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PERFORMANCE STUDY OF THE LORAN-C SYSTEM IN THE PRESENCE OF WIDEBAND INTERFERENCE

A thesis presented for the Master of Science Degree at
The University of Tennessee, Knoxville

Joshua Emory Fox

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DEDICATION

This thesis is dedicated to my beautiful wife, Tara Fox. Without her support, this past year's tribulations and trials would have all but consumed me and this work would not have been completed. Her steadfast love helped me through the troubled times and made me maintain focus towards my goal.

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ABSTRACT

Loran-C is a navigation and timing system with a long and interesting history. In its near half-century existence, it has proven its worth against newer and more technologically advanced navigation system designs. From the work done to improve Loran's capabilities, we look at alternative means of providing navigation solutions to future users while maintaining the current infrastructure. Specifically, we investigate the design implications for a new potential overlay system and possible replacement to Loran-C currently under development at Oak Ridge National Laboratory. The implications for existing users, in terms of performance and reliability, are explored. Through laboratory experiments and computer simulations, we show that interoperability can exist with this new design and legacy Loran receivers with little or no modifications. We also give an overview of previous methods used in interference analysis and their applicability here for wideband PSK signals. Lastly, we suggest some additional work necessary to obtain better quantitative results specific to certain receivers.

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Chapter 1 Introduction

The purpose of this thesis was to design and test a new navigational system that could potentially act as a supplement to the Loran-C system currently in use worldwide. These signals were to exist in the same allotted band as for Loran-C and peacefully coexist. Our role was to conduct a performance study on the Loran-C navigation system in the presence of the proposed overlay signals. From initial laboratory testing, presented in Chapter 4, we were able to show that acceptable performance could likely be maintained for existing receivers with little or no additional processing. These new signals were to piggyback on the existing infrastructure and provide additional messaging and navigation timing for capable receivers. The possibility of eventual replacement was even discussed and considered in the signal design. We will not cover much detail concerning motivations for the study or design criteria, but will give some qualitative and quantitative analysis of the signals themselves.

We give a short treatise in Section 2.2 on current and future Loran-C trends. One of the motivations for this project was the fact that all updates and changes to Loran systems have neglected the fundamental limitations of Loran's signal shape. It is a pulsed system and information is embedded into the envelope of the signal. Channel distortion can greatly perturb this and receiver bandpass filters can even cause unacceptable errors here as well. For its time, the design was actually quite ingenious. Given the increased technological capabilities and computing power, we now can design and generate much more robust and advanced communication signals. Embedding critical information in the signals' envelope was one important factor in this determination.

Chapter 2 gives a brief account of Loran's history, including the origins of hyperbolic navigation. The basic operation of the Loran-C system and receiver functionality are given. Discussion then follows with the political and economical struggles Loran-C and its proponents have endured in the recent past, including the attempts by the navigation community to regain public support and continued use of Loran and its infrastructure.

Chapter 3 gives a review of the literature covering interference. Much of the published work was helpful in understanding the factors controlling how adjacent signals degrade Loran's accuracy and availability. For our work, this formed an important foundation that extended in the work presented in Chapter 4.

Chapter 4 gives laboratory and simulation data concerning the effects of introducing our proposed wideband overlay signals into the Loran band. Statistical analysis begun in [4] and Section 3.3 is continued in the simulation models here. The effects of filtering are the primary focus here also. Future work studying other receiver functions and signal structures would be beneficial. The inclusion of an accurate channel model would greatly improve the applicability and accuracy of results in later analysis. Many advanced simulation models designed by several of the authors referenced here would also be helpful if available to designers in the future.

Chapter 2 Loran-C Primer

2.1 System Model

Loran-C is a high-powered, terrestrially based, pulsed navigational system. It operates primarily on the principle of hyperbolic navigation. Although position location is its main function, Loran-C also serves many others roles, such as a timing reference [28]. We now give a basic model of hyperbolic methods and how Loran-C implements them.

In principle, hyperbolic navigation is a very simple idea. A hyperbolic system utilizing a Time Difference of Arrival (TDOA) method consists of at least three stations. There are other possible methods of deriving position solutions from the signals, but those are not covered here. The TDOA method is assumed. The Loran-C system is partitioned into 'chains' of three or more stations. Within a given chain, there exists one master station and two or more secondary stations, or slaves. All stations are synchronized to a common time source. Each emits its signal in a predetermined manner. All subsequent timing from the secondary stations is referred back to the TOA of the master signal. The receiver forms hyperbolic Lines of Position (LOPs) based on the TDOA for each master/slave pair within the chain. Each line represents all possible points in space which satisfy the difference in arrival times for the particular master/slave pair. The point in space where the LOPs overlap is the determined user position. All stations in Loran-C transmit, by convention, a sequence of eight equally spaced pulses and then hold for a predetermined period. This period is unique to each chain and is used for identification. Stations do not transmit simultaneously within a chain. There is a wait period between the successive transmission times for each station within a chain. Two terms of interest are Coding Delay and Emission Delay. They are defined here [39]:

Coding Delay (CD): the interval of time after reception of the master's transmission that a secondary station waits prior to transmitting its own signal. The coding delay is added to the baseline transmission delay in order to develop the Emission Delay (ED). The Coding Delays, the electrical lengths between the stations, and the lengths of the signal groups are used to develop the minimum allowed

Group Repetition Interval (GRI) for a chain. The Coding Delay assigned to each secondary station allows stations of a chain to transmit sequentially in time and to prevent overlap of the different signal groups anywhere in the system.

Emission Delay: the interval of time (in microseconds) between the beginning of the first pulse from the master station and the beginning of the first pulse from the secondary station in the same chain (both stations using a common time reference). The emission delay equals the sum of the baseline travel time plus the secondary coding delay.

The next term to be defined is the Group Repetition Interval (GRI). It, along with the delay terms defined above, sets the basic structure for the transmission timing of Loran-C pulses within a chain. The GRI definition is given [39]:

Group Repetition Interval (GRI): the time interval between successive pulse groups measured from the third cycle (or zero cross-over) of the first pulse of any one station in the group to the third cycle of the first pulse of the same station in the following pulse group. All stations in a chain have the same GRI, and the GRI expressed in tens of microseconds is the identifier for that chain and is called the chain "rate". GRI's may range from 40,000 microseconds to 99,990 microseconds, in increments of 10 microseconds.

A single Loran-C pulse [39] and its Fourier Transform are given here and plots are shown in Figures 2.1 and 2.2, respectively:

$$x_o(t) = A \cdot t^2 \exp\left(\frac{-2 \times 10^6 t}{65}\right) \sin(2\pi \times 10^5 t) \cdot u(t) \quad (2.1)$$

$$X_{0BB}(f) = F\left\{A \cdot t^2 \exp\left(\frac{-2 \times 10^6 t}{65}\right) \cdot u(t)\right\} = \frac{2A}{(\alpha + j2\pi f)^3} \quad (2.2)$$

