_d)} = \int_{-\infty}^{\infty} S_d(f) |H_1(j2\pi f)|^2 df \quad (2.4-31)$$

Equation (2.4-29) may then be written as

$$\frac{1}{T_c} \frac{d\epsilon}{dt} = \frac{1}{T_c} \frac{d\tau_d}{dt} - k_F(p) [2P_s D_m] \left\{ g_n \left(\frac{\epsilon}{T_c} \right) + \frac{n_e(t, \epsilon/T_c)}{2P_s D_m} \right\} \quad (2.4-32)$$

To analyze the statistical nature of the equivalent additive noise, its autocorrelation function $R_e(\tau, \epsilon/T_c)$ must be determined

$$R_e(\tau, \frac{\epsilon}{T_c}) = E \left\{ n_e(t, \epsilon/T_c) n_e(t+\tau, \frac{\epsilon}{T_c}) \right\}$$

Following conventional methods, $R_e(\tau, \epsilon/T_c)$ can be shown [12] to reduce to

$$R_e(\tau, \epsilon/T_c) = 8 R_{Nf}^2(\tau) + 4 P_s R_{df}(\tau) R_{Nfb} \left(\frac{\epsilon}{T_c} \right) \quad (2.4-33)$$

where

$$R_{Nf}(\tau) = \frac{N_0}{2} \int_{-\infty}^{\infty} |H_1(j2\pi f)|^2 \exp(j2\pi f \tau) df$$

$$R_{df}(\tau) = \frac{N_0}{2} \int_{-\infty}^{\infty} S_d(f) |H_1(j2\pi f)|^2 \exp(j2\pi f \tau) df$$

$$b \left(\frac{\epsilon}{T_c} \right) = R_{PN/E}^2 \left(\frac{\epsilon}{T_c} \right) + R_{PN/L}^2 \left(\frac{\epsilon}{T_c} \right)$$

$$= \begin{cases} 0 & , \quad \epsilon/T_C \leq -3/2 \\ (3/2 + \epsilon/T_C)^2 & , \quad -3/2 \leq \epsilon/T_C \leq -1/2 \\ 1/2 + 2(\epsilon/T_C)^2 & , \quad |\epsilon/T_C| \leq 1/2 \\ (3/2 - \epsilon/2)^2 & , \quad 1/2 \leq \epsilon/T_C \leq 3/2 \\ 0 & , \quad \epsilon/T_C \geq 3/2 \end{cases} \quad (2.4-34)$$

The loop filter is usually chosen to be much more narrow than the equivalent noise bandwidth of $H(s)$. The equivalent single-sided noise spectral density can thus be approximated as

$$\begin{aligned} n_e(\epsilon/T_C) &= 2 \int_{-\infty}^{\infty} R_e(\tau, \epsilon/T_C) d\tau \\ &= 4 N_0^2 \int_{-\infty}^{\infty} |H_1(j2\pi f)|^4 df + \\ &\quad 4P_S N_0 b(\epsilon/T_C) \int_{-\infty}^{\infty} S_d(f) |H_1(j2\pi f)|^4 df \\ &= 2 P_S N_0 D_m [2k_D b(\epsilon/T_C) + 2k_L / \rho_i D_m] \end{aligned} \quad (2.4-35)$$

where, following Simon's definitions [13],

$$k_L = \frac{\int_{-\infty}^{\infty} |H_1(j2\pi f)|^4 df}{\int_{-\infty}^{\infty} |H_1(j2\pi f)|^2 df} \quad (2.4-36)$$

$$k_D = \frac{\int_{-\infty}^{\infty} S_d(f) |H_1(j\pi f)|^4 df}{\int_{-\infty}^{\infty} S_d(f) |H_1(j\pi f)|^2 df} \quad (2.4-38)$$

The k_L term is a constant which is dependent on the filter type whereas k_D is a constant dependent on both the baseband data power spectrum and the filter characteristic. The single-sided noise bandwidth of the equivalent lowpass filter $H_1(j\omega)$ is given as

$$B_i = \int_{-\infty}^{\infty} |H_1(j\pi f)|^2 df$$

The signal to noise ratio following the bandpass filter is defined as

$$\rho_i = P_s / (N_0 B_i) \quad (2.4-38)$$

For large ρ_i , the DLL operates in the linear region where

$$g_n(\epsilon/T_c) = \epsilon/T_c$$

Tracking Performance

If the assumption that $d\epsilon/dt=0$ is made, then from equation (2.4-32) the mean-squared jitter performance can be derived simply from the bracketed terms as

$$\left(\frac{\sigma_\epsilon}{T_c}\right)^2 = \frac{\epsilon^2}{T_c^2} = \frac{\eta_e(t, \epsilon/T_c) B_L}{(P_s D_m)^2} \quad (2.4-39)$$

Making use of earlier definitions, equation (2.4-39) is transformed using (2.4-34) and (2.4-35) to

$$\begin{aligned} \left(\frac{\sigma_E}{T_C}\right)^2 &= \frac{B_L}{(2P_S D_m)^2} (2P_S N_O D_m) \left[2k_D \left(1/2 + 2\left(\frac{\sigma_E}{T_C}\right)^2\right) + 2 \frac{k_L}{\rho_i D_m} \right] \\ &= \frac{B_L N_O}{2P_S D_m} \left[k_D (1 + 4(\sigma_E/T_C)^2) + 2 \frac{k_L}{\rho_i D_m} \right] \end{aligned}$$

Rearranging and solving for σ_E/T_C ,

$$\left(\frac{\sigma_E}{T_C}\right)^2 = \frac{1}{2\rho_L} \left[\frac{k_D + \frac{2k_L}{\rho_i D_m}}{D_m + 2k_D/\rho_L} \right] \quad (2.4-40)$$

where $\rho_L = P_S/(N_O B_L)$ is defined as the loop signal to noise ratio, a misleading term at best.

With the assumption that $\rho_L \gg 1$, it follows that a first approximation of the tracking jitter in the linear analysis scenario is given by

$$\begin{aligned} \left(\frac{\sigma_E}{T_C}\right)^2 &= \frac{1}{2\rho_L} \left[\frac{k_D + \frac{2k_L}{\rho_i D_m}}{D_m} \right] \\ &= \frac{1}{2\rho_L S_L} \end{aligned} \quad (2.4.41)$$

where S_L is referred to as the squaring loss of the DLL. The squaring loss can be related to the signal to noise ratio in the information bandwidth ρ_d ($\rho_d = P_S T_S / N_O$) with a symbol period of T_S and rate of R_S as

$$S_L = \frac{D_m}{k_D + k_L} \frac{2B_i/R_S}{\rho_D D_m} \quad (2.4-42)$$

Some general comments on S_L can be made. D_m can be made to approach unity by widening the bandpass filter (as might be required if the Doppler uncertainty is considerably larger than the data modulation spectrum) and shaping the symbols to obtain a relatively narrowband data modulation spectrum. As the number of poles of the filter increases, k_L also approaches unity. The constant k_D will approach unity when both the above characteristics are observed. Simon [13] shows an example of the squaring loss for a two-pole Butterworth filter for various combinations of B_L/R_S and ρ_d . Some general conclusions can be drawn. The filter bandwidth B_L that minimizes the squaring loss lies between two and three times the symbol rate for NRZ data with the deviation from this range becoming less significant for values of ρ_d approaching 10 dB. Holmes [2] states that at low SNR's the optimum 3 dB bandwidth is $1.4 R_S$ for NRZ data and $2.8 R_S$ for Manchester symbols.

The nonlinear analysis for the DLL tracking performance (resulting in the steady state pdf for a normalized timing error) can be accomplished using Fokker-Planck techniques. In [12], Simon shows that for a first order loop the linear and nonlinear analysis are essentially equivalent when the tracking jitter is less than 0.1 chips, particularly when the symbol rate to loop bandwidth approaches 1000 and above.

The DLL can also be configured as in Figure 2.14 and explained as follows. The discriminator characteristic for the DLL has been defined by $x_E^2 - x_L^2$; however, it can also be written as

$$\begin{aligned} x_E^2 - x_L^2 &= (x_E + x_L)(x_E - x_L) \\ &= x_\Sigma x_\Delta \end{aligned}$$

where $x_\Sigma = x_E + x_L$ is the sum of the early and late crosscorrelations and $x_\Delta = x_E - x_L$ is the difference between the crosscorrelation terms. The tracking bias caused from the filters and amplifiers in the traditional DLL of Figure 2.13 is often severe enough to warrant the latter configuration. Since the early and late crosscorrelations are summed and differenced before filtering and amplification there is no bias towards the early or late signal. The penalty for such an approach, however, is that the phase delays may become significant.

Figure 2.14 : Reconfigured DLL.

Loop Acquisition

The performance of the DLL in the presence of transients is now considered. The time to acquire PN code lock (assuming the propagation delay error ϵ is within the loop's pull-in range) is derived using phase-plane techniques. The analysis parallels that of the more familiar phase-locked loop acquisition phenomena. The maximum PN code clock offset allowable without losing lock is also deduced. This is especially important in the design of the DLL loop filter for the serial synchronization method outlined in section 2.3.2 (recall that the transmitter and receiver PN code clocks are deliberately offset).

To perform a phase-plane analysis of the DLL in the noiseless scenario, equation (2.4-32) is normalized and rewritten as [12]

$$p(y-x) = GF_0(p_0)g_n(x) \quad (2.4-43)$$

where

$$p_0 = \frac{1}{w_n} \frac{d}{dt} = \frac{p}{w_n}$$

$$x = \epsilon/T_c$$

$$y = \tau_d/T_c$$

$$G = 2kP_s$$

$$F_0(p_0) = F_0(p/w_n) = F(p)$$

The natural angular frequency of the loop is w_n .

For a second order loop with phase lead correction, as shown in Figure 2.15, the filter transfer function is given by

$$F(p) = \frac{1 + \tau_1 p}{1 + \tau_2 p} \quad (2.4-44)$$

where $\tau_1 = (R_1 + R_2)C$ and $\tau_2 = R_2 C$.

Figure 2.15 : Second order loop with lowpass filter and phase lead correction.

The damping constant ξ and natural angular frequency for a loop of this type are given by

$$\xi = 1/2 \omega_n (\tau_2 + 1/G) \quad (2.4-45)$$

$$\omega_n = \sqrt{2/\tau_2} \quad (2.4-46)$$

The corresponding closed loop transfer function $H(p)$ becomes

$$H(p) = \frac{GF(p)}{p + GF(p)} \quad (2.4-47)$$

$$= \frac{1 + p(2\xi/\omega_n - 1/G)}{1 + p(2\xi/\omega_n) + (p/\omega_n)^2} \quad (2.4-48)$$

Assuming $G \gg \omega_n / (2\xi)$,

$$H(p) = \frac{1 + p2\xi/\omega_n}{1 + p(2\xi/\omega_n) + (p/\omega_n)^2}$$

With the definitions

$$\dot{y} = \frac{1}{w_n} \frac{dy}{dt}$$

$$\dot{x} = \frac{1}{w_n} \frac{dx}{dt}$$

$$G_n = G/w_n$$

equation (2.4-43) can be written as

$$\frac{1}{G_n} \ddot{y} - \frac{1}{G_n} \ddot{x} + \ddot{y} - \ddot{x} = g_n(x) + 2\xi \frac{d}{dx} g_n(x) (\dot{x}) \quad (2.4-49)$$

Noting that

$$\frac{\ddot{x}}{\dot{x}} = \frac{d\dot{x}}{dx}$$

equation (2.4-49) can thus be reduced to a standard phase-plane format

$$\frac{d\dot{x}}{dx} = \frac{-g_n(x) - 2\xi \frac{d}{dx} g_n(x) \dot{x} + \ddot{y} - 1}{G_n} (\dot{x} - \dot{y}) \quad (2.4-50)$$

The acquisition trajectories for a one-chip noncoherent DLL are shown in Figure 2.16 from Holmes [2]. Increasing the gain marginally widens the pull-in range. If the expected initial code clock offset is expected to be large (as for serial synchronization), a gradual reduction of the loop filter bandwidth will likely be required. For example, a one-chip DLL with $G_n = \infty$ only has a pull-in range of

$$\frac{dx}{dt} = w_n$$

In other words, the code loop frequency error must be less than w_n for successful code acquisition. Another useful graph is one that presents an estimation of the time to acquire lock, shown in Figure 2.17, also reproduced from Holmes [2].