$$X_0(f) = \frac{j}{2} (X_{0BB}(f + 10^5) - X_{0BB}(f - 10^5)) \quad (2.3)$$

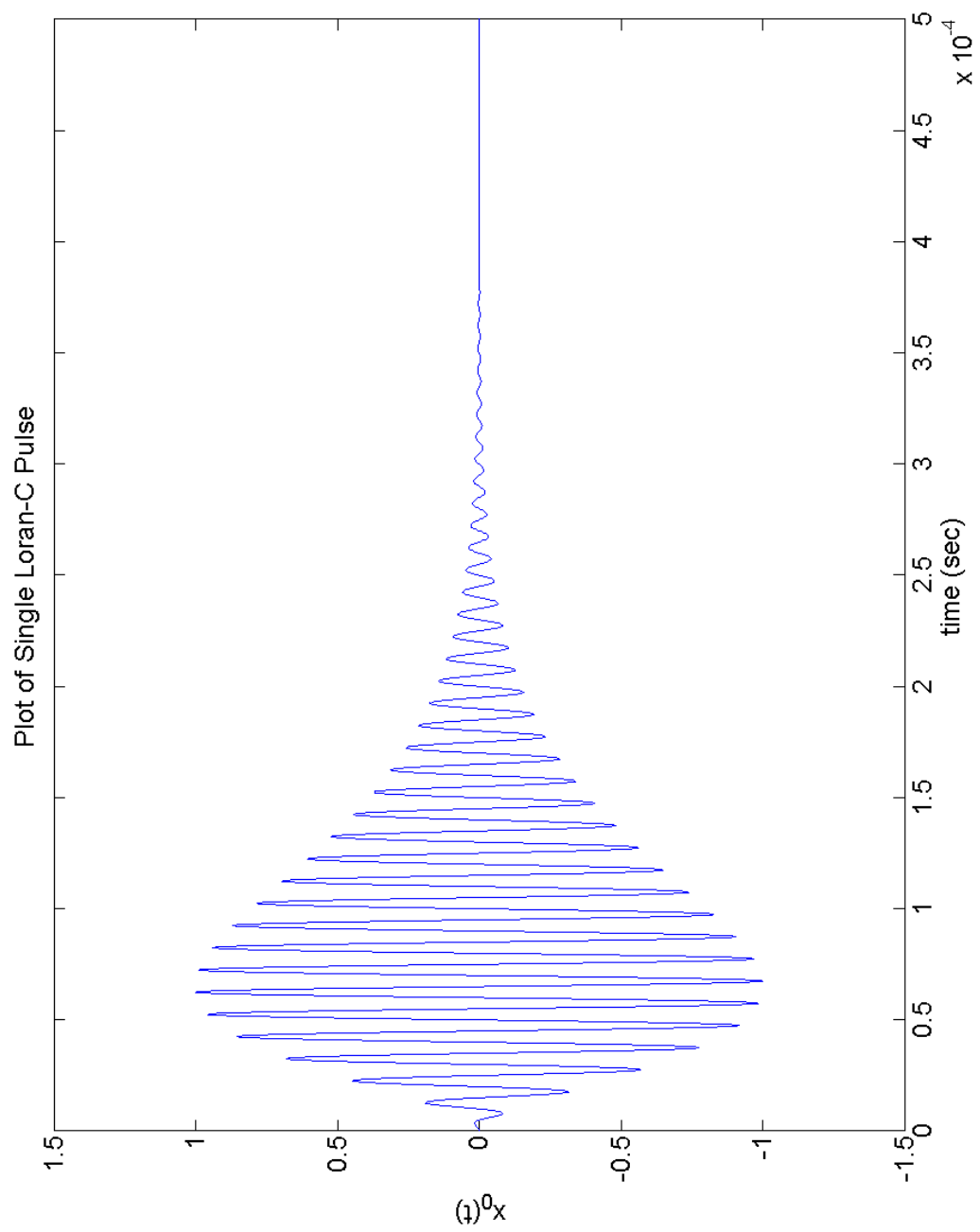


Figure 2.1 Plot of a single Loran-C pulse

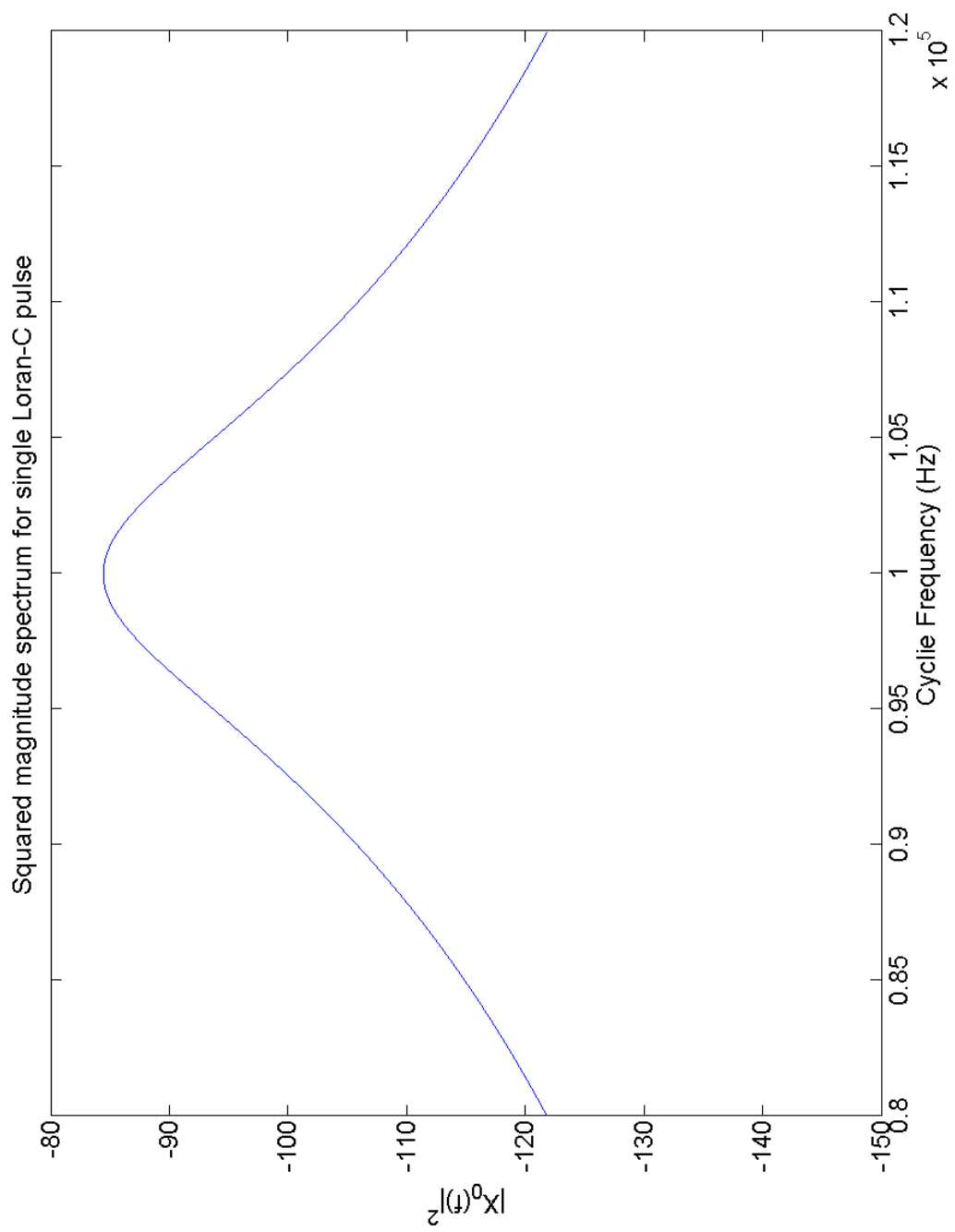


Figure 2.2 Energy Spectrum for single Loran-C pulse

where A is a normalization constant determined by the transmitter power, $\alpha = \frac{2 \times 10^6}{65}$ is the damping coefficient of the exponential function in (2.1), and for unity peak pulse amplitude, $A = \left(\frac{\alpha \cdot e}{2} \right)^2 \approx 1.75 \times 10^9$. The baseband pulse spectrum is given in (2.2), and its carrier modulated counterpart in (2.3). The pulses from each station are bi-phase coded by ± 1 . The phase code is identical for all secondary stations; master stations have a separate code. All codes repeat over 2 GRIs. The purpose of the phase coding is mainly to attenuate interference. Chapter 3 will describe this in more detail. Figure 2.3 illustrates the effects of phase coding on the spectrum of Loran-C. Slight spectral spreading occurs due to the effective BPSK modulation of the code. The definition for the entire Loran-C pulse train emanating from a single station is given by:

$$x_L(t) = \sum_{n=1}^N \left[\sum_{m=1}^{16} x_0(t) * \delta(t - t_m) pc(m) \right] * \delta(t - t_n) \quad (2.4)$$

where $x_0(t)$ is defined in (2.1), δ is the Dirac delta function, and $pc(m)$ is the phase coding. The time shifts t_m and t_n are defined:

$$t_m = \begin{cases} (m-1)T, & \text{for } 1 \leq m \leq 8 \\ (m-1)T + \text{GRI}, & \text{for } 9 \leq m \leq 16 \end{cases} \quad (2.5)$$

$$t_n = 2\text{GRI} \cdot (n-1) \quad (2.6)$$

where T is the 1 ms spacing between consecutive pulses in a pulse train. The spectrum of (2.4) is given by:

$$X_L(f) = X_0(f) \cdot \sum_{n=1}^N \sum_{m=1}^{16} [\exp(-j2\pi f(t_m + t_n)) \cdot pc(m)] \quad (2.7)$$

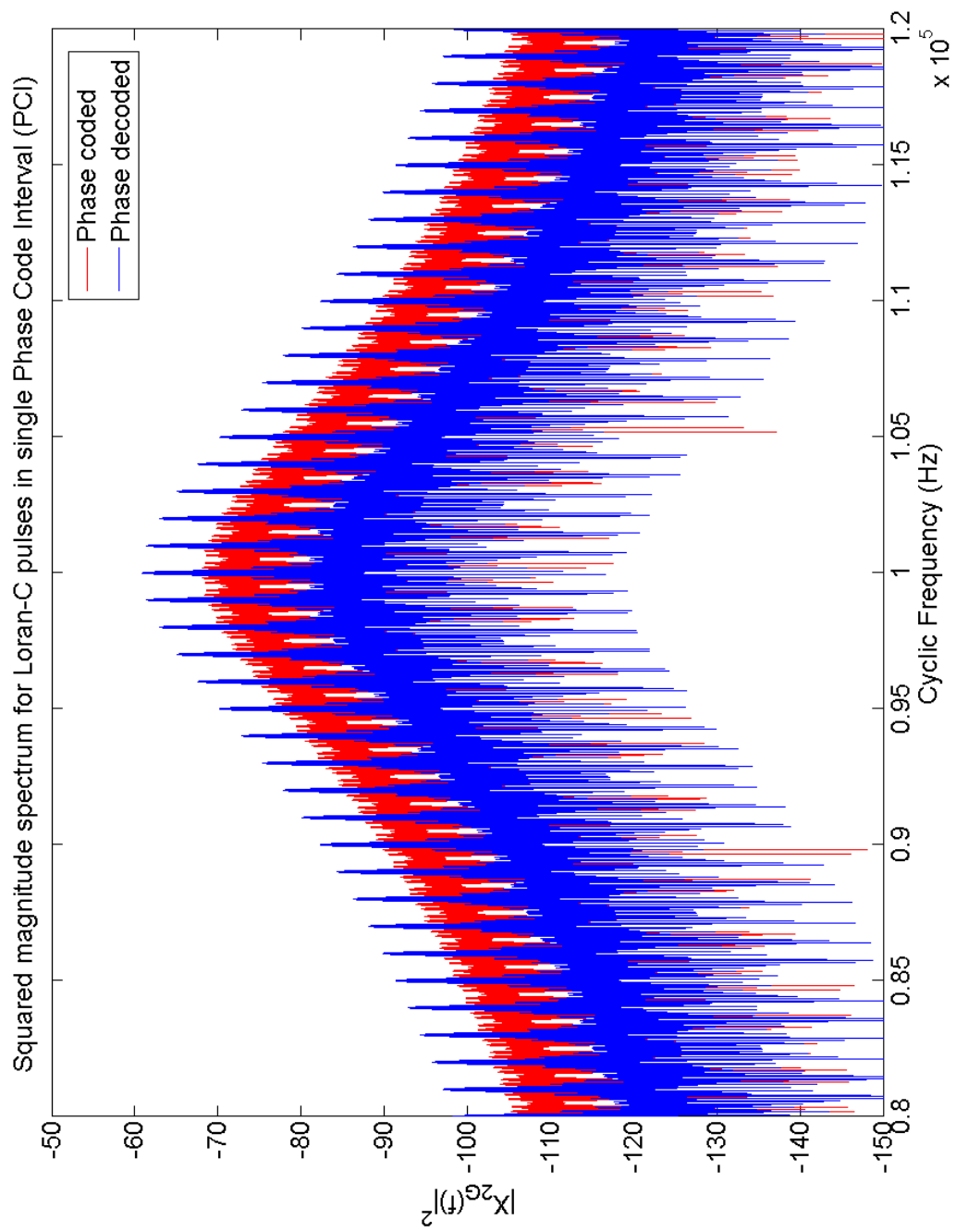


Figure 2.3 Energy Spectrum for pulses over 2 GRI (1 PCI)

The squared magnitude of (2.7) for $n=1$ is shown as the phase-coded portion in Figure 2.3.

Essentially, as the signal propagates, it encounters distortion of many kinds, including a nonlinear group delay [14]. The latter effect becomes more pronounced as the user's distance from the transmitter increases. For completeness, it is included here but left ideal in later analyses. As stated, each chain has a unique identifier, its GRI. Since all Loran-C stations transmit with the same carrier, interference from neighboring chains necessitates the use of this quantity [16].

Loran pulses arrive at the receiver along many paths. The path of primary interest is that of the groundwave. Due to a Loran-C signal's large wavelength of 3 km it can diffract across the earth's surface and travel great distances, penetrating even large obstructions with ease. However, this path is quite different from free space propagation. Due to the variations in the propagation characteristics, spatially and diurnally, a level of uncertainty is created. This translates to a limit in the absolute accuracy of Loran-C [15].

Typically, two factors are used to arrive at the correct propagation time. The primary factor (PF) is the increment of time for traversing an all seawater path. Second is additional secondary factor (ASF), which accounts for propagation delays over heterogeneous earth. PF is solely dependent on distance while ASFs need to be measured or modeled. More accurate models or measurements result in more accurate range measurements and, hence, more accurate position solutions. [14]

The other paths afforded Loran-C signals are by means of ionospheric refractions. These received signals are termed 'skywaves' and are part of nominal system operations. In fact, part of the design of the pulse rising edge was to reduce the likelihood of skywaves arriving early enough to prevent proper carrier tracking. These are less desirable due to their instability. The ground path, however, is and the geometry is greatly simplified in its case. Unless otherwise stated, we assume exclusively a groundwave path for signals of interest.

We present a brief set of additional definitions [20] of other system parameters to better understand the basic mechanisms of Loran-C operation:

Standard Sampling Point: the point on the Loran-C pulse envelope 25 microseconds after the beginning of the pulse to which far-field field strength calculations or measurements are referenced. For the standard Loran-C pulse with 0.0 ECD, the amplitude at the standard sampling point is 0.506 time the peak amplitude.

Standard Zero Crossing: the positive zero crossing at 30 microseconds of a positively phase coded pulse on the antenna-current waveform. This zero crossing is phase-locked to the Loran-C station's cesium time reference. The standard zero crossing is used as a timing reference for measurement of Loran-C signal specifications.

Time Difference (TD): the interval in time between the receipt of a master station's signal and secondary station's signal from the same rate. Controlling Standard Time Difference (CSTD) is the reference standard against which the control station compares the observations of the System Area Monitor.

Signal to Noise Ratio (SNR): the ratio of the RMS amplitude of the Loran pulse at the standard sampling point to the RMS value of the noise present at that time.

Nominal ECD: the calculated ECD held at the transmitting station which, given the identical propagation conditions which existed during the chain calibration, would result in the assigned CSECDs being observed at the monitor sites. The Nominal ECD is determined from the transmitting station's antenna current waveform.

Blanking (Priority, Alternate): the suppression of pulses on one rate due to the periodic phenomenon that occurs when a dual-rated transmitting station has to transmit two pulse groups of different rates at the same (or nearly the same) time. During the period of overlap, the one rate's pulses are suppressed. Priority blanking occurs when the same rate is always blanked while alternate blanking occurs when the two rates are blanked in an alternating manner.

Controlling Standard Envelope-to-Cycle Difference (CSECD): Envelope-to-Cycle Difference that is maintained at the monitor site as determined by chain calibration.

The basic Loran-C signal and system structures have been described. One last parameter of importance is the Envelope to Cycle Difference (ECD). It represents the offset of the carrier to the pulse envelope from the ideal case. It is assumed zero but is essential to understand its effects on timing results. Chapter 3 gives greater detail of this parameter.

2.1.1 Receiver Basics

A review of basic Loran-C receiver basics is given, including those functions which are common to all and which form the basis of essential receiver operation. A diagram of a basic Loran-C receiver is given in Figure 2.4.

At the receiver front-end one can see a set of filters for initial noise reduction and interference mitigation. The bandpass filter (BPF) has traditionally been taken as a 5th order Butterworth whose passband is taken as 90-110 kHz. This filter type was adopted due to its maximally flat pass band amplitude, and approximately linear group delay within the passband [2].

Loran-C is a pulsed navigation system. It relies on the pulse rising edge for tracking purposes. This filter puts an upper limit on the pulse rise time and nonlinear phase is prevalent outside the passband. Digital filters offer complete linear phase solutions which are very important for timing calculations. Analog filtering would still be desirable to prevent aliasing, minimize noise hazards, maximize dynamic range by rejection of adjacent band signals, etc. The bandwidth of the analog filter could be relaxed in order to further minimize its effects on the Loran-C pulse while maintaining interference and noise rejection. It is the phase, not the amplitude, which is critical in maintaining acceptable timing performance. By minimizing the dependency on analog filtering, parameter drift due to temperature and time variations in the components will have less effect on the Loran-C pulses. Thus, greater possible accuracy and consistency may result.