Figure 2.16 : Acquisition trajectories for a one-chip DLL with a second order loop (from Holmes [2]).

Figure 2.17 : Time to acquire lock for the one-chip DLL with a second order loop (from Holmes [2]).

DLL Mean Time To Lose Lock

A desirable quantity to know is the DLL's mean time to lose lock. Holmes and Biederman [14] have analyzed the performance of the one-chip DLL in this respect. The DLL is said to have lost lock once the code timing error exceeds $3/2$ chips since crosscorrelation with both the early and late signals is zero. In [14], the mean time to lose lock for a second order loop is derived from the tracking error of the DLL for a first order loop. The results are shown in Figure 2.18. As a general rule a tracking jitter of less than 0.2 radians will result in minimal code tracking drop-out while jitter values exceeding 0.3 radians will result in constant loss of lock and a corresponding poor despread SNR.

Figure 2.18 : DLL mean time to lose lock for a second order loop
(from Holmes and Biederman [14]).

2.4.2 Code Tracking Loops for High SNR's

When the SNR of the despread signal is relatively high, a number of other code tracking loops may be preferred. A brief description of three less hardware-intensive code tracking loops that give adequate performance under the high SNR assumption will be presented.

Time Shared Noncoherent Code Tracking Loop

One complete processing channel can be avoided with the time-shared code tracking loop. This loop splits the use of the remaining channel between the early and late signals, as shown in Figure 2.19. An equal amount of time ($T_D/2$) is allotted for correlation with the early and late signal. The dither frequency $1/T_D$ must be appropriately chosen. The dither rate should be much less than the symbol rate. But $1/T_D$ must also be much greater than the loop filter bandwidth, so that the discrete lines from the dither will not deteriorate the loop performance. Considering the time sharing of the correlator, it is not surprising to find that the timing error variance is 3 dB higher than that of the DLL [2].

Figure 2.19 : Time shared noncoherent code tracking loop.

Tau-Dither Code Tracking Loop

The tau-dither loop is similar to the time sharing loop with the exception that only one correlator is used for both the despreading and tracking, as shown in Figure 2.20. A dither generator controls which signal (early or late) is to be crosscorrelated with the received signal in a manner similar to the time shared tracking loop with the accompanying restrictions on T_D . The generation of the error signal from the tau-dither loop (TDL) is illustrated in Figure 2.21. When the codes are misaligned by less than a chip the correlator output signal envelope has an amplitude modulation at the dither rate with a magnitude that indicates the degree of timing error. The signal envelope is multiplied by the dither signal to obtain the proper polarity for the error signal and then passed through the loop filter to control the voltage controlled clock (VCC). When the code is exactly aligned the filtered error signal voltage is at zero.

The discriminator curve for the TDL is quite different from the DLL, as would be expected considering the reduction in hardware. As the dither offset D is increased, the despread signal power is seen to decrease by a factor of $(1-D)^2$ [2]. A dither of $D=0.1T_C$ is often used, giving a 0.92 dB decrease in signal power. The TDL discriminator characteristic for dither offsets of $0.1T_C$, $0.2T_C$ and $0.5T_C$ are shown in Figure 2.22 together with the signal power loss.

Figure 2.20 : A tau-dither code tracking loop.

Figure 2.21 : Generation of the TDL error signal.

Figure 2.22 : Discriminator characteristic for the TDL.

Modified Code Tracking Loop

Yet another form of tracking loop with hardware simplicity approaching that of the TDL and jitter performance approaching that of the DLL is the modified code tracking loop (MCTL) [15,16]. The MCTL is shown in Figure 2.23. When compared to the reconfigured DLL of Figure 2.14, it is observed that the error signal for the DLL is the product of the sum and difference signals while the error signal for the MCTL is just the difference signal times the punctual signal. The elimination of an entire IF processing channel points to a lower timing error for the MCTL over the DLL, particularly for the lower SNR scenario. The price paid, however, is a decreased tracking error sensitivity for timing errors exceeding 0.2 chips (for a one-chip MCTL) and a decrease in pull-in capacity by 33%. Since the discriminator curve of the MCTL,

shown in Figure 2.23, yields a lower error voltage than the DLL the probability of loss of lock increases [16]. When no filtering effects are considered equation (2.4-41), the DLL timing error, can be written as [16]

$$\frac{\sigma_E}{T_C} = \sqrt{\left(\frac{1}{2(S/N)} + \frac{1}{(S/N)^2} \right) \left(\frac{B_L}{B_{IF}} \right)} \quad (2.4-51)$$

Figure 2.23 : (a) MCTL and (b) discriminator curves for MCTL and DLL.

where S/N is the SNR following the narrowband filter (B_{IF}). The MCTL timing performance is derived as [16]

$$\frac{\sigma_E}{T_C} = \sqrt{\left(\frac{1}{2(S/N)} + \frac{1}{2(S/N)^2} \right) \left(\frac{B_L}{B_{IF}} \right)} \quad (2.4-52)$$

Hence, the MCTL is observed to rival the jitter performance of the DLL. The degree of complexity of the MCTL is nearly equivalent to the TDL but does not share its tracking performance degradation.

2.5 CARRIER SYNCHRONIZATION

Carrier synchronization is seldom a trivial exercise. One must consider the synchronization range, acquisition range, synchronization jitter, transient effects, the time to acquire lock and the mean time to lose lock. Coherent demodulation does, however, become a necessity to achieve a reliable system performance in a low SNR environment. The SNR at the output of a filter whose bandwidth is much greater than the information bandwidth (due to doppler uncertainty or oscillator instability) can be increased by lowpass filtering the coherently demodulated signal.

2.5.1 Phase-Locked Loop Operation

Consider the phase-locked loop (PLL) shown in Figure 2.24.

Figure 2.24 : Basic PLL architecture.

The input and VCO output signal are denoted by y_i and y_o , respectively, and are given as

$$\begin{aligned} y_i(t) &= A \cos(\omega_i t + \theta_i) \\ y_o(t) &= B \cos(\omega_o t + \phi_o) \end{aligned} \quad (2.5-1)$$

The VCO central angular frequency is ω_o . For a sinusoidal phase detector, the output $u_1(t)$ can be expressed as [17]

$$u_1(t) = k_1 \cos[(\omega_i - \omega_o)t + \theta_i - \phi_o]$$

where k_1 is the phase detector gain. The two signals are initially taken as being out of synchronization. Now, assuming the initial frequency offset is within the PLL's acquisition range, the two signals will become synchronous. The VCO output signal can then be written as

$$y_o(t) = B \cos(\omega_i t + \psi_o) \quad (2.5-2)$$

The quantity ϕ_o has hence been denoted as a linear function of time,

$$\phi_o = (\omega_i - \omega_o)t + \psi_o \quad (2.5-3)$$

Since $y_i(t)$ and $y_o(t)$ are now considered synchronous, the phase detector output signal will consist of a dc term and a term at frequency $2\omega_i$. Considering only the lowpass signal (with $k_2=1$),

$$u_2 = u_1 = k_1 \cos(\theta_i - \psi_o) \quad (2.5-4)$$

The instantaneous frequency output of the VCO is assumed to be a linear function of the control voltage u_2 ,

$$\omega_{\text{inst}} = \frac{d}{dt} (\omega_o t + \phi_o) = \omega_o + k_3 u_2 \quad (2.5-5)$$

where k_3 is the VCO sensitivity. From the previous equality,

$$\frac{d}{dt} \phi_o = k_3 u_2 \quad (2.5-6)$$

Making substitutions (equations (2.5-3) and (2.5-3)),

$$w_i - w_o = k_1 k_3 \cos(\theta_i - \psi_o) \quad (2.5-7)$$

By rearranging, the phase of the VCO signal can be written as,

$$\psi_o = \theta_i - \cos^{-1} \left[\frac{w_i - w_o}{k_1 k_3} \right] \quad (2.5-8)$$

When the input frequency is close to the central frequency of the VCO and the loop gain $k=k_1 k_3$ is relatively large, the two signals are seen to be in phase quadrature at the phase detector when synchronous. Strictly speaking, the two signals are only in phase quadrature when the input signal frequency is equal to the VCO central frequency. But, if the loop gain is large the deviation from phase quadrature is minimal and the two signals are thus considered to be 90° out of phase.

A more conventional form for the input signal and the VCO output signal is given as [17]

$$\begin{aligned} y_i(t) &= A \sin [wt + \phi_i(t)] \\ y_o(t) &= B \cos [wt + \phi_o(t)] \end{aligned} \quad (2.5-9)$$

From equation (2.5-9) the two signals are observed to be in phase quadrature when $\phi_i(t) = \phi_o(t)$. Any difference in frequency can be accounted for in $\phi_i(t) - \phi_o(t)$. For a sinusoidal phase detector, equation (2.5-6) may be described by

$$\frac{d}{dt} \phi_o = k [\sin \phi_e(t)] * f(t) \quad (2.5-10)$$

where $f(t)$ is the impulse response of the loop filter and the instantaneous phase error is given by $\phi_e(t) = \phi_i(t) - \phi_o(t)$. Equation (2.5-10) describes the general time-domain response of the PLL with a sinusoidal phase detector. When the phase error is small, equation (2.5-10) can be closely approximated by

$$\frac{d\phi_o}{dt} = k \phi_e(t) * f(t) \quad (2.5-11)$$

Denoting the Fourier transforms of $\phi_o(t)$, $\phi_i(t)$, $\phi_e(t)$ and $f(t)$ by $\phi_o(p)$, $\phi_i(p)$, $\phi_e(p)$ and $F(p)$, respectively,

$$\begin{aligned} p\phi_o(p) &= k [\phi_e(p)] F(p) \\ &= k [\phi_o(p) - \phi_i(p)] F(p) \end{aligned} \quad (2.5-12)$$

The general linearized transfer function of the PLL is consequently described by

$$H(p) = \frac{\phi_o(p)}{\phi_i(p)} = \frac{kF(p)}{p+kF(p)} \quad (2.5-13)$$

and the error function of the PLL is given by

$$\frac{\phi_e(p)}{\phi_i(p)} = \frac{p}{p+kF(p)} = 1 - H(p) \quad (2.5-14)$$

The transient response of the PLL can subsequently be deduced from equation (2.5-14). Blanchard [17] analyzes the natural acquisition of the input signal and derives for the second order loop of ~~Figure 2.15~~ an angular frequency acquisition range of

$$\Omega_{acq} = 2\sqrt{k\xi\omega_n} \quad (2.5-15)$$

and an angular frequency synchronization range of

$$\Omega_{syn} = k \pi/2 \quad (2.5-16)$$

for a sinusoidal phase detector. The loop natural angular frequency is denoted by ω_n . Using phase-plane techniques, Viterbi succeeded in calculating the approximate acquisition time, given as [17]

$$T_{acq} = \frac{\Omega_o^2}{2\xi\omega_n^3} \quad (2.5-17)$$

where Ω_o the angular frequency offset (i.e., $|w_i - w_o|$). Recall that Ω_o must satisfy the constraint given in equation (2.5-15) for natural acquisition to occur.