The next block is the notch filtering. Depending upon the implementation, these may be fixed by the user or adaptive in nature. These filters are set to reduce the most harmful interfering signals. They can be implemented in analog, digital, or both forms. As will be shown later in Chapter 3, some signals can cause more problems than others, even at much lower amplitudes. Next is the carrier tracking loop. The first 35-65 μ s of the rising pulse edge are generally used for tracking and timing.

Traditionally the timing point on each pulse is the 'standard zero-crossing', the third positive-going zero-crossing of the 100 kHz carrier, 30 μ s after the start of the pulse. However, many receivers make their time measurements at later zero crossings. [6]

Later in the pulse, skywave presence makes it difficult to obtain accurate

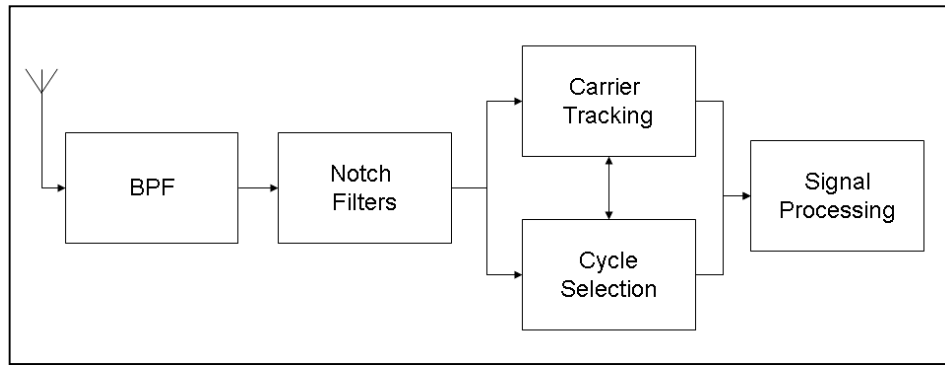


Figure 2.4 – Block Diagram of Loran-C Receiver

measurements [6]. The trailing edge of the Loran-C pulse is not employed for tracking. In fact, the definition in (2.1) may not always hold. The trailing edge shape is primarily for spectral containment. Its exact shape may vary from chain to chain [20]. Depending upon the application, the tracking bandwidth of the carrier tracking loop will vary, but it is generally modeled as a second order digital phase-locked loop (DPLL). However, in some cases, a first order PLL may be used for stationary tracking [9].

The adjacent block performs a similar operation on the received envelope. The cycle selection block is responsible for analyzing the receiver envelope and determining the proper cycle which is to be tracked in the phase tracker loop above. The Loran-C pulse's shape was designed to yield a unique gradient at each point of the rising envelope edge. To take advantage of this, measurements of the cycle peak amplitudes about each zero crossing are taken and are compared against stored values. This process is critical for performance because if it fails, a cycle slip would produce a 10 μ s offset in the reported TOA resulting in approximately a 3 km bias for that particular station, with an average total bias of 1.5 km from the correct position. This is one disadvantage of Loran. Information is also embedded into the pulse amplitude. One comparison of this modulation type to GPS signaling would be between AM and FM radio.

Fundamentally, Loran-C will degrade like AM, in a linear fashion over distance. CW signals like GPS, are less dependent on the amplitude, and will generally follow a threshold-like effect, similar to FM radio, over distance.

In addition to these operations, a phase decoder is present in each block which attempts to decode the Loran-C component and return it to its original state while encoding the interference component. The receiver then makes estimates of its position from the resultant of these operations. The basic operations of a Loran-C receiver have been given.

2.1.2 Advancements in Receiver Design and Performance

In addition to the fundamentals, here follows a brief discussion of some of the improvements and advancements in Loran-C receiver design. Until quite recently, Loran-C receivers were capable of only tracking a single chain at a time. Recent advancements in digital signal processing, computing power and hardware costs

have progressed receiver capabilities to the point where “all-in-view” reception is now capable, where multiple chains can simultaneously be tracked [10]. In North America, 20-30 transmitters can often be accurately tracked [12]. With the increased availability and lower cost of memory, receivers can now also contain large databases with expected coverage reporting, more accurate secondary correction factors and other data which can better control receiver decisions. The coverage reporting is especially important in fringe areas where transitions in the ranking of incoming signals can play a role in the reported user position [10]. A close parallel to cellular systems and the advent of handovers exists in this respect. Prior to handover capabilities, a user would continue a call until the particular cell in use could no longer support voice transmission. The user would then have to reacquire in the new cell and begin a new call [34]. The same problem existed for early Loran-C receivers. Manual resets of the desired chain were required and were quite troublesome to users. There are many advantages of increased chain reporting in the navigational solution. First, diversity in geometry is afforded to the user. As the LOP crossing angle for two station pairs decreases, the user’s position uncertainty grows along the axis of that angle [10]. With multiple chains, the uncertainty region of user position generally diminishes. This is especially helpful in areas about the baseline for a given station where noise and interference could have caused the most trouble. Second, the total availability increases which will be shown to be quite important for aviation and other applications. Within a chain or Loran-C stations, a single outage can be disastrous for the user. If possible, manual reassignment of a further chain may be possible but would be troublesome at best. With automatic adjustments and multichain solutions, the total degradation in performance is less for contemporary receivers. Lastly, the redundancy in reporting of user position allows for more advanced processing. Using the databases available of correction factors, one can weight the contribution of each station adaptively, according to confidence, to the total position solution. This helps greatly in obtaining increases in both repeatable and absolute accuracy. More details of performance enhancements with respect to interference mitigation will be given in Chapter 3.

A list of Improvements in Loran-C receiver performance from [23] is given:

1. *Signal to Noise (SNR) improvement over hard limited receivers exceeding 20dB.*
2. *Short term root mean square (RMS) noise reduction of 3-5x.*
3. *Envelope to cycle difference (ECD) noise reduction of 4-5x, which typically reduces averaging times 25x and eliminates cycle slips.*
4. *An increase in range (and associated coverage) of approximately 300km.*
5. *At least one order of magnitude of increased availability, even in urban environments.*
6. *Virtual immunity to noise interference and burst noise.*
7. *Use of multichain, multistation navigation to improve absolute and repeatable accuracy in marginal coverage areas.*
8. *Improvement of statistical jitter in timing applications.*

A list of hardware improvements which result in performance gains from [23] is given:

1. *Linear, digital receivers with DSP chipsets.*
2. *Adaptive digital filtering through DSP.*
3. *Smaller, active H-Field antennas.*
4. *Increased processing speeds.*
5. *GPS integrated antenna and receiver.*

2.1.3 Recent events in Loran-C

Over the years, the role and status of Loran-C have both changed tremendously. Since its inception, some new navigational systems have come into operation which many feel make Loran-C an obsolete and dated system. Satellite-based GPS and GLONASS provide worldwide coverage to users, a capability Loran-C has never had. In addition to its critics, Loran-C has had many proponents for its continued use. An excerpt from [10] provides a somewhat detached view of the turmoil concerning the potential phase-out of Loran-C operations over the past few years:

As each generation of navigation systems has been developed and deployed, there have been concerns that the preceding generation system not be prematurely abandoned. Although each succeeding generation of navigation system has ultimately proven to be an overall

improvement upon its predecessor, there have been areas in which the improvement may not have been as dramatic. These areas have often led to disagreement regarding the transition decisions to be made. [10]

The development and growth of GPS in the United States has created many research dollars. Its proponents have been the major driving force in transitioning to GPS in areas like timing control and even for use as a sole means navigational solution. The public was made aware of this decision after publication of the DOT FRP in 1994. It stated that, by the year 2000, Loran-C operations in the United States would cease [10]. This came as quite a shock because the last FRP in 1992 claimed continued support for U.S. Loran-C stations at least through 2015 [10]. One interesting note is that the FRP is actually privately distributed to the JCS under the title "Master Navigation Plan" much earlier than the public release. We cannot be sure, but, it is suspected that news of this magnitude made it difficult to keep under wraps for long. "Such controversy surrounded the 1994 FRP that it was delayed release until summer 1995 [10]."