The VCO phase modulation variance, denoted as $\psi_o^2(t)$, can be derived using equation (2.5-13). First consider the noise corrupted input signal

$$\begin{aligned} y_i(t) &= A \sin(\omega t + \theta_i) + n(t) \\ &\approx A \sin[\omega t + \theta_i + n'(t)] \end{aligned} \quad (2.5-18)$$

where $n'(t)$ is the equivalent modulation, representing the effective influence of noise $n(t)$ on the signal. The VCO output signal is described by

$$y_o(t) = B \cos[\omega t + \psi_o(t)] \quad (2.5-19)$$

where $\psi_0(t)$ is the instantaneous phase error to be analyzed. The process $\psi_0(t)$ is the output signal of a filter having a transfer function $H(j2\pi f)$, driven by the noise process $n'(t)$ [17]. The mean value of $n'(t)$ is zero and its variance is given as

$$\overline{n'^2(t)} = \frac{n^2(t)}{A^2} = \frac{N_0 W}{A^2} = \frac{N_0 W}{2S} \quad (2.5-20)$$

where $N_0/2$ is the double-sided noise spectral density of $n(t)$, S is the signal power, and W is the IF filter bandwidth preceding the input signal. Consequently, $n'(t)$ is a stationary noise process with a uniform spectral density of N_0/A^2 (in rad^2/Hz) in the W Hz bandwidth.

The VCO rms phase modulation $\psi_0(t)$ is thus also a stationary gaussian noise process with a spectral density (in rad^2/Hz) defined by

$$S_{\psi_0}(f) = \frac{N_0}{A^2} |H(j2\pi f)|^2 \quad (2.5-21)$$

The variance of $\psi_0(t)$ (in rad) is simply the product of the transfer function and phase modulation spectral density, integrated over the IF bandwidth W . Representing the IF filter by an equivalent lowpass filter (from $-W/2$ to $W/2$), the variance is given as

$$\sigma_{\psi_0}^2 = \frac{N_0}{A^2} \int_{-W/2}^{W/2} |H(j2\pi f)|^2 df \quad (2.5-22)$$

Since $|H(j2\pi f)|^2 \approx 0$ for $|f| > W/2$, equation (2.5-22) can be simplified as

$$\sigma_{\psi_0}^2 = \frac{N_0}{A^2} \int_{-\infty}^{\infty} |H(j2\pi f)|^2 df \quad (2.5-23)$$

The integral in equation (2.5-23) represents the noise equivalent bandwidth of a loop having a transfer function given by $H(j2\pi f)$, and is conventionally denoted as [17]

$$(2B_n) = \int_{-\infty}^{\infty} |H(j2\pi f)|^2 df \quad (2.5-24)$$

It follows from symmetry that the unilateral noise band is given by

$$B_n = \int_0^{\infty} |H(j2\pi f)|^2 df \quad (2.5-25)$$

The band $(2B_n)$ should be used when the noise spectral density is double-sided and, similarly, B_n should be used for a single-sided noise spectral density. In any event, equation (2.5-23) can be reduced to

$$\frac{\sigma_{\psi_0}^2}{A^2} = \frac{N_0}{A^2} (2B_n) = \frac{N_0 B_n}{S} \quad (2.5-26)$$

The equivalent SNR in the loop filter bandwidth is often used to describe the random phase modulation variance,

$$\rho_L = \frac{S}{N_0 B_n} = \frac{1}{\frac{\sigma_{\psi_0}^2}{A^2}} \quad (2.5-27)$$

The second order loop filter presented in section 2.4.1 (lowpass filter with phase lead correction) is usually chosen because of the flexibility in choosing k , ω_n and ξ independently. The noise equivalent bandwidth for a loop filter of this type is given by [17]

$$(2B_n) = \frac{\omega_n}{4\xi} (1+4\xi^2) \quad (2.5-28)$$

or, for the unilateral noise bandwidth,

$$B_n = \frac{w_n}{8\xi} (1+4\xi^2)$$

(2.5-29)

2.5.2 Suppressed Carrier Tracking Loops

A modulation scheme with no residual carrier is described by

$$s(t) = A m(t) \sin(w_0 t + \theta_0) \quad (2.5-30)$$

where $m(t) = \pm 1$ during the symbol period, w_0 is the carrier angular frequency and θ_0 is the unknown phase. Since there is no residual carrier at w_0 , a conventional PLL will not suffice.

There have been several methods proposed over the years for suppressed carrier tracking. The squaring loop, Costas loop and decision-directed feedback loops are most commonly used. The decision-directed feedback loop can be described as follows: the signal modulation is first estimated, and then using this estimate, an attempt is made to eliminate the modulation and leave an unmodulated sinusoid for PLL tracking. The squaring loop, shown in Figure 2.25(a), has the same stochastic differential equation as the Costas loop [18], shown in Figure 2.25(b).

Consider the operation of the squaring loop. The received signal is bandpass filtered, squared to remove the modulation and then bandpass filtered at the double-frequency for PLL tracking. The VCO output signal is given a 90° phase shift and then frequency halved to provide a coherent reference for demodulation.

Figure 2.25 : Suppressed carrier tracking using (a) a squaring loop and (b) a Costas loop.

The performance of the squaring loop has been analyzed by numerous authors, including [2,18]. The essential difference between the squaring loop phase jitter and the PLL phase jitter given in equation (2.5-27) is the addition of a loss term due to the signal squaring and a term to represent the phase magnification,

$$\overline{\sigma_{2\psi}^2} = 4 \left(\frac{N_0 B_n}{S} \right) L_n \quad (2.5-31)$$

where $\overline{\sigma_{2\psi}^2}$ is the VCO phase jitter. L_n is the loss due to noise intermodulation, also referred to by Lindsey and Simon as the circuit squaring loss. The factor of four represents the doubling of the input signal phase. While equation (2.5-31) represents the VCO output phase jitter, the phase jitter of the carrier reference for demodulation will be reduced by of four (i.e., a factor of two for the rms jitter due to frequency halving)

$$\overline{\sigma_{\psi}^2} = 1/4 \overline{\sigma_{2\psi}^2} = \frac{N_0 B_n}{S} L_n \quad (2.5-32)$$

Hence, the squaring loop suffers a degradation of L_n over the linear performance of a PLL. The intermodulation loss is defined for a BPSK signal as

$$L_n = 1 + 1/(2 \rho_i) \quad (2.5-33)$$

where $\rho_i = S/(N_0 W)$ is the effective SNR following the W Hz BPF.

Combining equations (2.5-32) and 2.5-33), the phase jitter of the reference carrier is deduced as

$$\overline{\sigma_{\psi}^2} = \frac{N_0 B_n}{S} \left(1 + \frac{1}{\frac{2S}{N_{0W}}} \right) \quad (2.5-34)$$

The synchronization range, acquisition range and acquisition time for the squaring loop are described by the corresponding PLL values (with acknowledgement given, of course, to the doubled frequency signal).

A SPREAD SPECTRUM TIMING BEACON

3.1 INTRODUCTION

With the theoretical base having been provided in the previous chapter, the thesis can now be focussed on a particular application of spread spectrum. The overall system requirements and the environmental scenario are first presented in the next section. Subsequent sections discuss the means by which to obtain these specifications. PN code generation, code acquisition and tracking, carrier synchronization and bit recovery are all analyzed and designed to meet the system requirements.

3.2 SYSTEM OBJECTIVES AND ENVIRONMENTAL SCENARIO

As earlier mentioned, the transmit baseband signal consists of a 300 bps differentially-encoded PSK (DE-PSK) data signal digitally mixed with a 1 MHz direct sequence PN code. The combined time for both PN code acquisition and carrier synchronization should not exceed one minute.

The receiver must provide a timing mark (i.e. a unique PN code segment or vector) that can synchronize the receiver clock to that of the transmitter within an accuracy of 1 μ sec. The propagation delay between the transmitter and receiver is assumed to be approximately known. A

high system reliability (a BER performance exceeding 10^{-4}) is desired under the worst case scenario. The system must also be able to tolerate a maximum doppler uncertainty of ± 1 kHz.

The transmit signal has a power of 1 Watt. The proposed transmit antenna has a circular beamwidth of 10° and an efficiency of 55%. The downlink frequency is 20 GHz and the link distance is 41,756 km. The receive antenna has a circular aperture of diameter 0.46 meters and an assumed efficiency of 55%. The effective noise temperature of the receiver (including galactic noise, thermal noise, LNA noise figure, etc...) is given to be no higher than 1000° K. In addition, a 6 dB fade margin should be provided.

3.3 DOWNLINK SNR CALCULATION

The downlink receive power (P_r) can be calculated from the system parameters given in section 3.2,

$$P_r = \frac{P_t A_t A_r}{R^2 \lambda^2} \eta_t \eta_r \text{ [Watts]} \quad (3.3-1)$$

where P_t is the transmit power, A_t and A_r are the transmit and receive effective antenna areas with corresponding antenna efficiencies of η_t and η_r , λ is the carrier wavelength and R is the link distance. A 20 GHz signal has a free-space wavelength of 15 mm. The

receive antenna has a circular aperture with a diameter of 0.46 m, or an aperture area of 0.1662 m^2 . The directivity of the transmit antenna D_o can be related to the beamwidth using the empirical equation [20]

$$D_o = \frac{41253}{\phi_B \theta_B} \quad (3.3-2)$$

$$= 412.5 = 26.2 \text{ dB}$$

where ϕ_B and θ_B are the orthogonal beamwidths (i.e. E and H field directions) measured in degrees. The effective antenna aperture can be related to the directivity using the classical equation

$$A_t = \frac{\lambda^2 D_o}{4\pi} \quad (3.3-3)$$

$$= 7.39 \times 10^{-3} \text{ m}^2$$

Hence, the receive power is calculated as

$$P_r = \frac{(1) (7.39 \times 10^{-3}) (0.166)}{(4.1756 \times 10^7)^2 (0.015)^2} (0.55)^2$$

$$= -120.24 \text{ dBm}$$

The noise spectral density is calculated as

$$N_o = kT_o = -168.6 \text{ dBm/Hz}$$

where k =Boltzmann's constant= $1.38 \times 10^{-23} \text{ J/}^\circ\text{K}$ and $T_o=1000^\circ\text{K}$ is the assumed receiver noise temperature. For an ideal bandpass filter with a 2 MHz bandwidth, the noise power through the filter W_{ss} is

$$N = N_o W_{ss} = -105.6 \text{ dBm}$$

The effective input SNR of the spread signal is calculated as

$$\text{SNR}_i = P_r - N = -14.6 \text{ dB}$$

With a link fade margin of 6 dB, the expected worst case input SNR becomes

$$\text{SNR}_i / \text{worst case} = -20.6 \text{ dB}$$

Under ideal despreading and demodulation, the signal to noise ratio in the information bandwidth can be derived from the effective input SNR and the processing gain as

$$\begin{aligned} \text{SNR}_{\text{info}} &= \frac{2E_b}{N_o} = \text{SNR}_i + \text{PG} \\ &= -20.6 + 35.2 = 14.6 \text{ dB.} \end{aligned}$$

This represents the maximum SNR (under the worst case scenario) with no degradation due to squaring loss, partially coherent demodulation or implementation loss.

The system goal is to achieve a 10^{-7} BER, or an SNR in the information bandwidth of 12 dB for DE-PSK. At first glance, the allowable system degradation of only 2.6 dB appears to be a rather tight bound. Forward error correction using convolutional codes can ease this constraint considerably, providing as much as a 4 to 6 dB SNR reduction for the same BER [22].

3.4 PN CODE LENGTH AND GENERATION

As earlier alluded to, the length of the PN code helps to guide much of the system architecture. A code repeating once a day is certainly no more difficult to design and build than a code that repeats many times a second. However, the ramifications of the code length and its repetition rate influence the techniques and technologies used for acquisition and tracking loop design.

Since the desired acquisition time should not exceed one minute, a code repetition period of one minute is the upper bound for matched filtering acquisition. (At a 1 MHz clock rate, this represents a code length of 60 million bits.) Recall that the potential users are to receive timing information to synchronize their clocks to the satellite time; hence, ignorance of the time is assumed and a worst case wait of 60 seconds for acquisition is allotted. Note that the PN code timing mark will also only appear once in the code period. The analysis that follows in section 3.5.1 shows that MF techniques are not appropriate for this application. Since the chosen alternative is serial synchronization, a code repetition period of a much shorter duration should be considered.

First consider the lower bound on the PN code length. The timing mark should not appear more than once over the duration of one ASCII character. Since 10 bits are required for each ASCII character, the PN code period should not be any shorter than $1/30$ seconds (i.e. 10 bits at 300 bps). The spectral spacing for such a PN code length is 30 Hz. In

section 2.4.1 it was assumed that the loop filter bandwidth was larger than the spectral spacing yet much lower than the modulation rate. A code repetition rate of 30 Hz with a data rate of 300 Hz weakens the assumption of negligible modulation self-noise. A code that repeats once a second would satisfy the above constraints. The spectral spacing would be 1 Hz and an appropriate tracking loop filter could be realized without much difficulty. Allowing for a maximum of 50 seconds for the receiver to advance one full code period relative to the transmit code (the worst case wait), the required drift rate is 20 kHz. Unfortunately, the SNR improvement from a 20 kHz search rate will not increase the filtered cross correlation SNR to an acceptable level. A drift rate of no more than 2 kHz is shown to be required. A 2 kHz search that must take no longer than 50 seconds is realized with a code length of 100,000 chips. The repetition rate is 10 Hz and a total of 3 ASCII characters (30 bits) can be transmitted between timing marks. Thus, it is proposed that once the timing information has been loaded into a buffer, synchronization with the satellite time could be initiated with the next timing mark that follows a seldom used ASCII character reserved especially for this purpose. In other words, the receiver monitors the timing mark that appears every 0.1 seconds, but only uses it for synchronization following a specific ASCII character, as shown in Figure 3.1.

Figure 3.1.: A 1 μ sec accuracy timing mark for receiver synchronization.

A code that is an integral number of seconds long or repeats an integral number of times a second will simplify the timing synchronization exercise. (Recall that the length of a ML sequence is $2^n - 1$, hardly a convenient length.) This leads to the generation of a PN code segment. A shift register with 17 stages can be used to generate a ML sequence with a period of 131,071 bits. Hence, a code length of 100,000 bits can be realized using a shift register with 17 or more stages. All that is required is a "start vector" to load in the initial PN code and a "stop vector" that will interrupt the natural code progression and initiate loading of the start vector. The schematic for code segment generation is shown in Figure 3.2. The shift registers are usually in the output mode with Q_1 generated by an exclusive-OR of the appropriate feedback taps. When the generated PN code exactly matches the hard-wired stop vector the comparator goes high and the shift registers are held in input mode for one chip while the start vector is loaded from the latches. The latch is 3-state when not enabled.

Figure 3.2 : PN code segment generation.

Initial attempts to realize a PN code period of 60 seconds lead to the design of a 26 stage shift register. When it was finally decided to use a 100,000 bit length code due to acquisition time constraints, the existing wire-wrapped 26 stage shift register was used.

Caution must be exercised when using only a segment of a PN code. A "free running" ML sequence has only one more "1" than "0" in the code period. But an arbitrarily chosen code segment may very well have a much greater difference in ones and zeroes. This will lead to code unbalance, resulting in tones at the spread signal's $(\sin x/x)^2$ nulls. An example of the spread signal with and without code balance is shown in Figure 3.3.

(a)

(b)

Figure 3.3 : A 1 MHz direct sequence spectrum (a) with code balance and (b) without code balance. (Y-axis: 5 dB/div)

A 26 stage ML sequence can be realized using Q_1 , Q_2 , Q_6 and Q_{26} as the feedback taps [1,2]. To ensure code balance a simple Fortran program was run to search for a 100,000 bit sequence that had an equal number of ones and zeroes. The start vector (read from Q_1 on the left to Q_{26} on the right) was found to be given by

01001001100000110011101111

and the stop vector by

10101001111101011000001000

The complete PN generator circuit is shown in Figure 3.4. The baseband PN modulated data signal delivered by the transmitter is shown in Figure 3.5(a). The receiver generates an early, punctual and late signal, as shown in Figure 3.5(b).

3.5 PN CODE ACQUISITION

An attempt to use matched filtering techniques is first discussed. There are really just two technologies that can be considered for this application: the DMF and SAW TDL correlator. It will be shown that neither of these devices are suitable and that the serial search method should be used.

Figure 3.4 : PN generator circuit.

TRANSMITTER BALANCED PN GENERATOR

Figure 3.5 : (a) Transmitted PN code with data

RECEIVER BALANCED PN GENERATOR

Figure 3.5 (continued): (b) receiver generated code.

3.5.1 Matched Filter Acquisition

Digital Matched Filter

The noncoherent DMF SNR with multibit quantization was shown in section 2.3.1 to be upper bounded by that of the AMF. For $\text{SNR}_i \ll 1$ and $\text{SNR}_O \gg 1$, the matched filter output can be described by

$$\begin{aligned} \text{SNR}_O &= \frac{TW}{2} (\text{SNR}_i) \\ &= 10\log(TW) + \text{SNR}_i - 3\text{dB} \end{aligned} \quad (3.5-1)$$

The worst case input SNR was earlier calculated to be -20.6 dB. The resulting matched filter output SNR is

$$\text{SNR}_O = 10\log(TW) - 23.6\text{dB} \quad (3.5-2)$$

The SNR, P_{fa} , P_d curves of Figure 2.9 reveal that for $P_{fa}=10^{-3}$ and $P_d=0.5$ (a reasonable worst case expectation), the required SNR is 10 dB. A time-bandwidth product of 33.6 dB is the observed requirement. An effective integration of over 2290 chips for each quantization level will yield the necessary improvement. By sampling at the Nyquist rate one observes that over 4500 stages would be required for each quantization level in both channels. Furthermore, the 1 kHz doppler uncertainty can lead to an additional degradation of the filter output SNR. The result is a reduced effective integration time.

An offset of $\Delta\omega$ rad/sec for the noncoherent DMF of Figure 2.6 results in I and Q channel integrations of

$$I = \frac{1}{T_s} \int_0^{T_s} \cos \Delta\omega t \, dt = \frac{\sin \Delta\omega T_s}{\Delta\omega T_s} \quad (3.5-3)$$

$$Q = \frac{1}{T_s} \int_0^{T_s} \sin \Delta \omega t \, dt = \frac{1 - \cos \Delta \omega T_s}{\Delta \omega T_s} \quad (3.5-4)$$

where T_s is the integration time of the DMF. The sum is given by

$$I^2 + Q^2 = \frac{\sin^2 \frac{\Delta \omega T_s}{2}}{\frac{\Delta \omega T_s}{2}} \quad (3.5-5)$$

The sum is less than unity for $\Delta \omega \neq 0$. For example, a 1 kHz offset with an integration time of 2250 μ sec results in a 20 dB integration loss! This loss can be greatly reduced by summing a number of (reduced integration time) parallel matched filters whose total integration time is 2250 μ sec. But even this method has its problems. The sum of a number of noncoherently integrated "pulses" (in radar terms) results in a combining loss as illustrated by Skolnik [9]. Clearly, the DMF is not the preferred route for this application.

SAW TDL Correlator

Current state of the art in SAW TDL correlators offer an integration time of 60 μ sec. Since the ss timing beacon requires a bandwidth of only 2 MHz, the time-bandwidth product is only of the order of 21 dB, only enough to increase the output SNR to 0 dB. The manufacturing cost for SAW devices is also very expensive, providing even less incentive to use these devices when only small quantities are required.

Unless a preamble requiring much less bandwidth is used for acquisition, matched filtering is not applicable with the combined low input SNR and relatively narrow bandwidth.

3.5.2 Serial Synchronization

The degree of SNR improvement achieved with serial synchronization is dependent on the search rate (assumed equal to the narrowband filter bandwidth). By reducing the filter bandwidth following the crosscorrelation, a SNR improvement by as much as the system processing gain is theoretically possible. But recall that this is only achieved for a perfectly tracked signal.

The doppler uncertainty constrains the filter bandwidth to be no less than 2 kHz. A 30 dB decrease in noise power is observed for the 2 kHz filter. The SNR of the perfectly tracked signal is consequently increased by 30 dB. The worst case despread SNR following the 2 kHz BPF is hence upper bounded at 9.4 dB. However, the serial synchronization correlation pulse will have an SNR of 7.6 dB. The 1.75 dB decrease in SNR can be reduced by searching at a slower rate. Unfortunately, this also causes an increase in the code acquisition time. From Figure 2.9, an SNR of 7.6 dB combined with a false alarm probability of 10^{-3} yields a detection probability of 0.37. When the fade is 3 dB less severe, the detection probability increases to 0.85 for the same false alarm probability. If a considerably higher detection probability and lower false alarm rate are required for the worst case, the use of a preamble should be considered. A large decrease in the acquisition time could

only be achieved if the preamble started with a burst of carrier alone. This would minimize the otherwise potentially overwhelming integration losses associated with the doppler uncertainty.

Assuming that the worst case is rarely encountered, a search rate of 2 kHz is proposed. The acquisition loop need not be as robust as the tracking loop since the code only has to be detected and acquired once (i.e. the "duty cycle" of the acquisition process is very low).

When the receiver is in acquisition mode, a constant voltage is applied to the voltage controlled clock (VCC) to maintain the 2 kHz drift rate. The hardware realization of the acquisition IF channel is illustrated in Figure 3.6. (Section 3.6.2 discusses the need for filtering the receiver generated code.)

When the codes are not within a chip the envelope detected signal, $\tilde{x}_p^2$, will only exceed the threshold voltage during false alarms. Once the threshold has been exceeded, whether it is due to a false alarm or code alignment, the control voltage must then be driven by the tracking loop.

The time between false alarms can be calculated from equation (2.3-11) as

$$\begin{aligned} T_{fa} &= (P_{fa} B_{IF})^{-1} \\ &= (10^{-3} \times 2 \times 10^3)^{-1} = 0.5 \text{ seconds} \end{aligned}$$

Figure 3.6 : Serial search acquisition hardware implementation.

This will lead to an inordinate number of false alarms to chase up. A decrease in the false alarm rate to 10^{-4} will lead to an average time between false alarms of 5 seconds. This means that, on average, there will be 5 false alarms to chase up during acquisition. Note that the worst case scenario leads to a detection probability of only 0.18, a 3 dB fade has $p_d=0.67$ and no fade results in $p_d=0.993$.

The envelope detector used for acquisition was an HP8471A. Measurements of the input power and the output voltage were made, with the results shown in Figure 3.7. From these results, it can be observed that the optimum power of the despread signal at the detector should be in the range -24 dBm to -10 dBm (i.e., the square-law region).

The despread carrier power (i.e. no data) at the envelope detector was measured to be -19 dBm. The SNR at the detector has been estimated for the worst case at 9.4 dB. Consequently, the average noise power at the detector is -28.4 dBm. The rms noise voltage at the detector output is approximately 1.2 mV. The threshold voltage for an rms noise voltage of 1.2 mV and a false alarm probability of 10^{-4} is calculated using equation (2.3-10) to be 5.2 mV. An amplifier with a voltage gain of 100 precedes the voltage comparator. Hence, the threshold voltage for the synchronization detection should be set at 520 mV.