There were several results of this decision. First, commercial operations became almost non-existent. Many companies who manufactured Loran-C equipment quickly cut their losses and shifted into other ventures when capable. Second, research into improving and upgrading Loran-C waned because of the prospect of its impending demise. Third and most importantly was the immediate and outspoken campaign of Loran-C supporters to reverse this decision. The spectrum of those against this decision was quite wide. Those remaining in commercial Loran research and manufacturing were opposed for many reasons. Their livelihoods, which seemed quite secure two years prior, suddenly came into question. They also believed Loran-C to still be a viable system. Many from this group claimed that the system's true capabilities were not given full consideration; the majority of public knowledge concerning Loran-C was based on 10-20 year old technology and did not represent the system as it was then [27]. Others included groups of navigators who felt that the short term transitional costs for their end use would far outweigh any advantages GPS could provide, especially as it existed in 1994. For navigation use, sole means GPS was highly undesirable. Another level of redundancy was desired,

regardless of the primary means. Although its absolute accuracy was far greater, GPS could not compete with the high repeatable accuracy of Loran-C.

The main driving force behind reversing the U.S. DOT's position was the International Loran Association (ILA) [10]. They helped direct the tremendous letter writing and petitioning which eventually made some take interest. As one can see now, 2000 has passed and Loran-C is still in service in the U.S. The ILA communicating with Congress was a key for approaching their goal. Congress took note and, in 1996, ordered the U.S. DOT to reinvestigate the viability of Loran-C for navigational and timing purposes. The study [10] did show there were many potential hazards in solely utilizing GPS as a navigation source. One other deciding factor was the growing popularity of Loran-C worldwide, especially in the European community. The United States' popularity would have greatly waned within the navigational community if the 1994 FRP came to fruition. Since that time, many proposals for the continuation of Loran-C have been made and tested. One example is the Eurofix system. A brief overview of it is given in the next section. One last note of importance is the need for a long term plan detailing the role of Loran-C worldwide. The fear surrounding its possible demise has lessened commercial investment due to high risk. It must be better understood where Loran-C fits into the plans for worldwide navigation and timing in the future [10]. Systems like eLoran [22] and Eurofix [15] are a big step towards this aim.

2.1.4 Eurofix

We now give a brief description of the Eurofix modulation method and its capabilities. Ternary pulse position modulation is applied to the last six pulses of each Loran-C stations pulse train. A zero symbol indicates no change in the pulse. A $1\ \mu s$ shift in either direction represents one of the two antipodal symbols, ± 1 . This modification of the TOT (Time of Transmission) for the pulses is done in a balanced manner. It is illustrated in Figure 2.5. The purpose is solely to preserve maximum energy for existing Loran-C receivers not capable of demodulating the signals. This operation has no effect on how a Eurofix receiver operates. The modulation was designed to provide minimal loss to nominal Loran-C integrity. In fact, an average reduction of 0.79dB is reported [15].

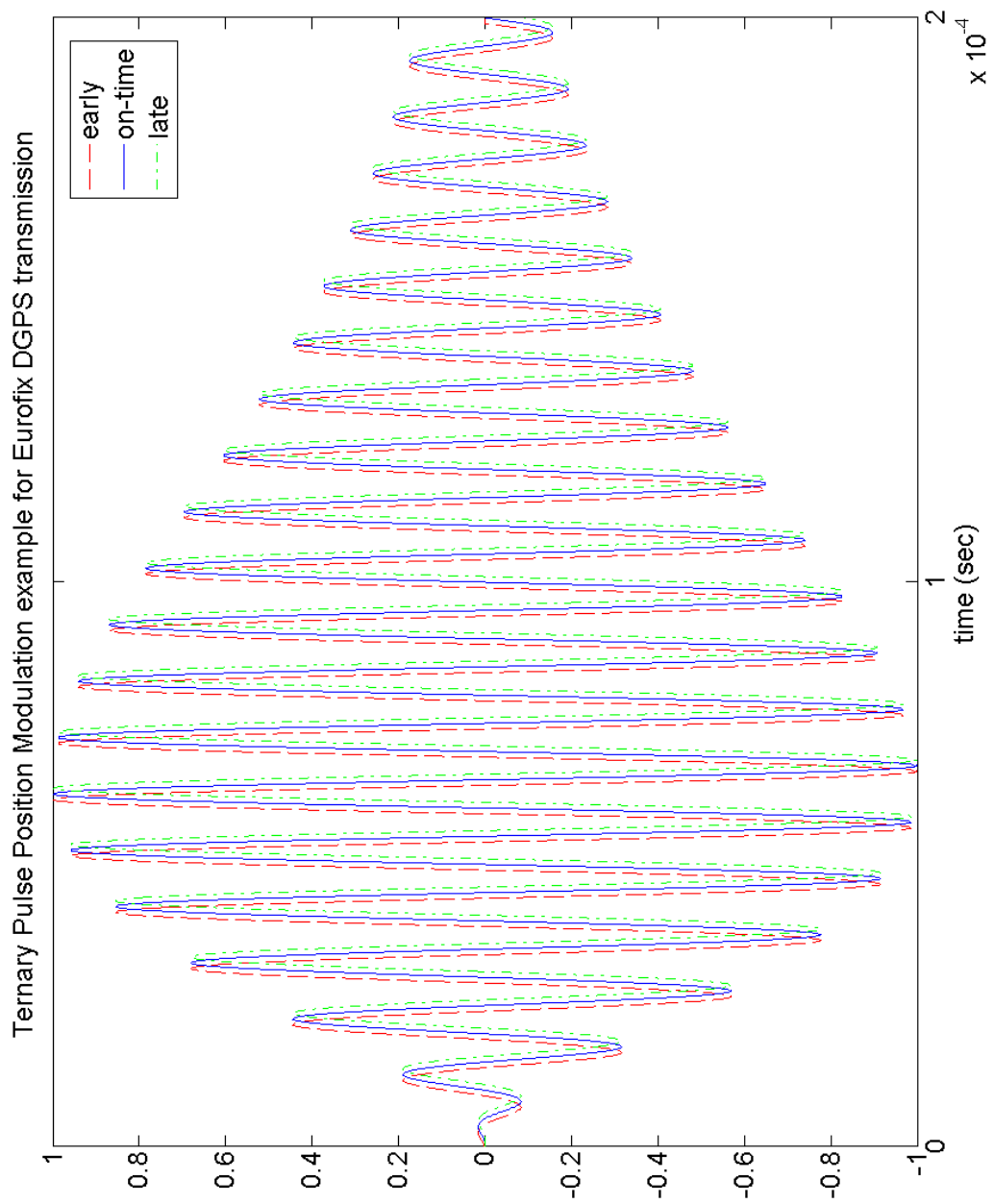


Figure 2.5 – Illustration of Ternary Pulse Position Modulation in Eurofix.

Compared to the contributions from inherent system degradations like blanking and CRI, this effect is very minor [13]. Prior to design of the modulation scheme, the following restrictions were set to ensure minimal degradation:

- 1. The blinking service must be preserved, which excludes the first two pulses of each Loran-C group of Eurofix modulation.*
- 2. The modulation is not allowed to induce tracking biases, which requires a balanced type of modulation.*
- 3. The modulation index must be kept small in order to prevent an undesirable loss in tracking signal power.*
- 4. The blinking service must be preserved, which excludes the first two pulses of each Loran-C group of Eurofix modulation.*
- 5. The modulation is not allowed to induce tracking biases, which requires a balanced type of modulation.*
- 6. The modulation index must be kept small in order to prevent an undesirable loss in tracking signal power.*

Prior to modulation, the data bits are channel encoded using a Reed-Solomon code and final frame is in RTCM (Radio Technical Committee for Maritime) format consisting of 56 bits. This channel code works well in ternary signaling and resists burst errors from CRI and blanking well. Its performance also fades gracefully as noise and interference increase. This method is quite similar to the format the USCG uses for its DGPS messaging with the exception of station identifiers. The Eurofix system is also capable of transmitting GPS integrity messaging to help monitor the primary navigation solution through an independent source [12].

There are many positive prospects of the Eurofix system for potential users. First, Loran-C can be used as a backup navigation system for GPS in the event of its failure. Second, added overall redundancy results since both systems can be treated as relatively independent. GPS is a high-frequency, low power, satellite based system while Loran-C is low frequency, high power, and terrestrially based.

“The two are also complementary [15].” GPS is known for high absolute accuracy but large variance about its mean. Loran-C, however, has more consistent results but generally with a much greater bias. One may use the DGPS corrections to calibrate out the bias of a given Loran-C position solution and update its ASF database accordingly. For GPS outages of minimal time span, the corrected Loran-C solution would be able to maintain acceptable accuracy until satellite acquisition can be reacquired. For Eurofix messaging, only the relative TOA of each pulse to a standard pulse is necessary. One does not need the absolute TOA of each pulse. Thus, with only a single Loran-C station equipped with Eurofix messaging, one can obtain differential corrections to the user’s satellite navigation system. This means that the data link can go beyond the given coverage area for the Loran-C service [13].