Figure 3.7 : Envelope detector characteristics for acquisition detection.

3.6 PN CODE TRACKING

The choice of tracking device should be based on the expected worst case SNR following the narrowband filter. A high SNR allows one the flexibility of using a less hardware intensive tracking circuit. The worst case effective SNR following the 2 kHz filter was earlier calculated to be 9.4 dB. Since the squaring loss is likely to be of the order of 2 dB, a degradation in signal power by this amount must be accepted. The hardware realization will introduce further losses such as correlation and implementation losses. Such losses may result from a tracking loop bias due to differences or imperfections of loop components (e.g. mixers, amplifiers, filters, envelope detectors). Assuming a 2 dB implementation loss, the expected worst case SNR following the narrowband filter is approximately 5 dB. The time-sharing loop and TDL should not be used with this low SNR. The choice of tracking loop is thus reduced to either the DLL or the MCTL. The MCTL performs very well at low SNR's; in fact, it can outperform the DLL in this range.

Consider the acquisition process for the proposed synchronization scheme. Threshold (for the noiseless case) is not exceeded until the two codes are aligned. But the group delay of the 2 kHz filter will be approximately 0.5 ms. Since the receiver PN code continues to drift past the incoming code until the presence of the code has been detected, the two codes will be almost a full chip out of synchronization by the time threshold is exceeded. Consequently, the tracking device must be fairly dynamic with a pull-in range of at least one chip. The MCTL has a relatively poor pull-in range and a reduced error signal when compared to

the DLL. The DLL was therefore chosen as the tracking device. Furthermore, when the pull-in range, search rate and discriminator characteristic are simultaneously considered in terms of the early and late signal separation, the one-chip DLL appears to be the best compromise. A close look at the relative timing offset upon threshold detection is essential to determine if the pull-in range of the one-chip DLL is sufficient. A 1.5- or 2-chip DLL will provide a wider pull-in range at the expense of tracking sensitivity.

3.6.1 Design of a One-Chip Delay-Locked Loop

The one-chip DLL timing performance, neglecting any filtering effects, is given by equation (2.4-51) as

$$\frac{\sigma_E}{T_C} = \sqrt{\left[\frac{1}{2(S/N)} + \frac{1}{(S/N)^2} \right] \frac{B_L}{B_{IF}}}$$

With a worst case S/N of 5 dB=3.16 and a narrowband 2 kHz filter, the timing performance can be simplified to

$$\begin{aligned} \frac{\sigma_E}{T_C} &= \sqrt{\left[\frac{1}{2(3.16)} + \frac{1}{(3.16)^2} \right] \frac{B_L}{2000}} \\ &= 0.0114 \sqrt{B_L} \end{aligned} \quad (3.6-1)$$

Recall the bounds on the loop filter. It must be well below the data rate (300 Hz) to ignore the modulation self-noise term and above the code repetition rate (10 Hz). It is also desirable to have an rms tracking jitter of less than 0.1 chips (see Figure 2.18). A tracking loop rms jitter of 0.1 chips is achieved with a 77 Hz loop filter. A 40 Hz loop filter will satisfy the constraints and still provide a reasonable margin of flexibility. Doubling or halving of the 40 Hz loop filter will still

result in adequate performance; consequently, nonideal components in the loop filter and discriminator that may alter the loop filter bandwidth can be tolerated. While a tracking loop filter of 40 Hz will satisfy the tracking performance, it will certainly fall far short of satisfying the acquisition requirement.

The acquisition process for the one-chip DLL (with $G=\infty$) was earlier described by

$$\frac{1}{w_n} \frac{d}{dt} \left(\frac{\epsilon}{T_c} \right) \leq 1.0$$

The term $d(\epsilon/T_c)/dt$ is the code loop frequency error (in chips/sec).

In acquisition mode the transmitter and receiver clocks are deliberately offset by 2 kHz. The loop natural angular frequency must then be greater than 2000 rad/sec for the tracking loop to "lock" onto the incoming PN code. If the second order loop filter shown in Figure 2.15 is used with $\xi=0.707$, the loop filter bandwidth can be related to the loop natural angular frequency by

$$B_L = 0.5303 w_n \quad (3.6-2)$$

The loop filter must then be greater than 1060 Hz for successful acquisition. To compensate for possible hardware degradations $w_n=3000$ rad/sec (with $B_L=1591$ Hz) was chosen. An interesting problem is observed: the tracking loop must be dynamic enough to respond to the 2 kHz frequency offset error signal using a 1600 Hz loop filter, yet it must also provide a 40 Hz loop filter to obtain the desired tracking performance. The proposed solution is to simultaneously generate 4 error signals, using different loop filters (the 1600 Hz and 40 Hz filters being the extremes). Digital circuitry would control which error signal is applied to the VCC. The simplified error signal generation is illustrated in Figure 3.8.

The tracking loop strategy is now summarized. A wide tracking loop filter is required for code acquisition and a narrow loop filter is required for adequate tracking performance. The loop filter will slowly be reduced from 1600 Hz to 40 Hz by stepping through four loop filters, each filter more narrow than the previous one. In the acquisition mode, the VCC is controlled by the drift voltage (a dc voltage that generates the 2 kHz offset). Once the acquisition threshold has been exceeded the control voltage of the VCC is supplied from the tracking loop. The output of the first loop filter controls the VCC for a predetermined length of time (the code acquisition time). The second, third and fourth loop filter are then consecutively stepped through, each supplying an error signal for a set length of time. As long as the threshold is still

Figure 3.8 : Error signal generation using 4 loop filters.

2 OF/DE 2

MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS
STANDARD REFERENCE MATERIAL 1010a
(ANSI and ISO TEST CHART No. 2)

exceeded, the narrow loop filter continues to supply the error signal to the VCC. In fact, a dropout of approximately 1 second is tolerated before the circuit is returned to acquisition mode. Consequently, if $\tilde{x}_p^2$ momentarily drops below threshold due to gaussian noise or some transient, the tracking loop should be able to maintain code synchronization.

The code acquisition time for the one-chip DLL can be estimated from Figure 2.17. Code acquisition is considered complete for $t=6/w_n$. The wide loop filter has an acquisition time of 2 ms while the narrow loop filter has an acquisition time of 80 ms. This provides the minimum duration for the switch controls of Figure 3.8.

The loop gain must be higher than the expected code frequency offset. Choosing $k=5$ kHz will allow for a reasonable degradation in the loop gain from the loop components. If the natural angular frequencies of the four loops are 3000, 850, 250 and 75 rad/sec a bandwidth reduction by a factor of 3.4 is observed between adjacent filters. The loop filter components can now be computed using the following equations for the second order loop

$$\tau_1 = \frac{k}{\omega_n^2} \quad (3.6-3)$$

$$\tau_2 = \frac{\tau_1}{k} \left(2\xi\omega_n - \frac{1}{\tau_1} \right) = \frac{2\xi}{\omega_n} \quad (3.6-4)$$

$$R_1 = \frac{\tau_1 - \tau_2}{C} \quad (3.6-5)$$

$$R_2 = \frac{\tau_2}{C} \quad (3.6-6)$$

Table 3.1 gives the required component values for the four loop filters.

Loop Filter No.	ω_n (rad/sec)	B_L (Hz)	τ_1 (ms)	τ_2 (ms)	C (μ F)	R_1 (k Ω)	R_2 (k Ω)
1	3000	1591	0.56	0.27	0.01	28.5	27.0
2	850	451	6.9	1.46	0.10	54.4	14.6
3	250	133	80	5.5	0.47	159	11.6
4	75	40	890	18.7	4.7	184	3.9

Table 3.1 : Loop filter component values.

The loop gain (k) is the product of the discriminator characteristic (k_D in volts/chip), the VCC sensitivity (k_{VCC} in chips/volt/sec) and the amplifier gain (k_2). The envelope detectors used for the early and late crosscorrelated signals were the HP8473B low-barrier Schottky diode detectors. A table of the V_{out} versus P_{in} values and a corresponding graph is shown in Figure 3.9.

The despread signal power at the envelope detector input for the early and late signals was designed to be at -7.5 dBm. In other words, a steady state error of $\pm T_c/2$ will result in a signal power level of -7.5

dBm at either the early or late envelope detector with the other detector receiving an uncorrelated (lower power) signal. The usual case, however, is for small errors, a 6 dB decrease in despread signal power (i.e. -13.5 dBm) at both the early and late envelope detectors. Although it is desired to operate the envelope detectors in the square-law region, consideration must be given to the magnitude of the expected detector output voltage. Circuit noise from the envelope detectors, discriminator IC's and supporting hardware cannot be overlooked if the signal voltages are only a few millivolts. Signal powers in the range -20 dBm to -10 dBm should suffice.

Figure 3.9 : V_{out} versus P_{in} for the early and late signal envelope detectors

The discriminator characteristic of Figure 2.12 can be extrapolated (neglecting any filtering effects) from Figure 3.9. The resulting DLL discriminator characteristic is shown in Figure 3.10. Note that the small error sensitivity is given by $k_D = 0.12$ volts/chip.

The VCC sensitivity is now considered. A high stability oscillator was required to maintain a code clock offset of 2 kHz in the acquisition mode and a relatively wide frequency deviation was required to switch from acquisition mode to track mode. The simplest way to meet these requirements was to implement the circuit shown in Figure 3.11.

Figure 3.10 : DLL discriminator characteristic.

Figure 3.11 : A stable, wide deviation voltage controlled clock.

Figure 3.12 : DLL discriminator with loop filters and digital control.

The VCC sensitivity was measured to be approximately 1060 chips/volt/sec.

The loop gain k was designed to be 5 kHz. Consequently, the amplifier gain is specified as

$$k_2 = \frac{k}{k_D k_{VCC}} = 39$$

The complete DLL discriminator and loop filters with the controlling digital circuitry is shown in Figure 3.12. Note that V_{DRIFT} of U37-1 is the VCC control voltage until threshold has been exceeded. Once exceeded, each error signal of U37-2 is consecutively used as the VCC control voltage.

3.6.2 Filtered PN Signal Correlation Loss

Bandpass filtering the received spread signal will result in both a distortion and a delay of the PN sequence. Since the leading edge of a rectangular pulse is due to the higher frequency terms, it follows that the filtered chip transitions will be less abrupt and delayed. Mixing the filtered signal directly with the receiver generated PN sequences will hence result in a correlation loss. In a straightforward analysis, it can be shown that the maximum output SNR of the correlator shown in Figure 3.13 is obtained with matched signals [2]. That is, the same degree of distortion should be introduced to the receiver generated signals.

Figure 3.13 : Correlation of the filtered PN code.

As one might expect, decreasing the filter bandwidth increases the correlation loss. This results simply from the averaging effect of the BPF. Holmes [2] shows that for the commonly chosen bandwidth of $2/T_C$, the correlation loss is 0.92 dB.

At the same time, one must recognize that the filtering leads to a reduction in the effective noise spectral density. The mixer output noise, spread by the PN sequence, can be written as

$$N(t) = n(t)PN(t) \quad (3.6-7)$$

with an autocorrelation function of

$$R_N(\tau) = E[N(t)N(t+\tau)]$$

The noise spectral density can be deduced [2] from the corresponding autocorrelation function. For an ideal BPF Holmes shows that the noise spectral density can be expressed by

$$S(w_0) = \frac{N_0}{2} \int_{-BT_C/2}^{BT_C/2} \left(\frac{\sin x}{x} \right)^2 dx \quad (3.6-8)$$

where B is the BPF bandwidth. The reduction in noise spectral density for a bandwidth of $2/T_C$ is 0.51 dB, so that the effect of correlation loss and noise reduction results in only a 0.41 dB reduction in SNR.

The 0.92 dB correlation loss for $BT_C=2$ quoted above assumed an ideal BPF. A nonlinear phase response of the BPF will cause an additional correlation loss. An amplitude variation across the bandwidth will add yet a further correlation loss. Linearity in both the amplitude and phase response of a filter is difficult to realize simultaneously. Gaussian or Bessel filters yield excellent group delay response at the expense of a rounded amplitude characteristic. Butterworth filters yield a maximally flat amplitude response but have poor group delay characteristics.

The above arguments lead to bandpass filtering of the receiver generated code. In an attempt to introduce the same degree of delay and distortion Bessel filters with the same specifications (except center frequency) were used for both the incoming and receiver generated signals. The PN code synchronizing circuits (acquisition and tracking) illustrating correlation with the remodulated, filtered PN codes are shown in Figure 3.14.

Figure 3.15 shows a photograph of the acquisition and tracking loop circuitry. The receiver generated PN code and digital control for the

Figure 3.14 : Receiver PN code synchronization circuits.

Figure 3.15: Remodulated PN sequence with code
Synchronization circuits.

tracking loop filters are located on the left side of the figure. An effort was made to separate the (low level) analog signals from the digital circuitry. By individually packaging the code acquisition detection circuit and tracking loop discriminator and loop filters, interference caused by 1 MHz square-wave radiation via the wire-wrap pins was minimized.

3.7 COHERENT DEMODULATION OF THE DESPREAD SIGNAL

The despread signal will consist of a 300 bps differentially encoded PSK NRZ modulated carrier with a ± 1 kHz doppler uncertainty. DPSK demodulation can be accomplished without actually generating a carrier reference at the receiver. An analog delay equal to the bit period can be used for mixing the signal to baseband, as shown in Figure 3.16. However, doppler uncertainty can significantly degrade this demodulation method. Since the delay is designed to represent an integral number of carrier cycles, a drifting carrier frequency will result in a relative phase offset at the mixer - the demodulation no longer appears coherent.

Alternatively, one can first demodulate the suppressed carrier signal, providing DE-PSK demodulation, as shown in Figure 3.17. A squaring loop can be used to coherently demodulate the signal.

Figure 3.16 : DPSK demodulation.

Figure 3.17 : DE-PSK demodulation.

Consider the squaring loop configuration of Figure 3.18. The VCO output signal provides a coherent reference for demodulation. Because of the squaring of the input signal, the frequency uncertainty in the loop is ± 2 kHz. For natural acquisition to occur, an acquisition range of at least 2 kHz must be provided. To keep the sinusoidal phase error small (thus reducing steady state phase errors [2]), a loop gain of the order of 100 kHz is required. In fact, the loop gain for the squaring loop was chosen to be equal to 115 kHz - for a sinusoidal phase detector the maximum phase error is 1° . The acquisition range was defined by equation (2.5-15) as $\Omega_{acq} = 2\sqrt{k\xi\omega_n}$. Choosing $\xi = 0.707$, ω_n is lower bounded by

$$\omega_n \geq \left(\frac{\Omega_{acq}}{2} \right)^2 \frac{1}{k\xi} = 485 \text{ rad/sec}$$

Figure 3.18 : Squaring loop.

The tracking loop jitter was defined earlier as

$$\begin{aligned}\overline{\sigma_\psi^2} &= \frac{N_0 B_n}{S} \left(1 + \frac{1}{\frac{2S}{N_0 W}} \right) \\ &= \frac{TB_n}{E_b/N_0} \left(1 + \frac{1}{\frac{2E_b/N_0}{TW}} \right) \quad (3.7-1)\end{aligned}$$

The worst case SNR following the narrowband IF filter was estimated to be 5 dB. The SNR in the information band is 5.2 dB higher (i.e. $10\log(1000/300)$) for a worst case $2E_b/N_0$ ratio of 10.2 dB. The carrier jitter may then be written in terms of the loop filter bandwidth B_n using equation (2.5-34) as

$$\overline{\sigma_\psi^2} = 4.18 \times 10^{-4} B_n \quad \text{rad}^2$$

The rms carrier jitter is given by

$$\sigma_\psi = 2.045 \times 10^{-2} \sqrt{B_n} \quad \text{rad}$$

An rms reference carrier jitter of under 0.1 radians results in minimal degradation from the ideal case. Consequently, B_n should be upper bounded by 24 Hz. But for $\xi=0.707$, the natural angular frequency is only 45 rad/s. The two constraints cannot be met simultaneously. (This is analogous to the problem observed with the DLL design - the conflict of satisfying code frequency acquisition and tracking performance.) The use of a frequency searching device will alleviate the need for such a wide natural acquisition range. The reference carrier jitter specification of no more than 0.1 radians can thus be pursued. The choice of $w_n=50$ rad/sec (and a loop filter of $B_n=26$ Hz) was made.

A form of the closed loop frequency searching circuit shown in Figure 3.19 was used to aid carrier acquisition. The linearly ramped VCO control voltage sweeps the VCO output frequency across the band. When the two signals, $y_i(t)$ and $y_o(t)$, are close in frequency the difference signal following the LPF increases in magnitude. Once the threshold has been exceeded, the sweeping generator voltage is held constant while the loop locks up.

Since $y_i(t)$ and $y_o(t)$ are in phase quadrature when the loop is locked, a 90° phase shift is required for $y_o'(t)$ to be in-phase (or anti-phase) with $y_i(t)$. The rate at which the VCO can be swept in terms of the acquisition probability has been analyzed by Frazier and Page [21] for a PLL. Recognizing that the PLL in the squaring loop operates on the squared input signal, the search rate is given by

$$R_{90} = \frac{(1 - \sqrt{2/\rho_L}) \omega_n^2}{2\pi(1+d)} \quad [\text{Hz/sec}] \quad (3.7-2)$$

where R_{90} is the sweep rate for a 90% acquisition probability, ρ_L is the effective loop bandwidth SNR and $d = \exp(-\pi\xi/\sqrt{1-\xi^2})$ for $\xi < 1$ and zero otherwise. For $\xi = 0.707$, the sweep rate is simplified to

$$R_{90} = 0.543 \left(1 - \frac{\sqrt{2}}{\sqrt{\rho_L}}\right) \omega_n^2 \quad (3.7-3)$$

Figure 3.19 : Closed loop frequency searching.

The SNR in the 26 Hz loop filter for the worst case is given by

$$\rho_L = (4\sigma^2 \psi)^{-1} = 23$$

Consequently, the search rate for a 90% acquisition probability is 260 Hz/sec. The doppler uncertainty in the squaring loop requires a sweep over a 4 kHz range. At a sweep rate of 260 Hz/sec, the entire band is swept in 15.4 seconds.

The approximate natural acquisition time for a 50 Hz offset for the designed squaring loop is 0.56 seconds. Choosing a LPF with a corner frequency of 10 Hz for the lock detector will ensure rapid carrier acquisition when the threshold has been exceeded and the sweeping terminated. Once the loop locks up, the VCO output signal will remain synchronized should the input signal slowly drift through the band. Allowing for no more than a 0.5 radian phase error due to frequency drift of the squared input signal, the maximum frequency ramp tolerated by the loop is calculated to be 190 Hz/sec (see Figure 5.7 of [17]). Since the loop operates on the double frequency term, the input signal should not have a drift rate exceeding 95 Hz/sec if synchronization is to be maintained.

Using the second order loop of Figure 2.15 the time constants, τ_1 and τ_2 are calculated to be

$$\tau_1 = \frac{k}{\omega_n^2} = 352 \text{ seconds}$$

$$\tau_2 = \frac{2\xi}{\omega_n} = 0.028 \text{ seconds}$$

Choosing a capacitor value of 9.4 μF , the resistor values are

$$R_1 = 37.5 \text{ M}\Omega$$

$$R_2 = 3.0 \text{ k}\Omega$$

The squaring loop synchronization range (k) is the product of three loop component sensitivities. Using a Watkins Johnson M1 mixer, the phase detector sensitivity for 7 dBm input signals was estimated at $k_D = 0.67$ volt/rad. The VCO sensitivity was measured to be $k_{VCO} = 550$ Hz/volt. Since the squaring loop of Figure 3.18 uses the squared VCO reference, the effective loop VCO sensitivity is $k_3 = 2k_{VCO} = 1.1$ kHz/volt. To achieve a loop gain of 115 kHz, the amplifier gain (k_2) is specified as

$$k_2 = \frac{k}{k_1 k_3} = 156$$

The coherently demodulated signal is then lowpass filtered for bit recovery. A symbol synchronizer will optimize the BER performance. However, the relatively high SNR of 10 dB in the information band lessens the need for a symbol synchronizer. By not using a symbol synchronizer, the acquisition time associated with the circuit is also avoided. The demodulated signal is already essentially a binary signal following the LPF; hard-limiting this signal with a Schmitt-trigger device will thus cause minimal degradation in the expected BER. The savings in hardware and acquisition time is seen to be very desirable. The degradation in BER performance can be more than compensated for by using forward error correction. The complete squaring loop design including frequency searching, lock detection and bit recovery is shown in Figure 3.20.

A photograph of the squaring loop is shown in Figure 3.21. Since the bandpass filters and other components of the two channels were not identical, delay lines were required to minimize the static phase offset at the mixer inputs. Two delay lines were required - one for the PLL error signal, generated by the double frequency terms (21.4 MHz) and one for the coherent demodulation and lock detection signals (10.7 MHz).

CARRIER TRACKING AND BIT RECOVERY
 - SQUARING LOOP WITH FREQUENCY SEARCHING
 AND LOCK DETECTION
 - ASYNCHRONOUS BIT RECOVERY

Figure 3.20 : Squaring loop and bit recovery circuits.

Figure 3.21: Data demodulation circuitry.

False Lock

There are a number of mechanisms that give rise to false lock [2]. These include spurious oscillations, inordinate delays in the carrier synchronization loop, amplitude and phase modulation of the received signal due to antenna movement and data modulation distortion introduced by the narrowband filters. Whenever the acquisition range of the carrier synchronizer exceeds the symbol rate, the possibility of false lock exists. The equivalent situation occurs when the natural acquisition range is small, but frequency searching is used to cover a much wider bandwidth. This is indeed the situation for the timing beacon. Hence, the possibility of locking up to any of six data related harmonics is possible. The possibility of false lock, however, is observed to decrease as the filter bandwidth increases [2]. This observation should not be surprising - as the filter bandwidth is increased the modulation distortion is reduced, resulting in a decrease in amplitude of the symbol rate harmonic terms. For filter bandwidths that are of the order of 6 or 7 times the data rate, the possibility of false lock is extremely low. (Holmes [2] calculates the relative magnitudes of the harmonic terms and shows a false lock protection of 14 dB or more for the present case.)

The simplified system configuration is illustrated in Figure 3.22. The power spectrum of the received spread signal is shown at various points in the system. The complete system hardware realization, including the transmitter, noise source, receiver despreading and demodulating circuits and supporting equipment is shown in Figure 3.23. (Absent from the photograph but playing an integral part in the SNR, BER

Figure 3.22 : Simplified system schematic.

Figure 3.23: Complete system hardware realization.

and spectrum measurements and analysis are the HP8568A Spectrum Analyzer and the HP9826 Computer.)

3.8 THE USE OF AGC

The complete system design was based on the worst case expectation of the noisy 1000°K receiver and a 6 dB fade. A lower noise temperature will result in a lower noise power and, consequently, a higher SNR. A reduced fade will result in a higher signal power and a similar increase in SNR. Some practical solutions to the fluctuating signal and noise powers are discussed.