2.2 Loran-C History

2.2.1 Hyperbolic Origins

The first proposed hyperbolic navigation system was invented in 1930 in Germany by Dr. Meint Harms. Strangely enough, the system bore a striking resemblance to the current Decca Navigation System now operating in Europe. It is believed that secrecy surrounding these types of patents during this period precluded any possible false claims and that they were developed independently. Three systems, based on hyperbolic methods, were actually designed and built by 1940. They were to be known as Gee, Loran, and Decca [31].

2.2.2 System Development

Although Loran-C did not officially exist until 1956 or go on air until 1957, it had been operating under the name Cytac since 1952. That system, nearly identical with the exception of very few modifications, was developed privately for the United States Air Force as a tactical long range navigational system [30]. At that stage of development, many improvements were made over previous Loran incarnations. The first main difference was the change to a carrier frequency of 100 kHz. The signal structure was also changed to the chain of phase coded pulses [31].

Loran-C offered several advantages over its predecessor including improved functionality, considerably more accurate positioning and longer range.[10]

These benefits will be discussed in greater detail in Chapter 3. The U.S. Coast Guard took over in 1956 when the Air Force lost interest and by 1957 had the first Loran-C chain operational with an approximate coverage area of 700 square nautical miles. Interest grew and the system began to expand over the next few years. By the early 1970s over 30 stations were operational, covering the Mediterranean, Hawaiian and Alaskan areas, Northeast Atlantic, and Southeast Asia. Also, at this time the U.S.S.R. had installed 8 independent but Loran-C compatible stations as well. A decade later Loran-C had officially been given the task of providing navigational coverage for the U.S./Canadian coastal confluence. It took much cooperation with the Canadian government to accomplish this goal. They even provided some of the additional stations needed to satisfy coverage requirements. By this time, some 41 official Loran-C stations were operational. In the early 1990s, the European community began to develop Loran-C chains, under initial guidance by the U.S.C.G. Much work has been done in Europe to advance Loran-C. As newer navigation systems have been implemented, Loran-C has shown to be competitive in performance and availability [20]. The advent of GPS and other satellite-based systems has not totally eliminated Loran. It has proven to be a good complementary and redundant system [27, 15, 20].

Chapter 3 Interference to Loran-C

This chapter provides an overview of the fundamental limitations of the Loran-C navigation system. There are many possible error sources, which can be either natural or manmade. We define natural interference here as any error source inherent to nominal Loran-C system operations, including any naturally occurring phenomena. Man-made interference is defined as any error source seen by a Loran-C receiver which is generated by a man-made source external to the system. A CWI navigation system like DNS is an example of man-made interference. Both types of error sources are presented. Analysis of the latter forms the basis for the work in Chapter 4. Thus, it will be given greater attention than effects of natural interference.

3.1 Natural Interference and Internal Error Sources

We present here a brief section discussing natural interference and its effects. There are many non-idealities present in Loran. Some may potentially contribute great error and availability hazards for users. Much attention has been given to understand and control these effects [5]. A list of the most prominent natural hazards to performance is given in Table 3.1. The cause for each source is listed along with different methods used to combat them. They are then given greater discussion.

One can observe from the table that, for each error source, there exist methods to combat their negative influences. Although it may seem unusual that it is so susceptible to interference, Loran-C is quite a complex system which has matured from decades of research, development, and design experience. Through time and technological advancement, the capabilities of the system have grown beyond what its designers envisioned [5]. However, it still has limits that need addressing in order to remain a viable navigation and timing source by current standards.

Table 3.1 – Natural Impairments to Loran-C -- Source: [5]

Error Source	Cause	Compensation Technique
1. Transmitter timing fluctuations	Cesium, timer, and transmitter variations	Accurate and stable time base frequency, phase adjustments on short- and long-term basis, cycle compensation loop
2. Temporal fluctuations	Refractive index changes along propagation path. Surface impedance variation along propagation path	Differential Loran-C and variations of this method
3. Spatial Effects	Bridges (such as Golden Gate); buildings; terrain elevation (islands, peninsulas in vicinity of harbor, river, etc.)	Conduct grid survey. Reflect warpage in grid. This is a one-time fix. Use position reference system or visual grid survey methods (PLAD)
4. Noise (atmospheric and manmade)	Electrical discharges in the atmosphere and power generation equipment	Band limiting and switched O in the receiver. Linear: filtering done at low-level ahead of amplifier and clipped linear amplifier. Hard limiter: all linear processing at low-level output has square-wave shape. Signal processing filters to minimize effects of interference and noise, shape the envelope, and minimize unwanted distortions. Narrowband switching of the filters is provided to gain SNR
5. Receiver	Error measurement technique	Linear and hard limiter amplifiers have wideband amplifier with low internal noise
6. Skywaves	Multipath	Loran-C pulse shape and signal processing
7. Cross-Rate Interference	Loran-C signals from multiple chains	Blanking pulses and actively separating signals

3.1.1 Transmission Errors

Errors are introduced through the transmitting facilities which place an upper limit on accuracy. Receiver techniques have advanced to the point where these limits are now approachable. To go beyond them, Roth [12] claims that upgrades to transmit facilities will be required. Further receiver capabilities will otherwise be thwarted by the transmission fluctuations in timing. The following error sources are present in transmitting facilities for Loran-C pulse generation: timing synchronization, pulse shape control, phase control, and parameter drift [5]. Until quite recently, all Loran-C stations were operating with Cesium frequency standards which were quite problematic. The standard in question is a Hewlett-Packard 5061A Cesium Beam Atomic Frequency Standard [5]. This essentially controls the time for pulse transmission. Significant errors here can be harmful to performance. However, there are methods to combat these effects. Both short-term and long-term corrections to drift in timing are made. It should be noted that each clock has a minimum setting of $\pm 10^{-13}$ Hz between settable frequencies. Also, the frequency of Cesium clocks varies randomly once set. Corrections can be quite difficult to make. Phase noise in both the clocks and the timer are also problematic [5]. It has been estimated that the short term variations of timing from these particular Cesium clocks and the timer together are approximately 10ns rms (root mean square). The combined effect with the transmitter itself is an estimated rms equipment error of nearly 15 ns [5]. Under legacy conditions, these variations were not observable [17]. This is before the signal has even entered the 'channel'. It is hoped that the current and future upgrades to the transmission facilities will help to alleviate this problem and allow for greater accuracy, absolute and repeatable.

3.1.2 Channel Errors

Accurate TOA reporting requires an accurate model of the path between transmitter and receiver. The fluctuations in time delay are the result of variations in propagation conditions. These propagation errors can be spatially and temporally distributed. The total deviation tends to increase with distance. The speed of the propagating pulses, and hence their time of arrival, depends upon several electrical properties of the groundwave channel. Briefly, these are surface impedance, surface

roughness, and the refraction indices of the atmosphere near and above the earth's surface [5]. These changes have been combated primarily with ASF corrections. ASF corrections adjust for variable propagation speed of both carrier and envelope over the channel. These effects are nonlinear in nature and are further complicated by large path obstructions like bridges and small islands, which can greatly distort the signal. The ASF corrections are a best fit, in linear terms, to the variations from ideal in the received waveforms. Williams [32, 33] has done much work in this area, and has used several methods to model and predict the channel errors. The choice of the propagation model to be used can affect ECD estimations [6]. A brief overview of four popular models for low frequency groundwave propagation is given in [6], along with the conditions under which they can be applied. In addition, Eurofix and eLoran, two modern Loran-C enhancements, use DGPS corrections to remove the bias in Loran-C position reporting.

3.1.3 Noise

Noise of many kinds can corrupt a Loran-C receiver. The primary source is atmospheric noise produced by lightning, which presents a high availability hazards for Loran-C users. It is unpredictable and impulsive in nature; it can travel great distances with the earth as a waveguide. Front-end filtering is absolutely essential to maintain maximum dynamic range in the receiver analog to digital converter (ADC). In addition, automatic gain control and some limiting are necessary to eliminate the highest noise peaks.

One other form of noise is what is termed p-static (precipitation static) noise. This type of noise results from the buildup and discharge of static electricity about sharp edges on aircraft during flight. Noise from receiver amplifiers and other thermal noise sources are also present. They, however, will be neglected here when considering modern receivers. They will generally contain much better low noise amplifier designs, which support this assumption.