An AGC device should be used following the 2 MHz input filter to keep the total power (primarily noise) at a constant level. This will stabilize the false alarm rate at 10^{-4} . Since a lower noise temperature and/or a reduced fade will effectively give a higher signal power in the acquisition and tracking circuits, the implications should be addressed. The probability of detection will increase with an improvement in the SNR. The signal power at the envelope detector may become quite high such that the amplified detector output voltage may damage the voltage comparator. To alleviate this problem a 4.7 volt zener diode was used. The power in the tracking loop crosscorrelation signals will also increase with an increase in SNR. The tracking loop sensitivity will increase (i.e., the discriminator slope for small errors k_D will become steeper) and a higher error signal voltage will be observed for synchronization errors near $T_C/2$. A corresponding increase in the loop gain will result. Since τ_1 and τ_2 of the tracking loop filter are fixed

by their component values, an increase of k_D by a factor of x results in an increase in ξ , w_n and B_n by a factor of $\sqrt{x}$. In other words, the increase in SNR of the despread signal will not be as great as the input SNR increase.

The minimum expected SNR of the despread signal is 5 dB. Since the AGC at the receiver front end normalizes the signal power relative to that of the noise, a fluctuating input SNR will be observed at the squaring loop as a fluctuating signal power. An AGC amplifier should be incorporated at the squaring loop input to keep the total power (primarily signal information) constant and allow the noise level to vary. By keeping the signal power relatively constant the phase detector sensitivity will remain constant. Hence, the squaring loop parameters will not be altered by AGC. One would then expect to achieve the full benefits of an increase in the squaring loop input SNR in terms of jitter performance and BER.

CHAPTER 4

EXPERIMENTAL RESULTS

4.1 INTRODUCTION

The direct sequence spread spectrum transmitter and receiver of chapter 3 was built and tested. The receiver performance was measured over a wide range of effective input signal to noise ratios. Included in the measurements were the despread SNR, squaring loss and system BER. Numerous observations such as the filtered correlation triangle, filtered tracking loop discriminator curve and squaring loop carrier acquisition, to name a few, are also presented to help illustrate the dynamic performance of the system. Before presenting the experimental results and comparing them with theory, any deviations from the system design of chapter 3 must be acknowledged.

4.2 SYSTEM REALIZATION DIFFERENCES

Consideration must be given to the filters and, in particular, their equivalent noise bandwidths. The input filter was assumed in the design to have a bandwidth of 2 MHz. In order to obtain the desired group delay

response, however, a sacrifice in the filter roll-off had to be accepted.

The equivalent noise bandwidth of the input filter was calculated using numerical integration to be

$$B_n = 2.76 \text{ MHz}$$

Consequently, the effective SNR at the filter output is 1.4 dB lower than previously assumed. Although there will now be 1.4 dB more noise in the crosscorrelation signals, the SNR in the information bandwidth will remain as previously calculated. Of course, this assumes ideal despreading and demodulation, both of which will be slightly more degraded due to the additional noise.

The bandwidth of the narrowband filters was in excess of 3 kHz. In fact, numerical integration of the narrowband filter resulted in an equivalent noise bandwidth of

$$B_n = 3.225 \text{ kHz}$$

The effective SNR of the despread signal following the narrowband filter is hence reduced by 2 dB over the SNR of the 2 kHz bandwidth assumed in the design. The acquisition statistics under the worst case scenario will be much more pessimistic than previously calculated. Clearly, if the system is to operate well under the full range of expected SNR, 2 kHz filters must be obtained. The tracking loop performance will also be degraded due to the additional 2 dB of noise. Hence, it should be kept in mind that the system performance could be improved with filters that more closely match the system design. On a positive note,

the wider BPF will distort the correlation triangle less, yield a lower squaring loss and result in a lower code offset upon detection of code synchronization.

4.3 EFFECTIVE SNR IN THE SYSTEM

The worst case expected SNR at the output of the (ideal) 2 MHz filter was earlier calculated to be ~~-20.6 dB~~. With an equivalent noise bandwidth of 2.76 MHz, the effective input SNR for the worst case is -22 dB. The measured signal power following the input wideband filter was measured at -20.6 dBm. The level of the noise power competing with the signal was controlled by a stepped attenuator. Table 4.1 shows the measured effective input SNR as a function of the attenuator setting.

ATTENUATOR SETTING (dB)	MEASURED NOISE POWER (dBm)	SNR _i (dB)
16	-4.6	-16.0
14	-2.5	-18.1
12	-0.5	-20.1
10	1.5	-22.1
8	3.5	-24.1
6	5.5	-26.1
4	7.5	-28.1

Table 4.1 : Effective input SNR as a function of the attenuator setting.

Figure 4.1: Transmitter and noise source.

It should be noted that the worst case SNR_i is obtained with an attenuator setting of 10 dB. The transmitter and noise source used to generate the values in Table 4.1 are shown in Figure 4.1.

The power spectrum of the noise-free spread signal is compared in Figure 4.2 to the low SNR_i received spread signal. Figure 4.2(a) illustrates the case in which the SNR_i is 6 dB above the worst case expectation, Figure 4.2(b) illustrates the worst case expectation and Figure 4.2(c) illustrates the received spectrum for a signal that is 6 dB below the worst case expectation. Although acquisition at $SNR_i = -28.1$ dB (the third case) could not be accomplished, successful code tracking and carrier synchronization was achieved even at this extremely low SNR_i .

(a)

(b)

(c)

Figure 4.2 : Noise-free spread signal compared to received signal for (a) $SNR_i = -16$ dB, (b) $SNR_i = -22.1$ dB and (c) $SNR_i = -28.1$ dB.

With ideal despreading, the signal energy will be collapsed into a 600 Hz bandwidth and the SNR at the narrowband filter output should increase by 29.3 dB over SNR_i . (This neglects the BT_c correlation loss and noise spectral density reduction discussed in section 3.6.2.) The worst case SNR following the narrowband filter is hence upper bounded at 7.2 dB. Squaring and implementation losses are expected to reduce the filter output SNR by approximately 3 or 4 dB.

4.4 CODE ACQUISITION PERFORMANCE

The code acquisition circuit was observed to work very well down to an input SNR of -17.1 dB. The additional 2 dB of noise following the narrowband filter made acquisition at lower SNR_i quite difficult. The noise-free filtered crosscorrelation triangle resulting from code synchronization is illustrated in Figure 4.3(a). The photograph was taken at the acquisition comparator input (i.e. following the amplified output of the envelope detector). The adjacent picture, Figure 4.3(b), illustrates the amplitude of the crosscorrelation envelope for the noise-free despread signal without data. The triangular pulse is observed to be rounded due to the filtering. The desired 2 kHz bandwidth would introduce an even greater degree of modulation distortion and, hence, a more rounded crosscorrelation triangular pulse. Comparison of Figure 4.3(a) and (b) confirms the analysis of section 2.3.2; the peak amplitude of the triangular pulse reaches a value that is one-third below the tracked despread envelope amplitude. The width of the triangular pulse is 1 msec, the time required for a relative receiver code advancement of two chips.

Figure 4.3 : (a) Crosscorrelation triangular synchronization pulse and
(b) crosscorrelation of the despread signal without data.

Figure 4.4(a) shows that upon synchronization detection, the receiver code is 350 nsec ahead of the transmitted code. The steady state time difference between the codes is seen in Figure 4.4(b) with the receiver code generated 390 nsec after the transmitted code. The steady state time delay is attributed to the designed delay of the transmit code (reclocking the data and code) plus any additional group delay from the components. (The difference in group delay of the wideband filters could only introduce as much as a 10 nsec offset.) Consequently, the required pull-in range of the tracking loop is 740 nsec, or approximately $3/4$ of a chip. The use of a 2 kHz filter bandwidth is expected to increase the required pull-in range to a full chip.

Timing Waveform Diagram

Sample Period
Magnification
Magnify About
Cursor Moves
[↓]

1.000 μ S/div
10.00 nS/clock
350.0 nS o to x

(a)

Timing Waveform Diagram

Sample Period
Magnification
Magnify About
Cursor Moves
[↓]

1.000 μ S/div
10.00 nS/clock
390.0 nS o to x

(b)

Figure 4.4 : (a) Code offset upon synchronization detection and
(b) steady state code tracked propagation delay.

4.5 TRACKING LOOP PERFORMANCE

The tracking loop was able to maintain code synchronization down to an effective input SNR of -28.1 dB, 6 dB below the worst case scenario. An even lower SNR could be tolerated by the tracking loop if the narrowband filters used on the crosscorrelated signals more closely resembled the specified 2 kHz bandwidth. The tracking loop was also observed to obtain lock for code clock offsets as great as 2.8 kHz.

In Section 3.6.1 the tracking loop discriminator characteristic curve was derived from the static response of the envelope detectors. The characteristic curve shown in Figure 3.10 did not account for any filtering effects. The discriminator characteristic, also referred to as the S-curve [2], for the filtered crosscorrelation signals following the amplifier is shown in Figure 4.5 (for a 2 kHz search rate).

Figure 4.5 : Tracking loop S-curve for a 2 kHz search rate.

The slope for small errors was earlier deduced from the static envelope detector response as $k_D=0.12$ volt/chip. Since the amplifier has a gain of 39, the S-curve slope for $\epsilon=0$ should be described by

$$k_D k_2 = 4.68 \text{ volt/chip}$$

The slope at $\epsilon=0$ for Figure 4.5 is approximately 2.5 volt/chip. Hence, the loop gain is observed to be lower than the earlier calculated result. Fortunately, there was a sufficient margin provided in the initial tracking loop design to overcome a reasonable loop gain degradation.

Squaring Loss

The relatively flat envelope observed in Figure 4.3(b) occurs for the case of no data. A reduction in the envelope will occur at bit transitions due to the modulation distortion caused by the narrowband filters. Figure 4.6 illustrates this degradation of the envelope for a noise-free "data" signal with a 1:1 pattern. Recall that it is this modulation distortion that gives rise to the squaring loss.

The squaring loss can be closely approximated by measuring the despread SNR of a signal with random data and one without data. The difference in SNR between the two must be attributed to the squaring loss. Measurements over the 2 kHz doppler uncertainty range for a number of different SNR_i were taken. Two examples of the measurements are shown in Figure 4.7.

Figure 4.6 : Envelope of the crosscorrelated (despread) signal with a 300 bps 1:1 pattern.

The noise equivalent bandwidth that is often used for filters can also be applied to a data modulation spectrum. It can be shown that the total power in a $(\sin x/x)^2$ spectrum is given by

$$P_s = 0.89(R_D)S_0$$

where R_D is the data symbol rate and S_0 is the maximum spectral density of the signal (e.g. at $x=0$ for NRZ modulation). Consequently, the difference between P_s and the despread carrier power (no data) yields the squaring loss. At the filter center (no doppler) a squaring loss of 2.0 dB was measured while a squaring loss of 1.5 dB was measured at ± 1 kHz from the filter center.

(i)

(ii)

(iii)

Figure 4.7 : (a) Despread signals for $SNR_i = -17$ dB and $f_D = 0$

Figure 4.7 (continued) : (b) despread signals for $SNR_i = -15$ dB and $f_D = 1$ kHz.

Despread SNR

The SNR of the despread random data signal (SNR_d) can be deduced from the power spectrums of Figure 4.8. The noise spectral density (in a 1 Hz band) and maximum signal spectral density (in a 30 Hz band) following the narrowband filter are presented for several values of SNR_i .

(a)

(b)

(c)

(d)

Figure 4.8 : Despread data spectrum (5 dB/div) (a) $SNR_i = -16$ dB, (b) $SNR_i = -18.1$ dB, (c) $SNR_i = -20.1$ dB, (d) $SNR_i = -22.1$ dB,

(e)

(f)

(g)

Figure 4.8 (continued) : (e) $SNR_1 = -24.1$ dB, (f) $SNR_1 = -26.1$ dB and
(g) $SNR_1 = -28.1$ dB.

Table 4.2 summarizes these results and compares the measured despread SNR with the theoretically derived upper bound. Since the noise power following the narrowband filter is reduced below that at the input by 29.3 dB ($10\log[2.76\text{MHz}/3.225\text{kHz}]$) and a 2 dB squaring loss has been measured, the despread SNR will be upper bounded by an effective increase of 27.3 dB over SNR_i . The measured SNR of the despread data signal is denoted by SNR_d/meas while the theoretical upper bound on SNR_d is denoted by $\text{SNR}_d/\text{theory}$. The difference in SNR_d is shown in the last column of Table 4.2. The difference, primarily due to correlation and implementation losses, is approximately 2 dB, showing relatively good agreement with theory.

SNR_i (dB)	NOISE POWER (dBm)	SIGNAL POWER (dBm)	SNR_d/meas (dB)	$\text{SNR}_d/\text{theory}$ (dB)	ΔSNR_d (dB)
-16.0	-30.0	-21.0	9.0	11.3	2.3
-18.1	-28.0	-21.0	7.0	9.2	2.2
-20.1	-26.2	-21.0	5.2	7.2	2.0
-22.1	-24.2	-20.5	3.7	5.2	1.5
-24.1	-22.9	-21.0	1.9	3.2	1.3
-26.1	-21.3	-21.5	-0.2	1.2	1.4
-28.1	-20.4	-23.0	-2.6	-0.8	1.8

Table 4.2 : Comparison of measured despread SNR and theoretical despread SNR.

A confidence interval of the order of 0.5 dB on the measured SNR_d values would seem appropriate considering the resolution of the photographs.

4.6 SQUARING LOOP AND BER PERFORMANCE

Successful carrier synchronization and bit recovery was achieved for SNR_i down as low as -28.7 dB, 6 dB below the worst case expectation. The "bottom line" of a communication system performance is usually described by the BER performance measured in the expected environment. The BER performance will indeed be presented, but a look at the locking behavior of the squaring loop and the VCO upper sideband noise spectrum is first considered.

Figure 4.9 shows the carrier synchronization and resulting baseband signal, a 1:1 pattern, representing the 300 bps data signal. The photographs (taken for a relatively high SNR_i) represent the signal at the bit recovery input. When the VCO is not synchronous with the despread signal, a beat frequency signal results, as shown in Figure 4.9(a). As the frequency searching circuit brings the VCO output frequency to that of the input signal, the beat frequency is observed to decrease and the amplitude to increase (Figure 4.9(b) and (c)). Once the squaring loop "locks-up", the demodulated signal is simply the baseband data signal shown in Figure 4.9(d). Phase distortion on the beat frequency signal due to the preceding narrowband filter can be observed at the bit transitions.

Figure 4.9 : Carrier synchronization of squaring loop for high SNR_i
 (x-axis : 1 ms/div, y-axis : 2 V/div).

A software package has been developed by Hewlett Packard to measure the upper sideband (USB) noise spectrum (see AN-270-2). An example of the USB noise spectrum of the VCO with $SNR_i = -19.1$ dB is shown in Figure 4.10. A small noise peak can be observed at a 300 Hz offset, the data rate. Lower values of SNR_i increase the level of the USB noise spectrum, making the presence of symbol rate harmonics difficult to locate. Conversely, higher values of SNR_i result in a lower noise spectral density making the symbol rate harmonics more visible. Although the program calculates the jitter based on the measured USB noise spectrum, the noise power in the loop filter is effectively left out of the noise power calculation. This occurs because the routine can only calculate the noise spectral density for offsets starting at 50 Hz. But the loop filter has an equivalent noise bandwidth of 26 Hz and a large portion of the noise will hence not be included in the noise power integration. Consequently, the routine was expected to yield unrealistically low values of jitter. This was indeed observed to be the case; at the worst case SNR_i , a VCO rms jitter of only 1.3 degrees was recorded by the HP software package. The rms jitter for the worst case was deduced in section 3.7 to be of the order of 5.8 degrees. The BER results to be given shortly more closely agree with the predicted higher rms jitter. Thus, the USB noise measurements produce a useful illustration of the noise spectrum, but caution should be exercised before using the jitter results.

Figure 4.10 : USB noise spectral density of the VCO for $SNR_i = -19.1$ dB, 3 dB above the worst case expectation.

BER measurements were made using an HP1645A Data Error Analyzer with an HP9826 computer monitoring and recording the results. Although the results are for PSK modulation, the performance for DE-PSK can be accurately predicted from these results.

The system performance assuming coherent reception of the PSK signal is first deduced. The probability of error for a binary PSK system is given as

$$P(e) = \text{erfc} \sqrt{2\rho_d}$$

where

$$\text{erfc}(x) = (2/\sqrt{\pi}) \int_x^{\infty} \exp(-y^2) dy$$

and ρ_d is the SNR in the information bandwidth. Assuming a 2 dB squaring loss across the filter bandwidth, the effective processing gain for the otherwise perfectly despread and demodulated signal is given as

$$PG_{\text{eff}} = 10 \log[W_{ss}/2R_d] = 34.6 \text{ dB}$$

In other words, the worst case input SNR of $SNR_i = -22.1 \text{ dB}$ yields an upper bounded SNR in the information bandwidth of 12.5 dB. Table 4.3 tabulates the theoretical BER performance for a range of SNR_i assuming coherent despreading and demodulation.

SNR _i (dB)	ρ_d (dB)	BER
-29.1	5.5	3.91×10^{-3}
-28.1	6.5	1.39×10^{-3}
-27.1	7.5	4.04×10^{-4}
-26.1	8.5	8.50×10^{-5}
-25.1	9.5	1.23×10^{-5}
-24.1	10.5	1.1×10^{-6}
-23.1	11.5	5.7×10^{-8}
-22.1	12.5	1.3×10^{-9}

Table 4.3 : Theoretical BER bounds.

Measurement of the system BER were taken for a range of SNR_i and doppler offsets. Table 4.4 lists the measured BER for no doppler and a -1 kHz doppler offset. Recall that the squaring loss at a distance of ± 1 kHz from the BPF center was 0.5 dB lower than at the center. Assuming that the implementation and partial coherent demodulation losses are fairly constant throughout the filter bandwidth, one would then expect to observe that a 0.5 dB increase in SNR for no doppler offset is required to obtain the same BER for doppler offsets of ± 1 kHz. This is indeed the situation observed in Figure 4.11, a plot of the BER results from Table 4.3 and Table 4.4. A BER of 4×10^{-6} for the worst case scenario can be observed. But this is for PSK modulation - the predicted BER for the DE-PSK modulated signal is 8×10^{-6} . Convolutional coding would improve the BER performance by a few orders of magnitude. A BER improvement by at least one order of magnitude should be achieved with better packaging (PC boards instead of wire-wrap) and better matched components for both the despreading and demodulation loops.

SNR _i (dB)	BER FOR -1 kHz DOPPLER OFFSET	BER FOR NO DOPPLER OFFSET
-28.1	1.4×10^{-2}	2.2×10^{-2}
-27.1	5.7×10^{-3}	9.0×10^{-3}
-26.1	2.1×10^{-3}	3.5×10^{-3}
-25.1	7.3×10^{-4}	1.5×10^{-3}
-24.1		3.0×10^{-4}
-23.1		6.2×10^{-5}
-22.1		4.8×10^{-6}

Table 4.4 : Measured system BER.

For a given BER, a difference of 2.5 to 3.0 dB is observed between the theoretical upper bound and the experimentally measured values. Alternatively, it can be stated that as much as a 3 dB SNR degradation results from nonideal despreading, partial coherent demodulation and asynchronous bit detection. But the SNR degradation due to the despreading alone was shown in Table 4.2 to be of the order of 1.5 to 2.0 dB. Consequently, the SNR degradation from the squaring loop and asynchronous bit detection is only 1.0 to 1.5 dB. As earlier deduced, asynchronous bit detection for signals with an SNR approaching 10 dB and above in the information bandwidth gives only a slightly worse BER than that achieved with a symbol synchronizer.

Figure 4.11 : Measured system BER results compared with the theoretical upper bound.

CHAPTER 5

CONCLUSIONS

5.1 SYSTEM EVALUATION

A hardware realization of a spread spectrum timing beacon has been shown to satisfy the essential system requirements. This relatively novel approach of obtaining a high resolution timing reference from a low data rate (and, consequently, a low power) signal can thus provide a power- and cost-effective solution. The system performance was observed to be in good agreement with the theoretical predictions. Throughout the thesis, an attempt has been made to underline the many trade-offs and constraints of the various subsystems. While there are a number of routes one could take in attempting to satisfy the system specifications, the approach followed does offer a viable set of compromises.

Although PN code acquisition was not achieved for input SNR's near the worst case expectation, this was, in fact, predicted by theory. Code acquisition near the worst case can, however, be achieved using a narrowband filter with a 2 kHz bandwidth. An increase in the transmitter power from 1 W to 2 W would also help increase the code detection probability considerably, especially for the worst case SNR_i .

The code tracking loop was shown to work at exceptionally low SNR_i , certainly well below the worst case expectation. The pull-in performance of the code tracking loop upon switching from acquisition to tracking mode

was observed to behave extremely well. After adjusting the (tracking loop) amplifier offset voltage to give the proper VCC control voltage for the 40 Hz loop filter, the loop filter bandwidth reduction circuitry performed superbly. An initial code offset of $3/4$ of a chip upon detection of code synchronization was reduced to less than $0.1T_C$ within 400 msec.

Carrier synchronization of the despread signal was also shown to be achieved at SNR_i below the worst case expectation. The frequency searching and lock detection circuits enabled the carrier synchronizer to achieve both a relatively quick acquisition time and a low jitter carrier reference. It should also be noted that false-lock was never observed to occur, thus confirming the analysis of section 3.7.

A system BER of 4×10^{-6} (for PSK modulation) was measured under the worst case expectation, thus giving a DE-PSK BER of approximately 8×10^{-6} . This was achieved in the absence of a bit synchronizer and coding. It is felt that the initial system BER of 10^{-7} could be achieved with convolutional coding. A bit synchronizer would surely provide some improvement in the BER, but at the expense of more hardware and an increased acquisition time.

5.2 FUTURE STUDIES

Supporting circuitry to the existing system would provide for some interesting analyses. An AGC amplifier following the input (wideband) filter and one following the tracking loop should be incorporated. The effects of slowly- and quickly-varying fades on the system could then be measured. Additional circuitry to perform convolutional coding/decoding and DE-PSK modulation/demodulation would (hopefully) illustrate that a BER exceeding 10^{-7} could indeed be achieved for the worst case scenario. And, finally, the inclusion of a device such as a UART, that could transform the recovered bit stream to ASCII characters and monitor the 1 μ sec timing mark, would clearly illustrate the system's ability to provide a highly accurate users' uplink timing reference.

A more detailed experimental analysis of the various subsystems would be enlightening. For example, the despread SNR could be measured for a range of code tracking loop filter bandwidths. Similarly, the difference in system BER could be measured for a range of squaring loop filter bandwidths. The change in BER due to a change in the data LPF bandwidth could also be experimentally determined. Each set of experiments would help "peak" the performance of the individual subsystems and, consequently, improve the performance of the entire system. Experimental comparisons of the existing tracking loop and the sum/difference implementation would also be worthwhile to help illustrate the potential gain/phase imbalances in the two channels.

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