3.1.4 Skywave Effects

In addition to the desired ground path, the channel also provides propagation via ionospheric refractions at the D and E layers. These signals are called 'skywaves'.

Historically, they have been problematic for Loran-C in obtaining accurate navigational fixes, even greatly limiting proposed coverage areas beyond noise limits. The path of a skywave is generally much longer than its corresponding groundwave. Thus, skywaves arrive at the receiver after the beginning of the ground pulse, usually $35\mu\text{s}$ or more. The two paths are illustrated in Figure 3.1. This arrival time does tend to change greatly with transmitter distance and over time. Transmitter distance changes the relative path lengths of both groundwave and skywave signals. When close to the transmitter, the user will generally see the greatest delay, where the reflection angle is least. Also, changes in the ionization intensity, both seasonally and diurnally, alter the ionosphere's effective height. For daytime use, 73 km is average, where 91 km is that for nighttime use. This, too, affects the geometry and, hence, the arrival time for skywaves. In addition, the ionosphere reflection coefficient varies diurnally where greater field strengths occur at night than daytime. Seasonal variations in the reflection coefficient also occur, such that the intensity during winter is much greater than that during summer. One might assume then that winter nights would be a worst-case scenario for the Loran-C user. Previously this had been assumed. However, further measurements and analysis support the case that during winter months the skywave interference is actually most severe in daytime hours. This is because the skywave intensity is overshadowed by the greater sensitivity to skywave delay. So, greater skywave intensity may be afforded if they arrive much later. Another interesting note of skywave intensity is that it is relatively independent of distance [6]. Since groundwave intensity falls with distance, the SGR (skywave to groundwave ratio) grows with distance. Thus, as a user travels farther from a particular transmitter, he will observe earlier and greater skywave interference.

As stated, skywaves arrive and distort the Loran ground pulses. Unless checked, this puts a limit on how many cycles can be reliably tracked. As time difference wanes, the latest available zero crossing's SNR is decreased. This produces a reduction in overall coverage. As technology and research have evolved together in this field, many techniques for mitigating these effects have been proposed and implemented [7, 18]. Roth [17] claims that their receivers, more contemporary lab grade devices, can readily eliminate skywave contamination. The implications are

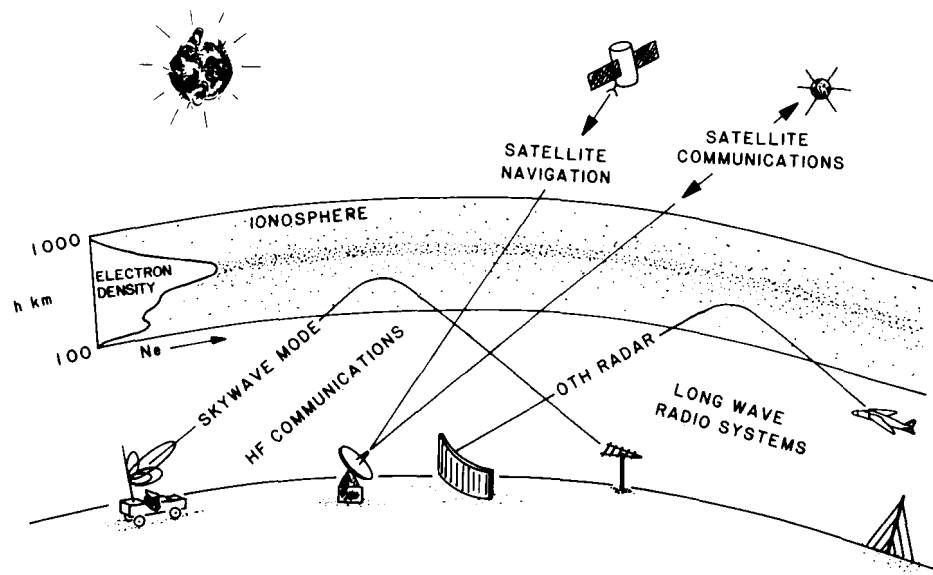


Figure 3.1 – Illustration of Skywave and Groundwave Paths – Source: [29].

that later zero crossings can be tracked, resulting in higher SNR for given transmission power. Thus, the available coverage is greatly increased.

The Loran-C pulse was designed to help combat the problem of skywaves, though it is limited in its ability. Navigation systems like DNS (Decca Navigation System) in Europe are continuous wave (CW) type and are highly susceptible to skywave contamination. DNS actually operates in the same band as Loran-C and its effective coverage is much less. Decca receivers cannot distinguish groundwaves from skywaves very well compared to Loran-C because of its signal shape. The coverage limits due to skywave interference in Loran-C are a function of both skywave intensity and relative delay. The former is the only parameter in determining limitations of Decca availability. Interestingly, DNS observes the worst degradation from skywaves during winter nights when the intensity is the greatest.

3.1.5 Cross-Rate Interference

All Loran-C chains operate simultaneously in the same band. Geographical separation is the only means of isolating signals. Cross-Rate Interference (CRI) occurs when adjacent and further chains signals overlap at the receiver with those signals within and cause distortion. Since all have identical carrier frequencies, the carrier phase can be greatly perturbed. This phenomenon can be quite problematic for the navigator because the reported time delays can be changed tremendously, reducing confidence in position solution and stability by an order of magnitude [16]. Fortunately, most modern receivers have the ability to greatly mitigate, or lock out, CRI to some extent [9]. Legacy receivers are still in use today which cannot combat nearly as well. To better understand the difference in performance between the two, we examine the fundamentals of CRI.

Let us first assume a stationary user. Over a small time interval, CRI is not random. The cross rate interference is periodic and is related to the proportion of the current chain's GRI and that of each interfering chain's GRI. The proportion determines the fundamental rate at which the interference will 'walk' through the desired signals. The carrier phase of each will remain constant since the Loran-C carrier and the GRIs are all integer related [16]. If one has the ability, the pulses being distorted can be omitted for ease of tracking user position. This requires the receiver to be able to

track or view multiple chains. If one cannot do so, great difficulty will result in attempts to predict CRI occurrences. Removing the assumption of a stationary user with a finite observation window reveals other interesting characteristics. First, spatial and time variations exist. Mobile users will generally see decorrelation effects and must have sufficient processing capabilities in order to maintain CRI lock out. Over time, propagation conditions may vary, resulting in changes even in stationary receivers. The time constant will depend upon the path taken to the user from the adjacent Loran-C chain transmitters [16]. One interesting approach the blanking of CRI is through an inverted comb filter for each GRI causing the distortion. In addition to the capability of blanking the affected pulses, some contemporary receivers can actively subtract the unwanted signals and isolate each component for its individual contribution to the navigation solution [13]. So, CRI has been introduced along with its performance implications and solutions.

3.2 CWI Fundamentals

From this point forward, the main focus of performance degradation will be carrier wave interference (CWI). This form of interference is present within, and adjacent to, the Loran-C spectrum. Much work has been done to eliminate their negative effects on navigation accuracy and availability [1, 2, 5, 4].

3.2.1 Interference Classification

Before entering into great detail on carrier wave interference (CWI) issues with Loran it is necessary to provide a classification of the types of interference presented to a Loran receiver. There are three main types of CWI. They are synchronous, asynchronous, and near-synchronous. There is also a fourth type called sub-synchronous which is only encountered in hard-limiting receivers. Since most contemporary designs are the linear type, this form is of less importance and is left to the reader if so desired. Below are two ways of defining carrier wave interference types as defined in [1].

$$\begin{aligned}
\text{Synchronous interference} \quad & \left| f_{\text{int}} - \frac{N}{2GRI} \right| = 0 \\
\text{Asynchronous interference} \quad & \left| f_{\text{int}} - \frac{N}{2GRI} \right| > f_b \\
\text{Near-synchronous interference} \quad & \left| f_{\text{int}} - \frac{N}{2GRI} \right| < f_b
\end{aligned} \tag{3.1}$$

where N is a non-negative integers, f_{int} is the interference frequency, f_b is the tracking bandwidth of the receiver. Another way of classifying CWI is to change f_{int} :

$$f_{\text{int}} = \frac{N + q}{2GRI} \tag{3.2}$$

where $-0.5 \leq q \leq 0.5$. Then, substituting into (3.1) gives:

$$\begin{aligned}
\text{Synchronous interference,} \quad & |q| = 0 \\
\text{Asynchronous interference,} \quad & |q| > f_b 2GRI \\
\text{Near-synchronous interference,} \quad & |q| < f_b 2GRI
\end{aligned} \tag{3.3}$$

One can see upon initial inspection that several factors determine the type of interference in question and thus the type of harmful effects on system performance it can cause. First, note that one interferer might have a different classification from one GRI to the next. It is the fundamental period of the Loran chain that determines where the spectral lines, the points of synchronicity, lie. Any interferer lying on one of these spectral lines is deemed synchronous, which is, as will be shown, the most troublesome. It can be viewed as aliasing the energy to zero frequency as a stationary bias. The near-synchronous frequencies lie about the lines and within the regions determined by the actual tracking bandwidth of the receiver being utilized.

This is shown in (3.3). So, as one can expect, this can vary from machine to machine. All other CWI is deemed asynchronous. This type generally results in a noise like effect [1].

3.2.2 System Model and Simplifying Assumptions

This section provides a simplified model of the mechanism of phase tracking and cycle selection errors in a Loran-C receiver in the presence of carrier wave interference. Van Willigen [1] and Last [2] are the primary resources here. The reader is directed there for further details as the author here provides only a brief overview. Several assumptions must be made prior to discussion and analysis. First, a single interferer is assumed present and as an unmodulated carrier. Second, it is assumed phase and amplitude stable over the channel during the observation time. Lastly, any filtering effects are withheld until a later section where analysis is appropriate in order to simplify discussion here.

3.2.3 Phase Tracking

Discussion and analysis of phase tracking must begin with a model of the sampling system and its phasor representation. Van Willigen [1] defines two timing frames to be observed. First is the small frame. One sample is taken from the tracking point of each Loran-C pulse over a 2GRI period. Thus, 16 total samples make up the small frame. The second frame is the large frame which is over multiple 2GRI intervals. So, the resultant of a small frame is a phasor in the large frame. Two important periods of interest now appear. The first period is the 1ms Loran-C pulse spacing within a pulse train. The second period is twice the GRI interval. The reasoning behind the 2GRI period is that the Loran-C phase code repeats every two GRI and will later simplify analysis of phase coding for interference reduction. So, small frame analysis will focus on phase coding and the attenuation of CWI due to this operation. Large frame analysis simply focuses on changes over larger periods, namely integer multiples of 2GRI. Figure 3.2 depicts a small frame and the phase codes over that interval used by master and slave stations in the system.

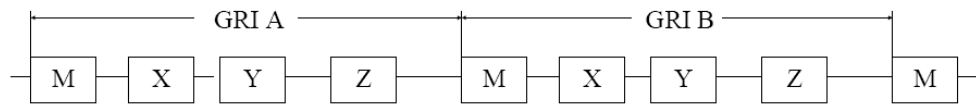


Figure 2. LORAN-C Group Repetition Interval

Master										Slave									
A					B					A					B				
+	+	-	-	+	-	+	-	+	-	+	-	-	+	+	+	+	+	-	-

Figure 3.2 -- Illustration of Loran-C Chain Structure and Phase Coding – Source: [1].

Now that sampling has been defined let us discuss the phasors themselves. First, all resultant signals in the phasor diagram are assumed multiplied with the receiver's phase decoder. This means the Loran component is decoded and the remainder is encoded. The angle between the Loran component and the resultant phasor is the error in phase reported by the receiver's carrier tracking loop.

3.2.4 TOA Errors

All phase tracking errors will translate into timing errors. Last [2] defines the contribution of timing error from a single CWI as:

$$\text{TOA error} = \frac{10}{2\pi} \sin^{-1} \left[\frac{I}{C} \sin(\phi_{\text{int}}) \right] \quad (3.4)$$

where ϕ_{int} is the relative phase of the CWI to the Loran-C carrier, I is the interferer's amplitude, $C = \sqrt{I^2 + S^2 + 2IS\cos(\phi_{\text{int}})}$, and S is the amplitude of the Loran-C envelope at the given zero crossing. For large SIR, the maximum error from (3.4) becomes a function of SIR. This maximum occurs when $\phi_{\text{int}} = \frac{\pi}{2}$.

3.2.5 Phase Decoding and Averaging

Last [2] rigorously derives a method to obtain the desired quantitative representation of the attenuation of CWI by phase decoding and averaging. A mathematical representation of the Loran signal during the observation window, $t \in [0, N \cdot 2GRI]$, is given:

$$x_L(t) = \sum_{n=1}^N \left[\sum_{m=1}^{16} x_0(t) * \delta(t - t_m) pc(m) \right] * \delta(t - t_n) \quad (3.5)$$

where $x_0(t)$ is a standard Loran-C pulse (2.1), $pc(m)$ is the phase coding function shown in Figure 3.2, and δ a CT impulse whose sifting property is employed for pulse and pulse group repetition, respectively. The interference and composite signals are defined:

$$x_{\text{int}}(t) = I \sin(2\pi f_{\text{int}} t + \phi_{\text{int}}) \quad (3.6)$$

$$x_c(t) = x_L(t) + x_{\text{int}}(t) \quad (3.7)$$

where f_{int} is the carrier frequency of the interferer and ϕ_{int} represents the phase of the CWI relative to the Loran-C carrier. The variable I in (3.6) represents the amplitude of $x_{\text{int}}(t)$. So, (3.7) is an ideal representation of the signal arriving at the receiving antenna. The next step in the process is to understand how phase decoding and averaging change this signal, given as $y(t)$ in (3.7):

$$y(t) = \frac{1}{16N} \sum_{n=1}^N \left[\sum_{m=1}^{16} x_c(t + t_m + t_n) pc(m) \right] = y_1(t) + y_2(t) \quad (3.8)$$

$$y_1(t) = \frac{1}{16N} \sum_{n=1}^N \left\{ \sum_{m=1}^{16} x_L(t) * \delta(t + t_m + t_n) pc(m) \right\} = x_0(t) \quad (3.9)$$

$$y_2(t) = \frac{1}{16N} \sum_{n=1}^N \left\{ \sum_{m=1}^{16} x_{\text{int}}(t) * \delta(t + t_m + t_n) pc(m) \right\} \quad (3.10)$$

From (3.7) and the principle of linearity, analysis of these effects on CWI attenuation can be performed. By splitting the received signal, $y(t)$, into its two respective components, $y_1(t)$ & $y_2(t)$ in (3.9) and (3.10), one can isolate each signal and process the effects separately. In the absence of any noise or other outside influences, (3.9) shows that the received Loran component will decode and return to

its original defined pulse shape, (2.1). Also, averaging does not change the Loran signal once it has been phase decoded. The CWI component in (3.10), $y_2(t)$, describes the interfering portion of the received signal after averaging and phase coding. The use of the Fourier transform, and its convolution property, on $y_2(t)$ shows quantitatively in (3.11) the amount of rejection of interfering energy.

$$Y_2(f) = F[y_2(t)]$$

$$Y_2(f) = X_{\text{int}}(f) \left\{ \left[\frac{1}{N} \sum_{n=1}^N \exp(+j2\pi f t_n) \right] \times \left[\frac{1}{16} \sum_{m=1}^{16} \exp(+j2\pi f t_m) pc(m) \right] \right\} \quad (3.11)$$

$$rejection(f_{\text{int}}) = \left[\frac{1}{N} \sum_{n=1}^N \exp(+j2\pi f t_n) \right] \times \left[\frac{1}{16} \sum_{m=1}^{16} \exp(+j2\pi f t_m) pc(m) \right] \quad (3.12)$$

where $X_{\text{int}}(f) = F[x_{\text{int}}(t)]$. Now one can see that the phase coded interference component, $X_{\text{int}}(f)$, is scaled by a frequency dependent constant, (3.12). Last [2] terms this the ‘rejection’ and it is a function of the interfering frequency. This yields a measure of the amount of attenuation to $x_{\text{int}}(t)$ by the phase decoding and receiver averaging operations. Continuing further, one can take each component of (3.12) and analyze independently by linearity. The contributions from averaging and phase coding are given as $R_1(f_{\text{int}})$ and $R_2(f_{\text{int}})$ in (3.13) and (3.14) respectively:

$$R_1(f_{\text{int}}) = \left| \frac{1}{N} \sum_{n=1}^N \exp(+j2\pi f t_n) \right| \quad (3.13)$$

$$R_2(f_{\text{int}}) = \left| \frac{1}{16} \sum_{m=1}^{16} \exp(+j2\pi f t_m) pc(m) \right| \quad (3.14)$$

$$R_1(f_{\text{int}}) = \left| \frac{1}{N} \sum_{n=1}^N \exp[+j2\pi q_{\text{int}}(n-1)] \right| \quad (3.15)$$

$$R_2(f_{\text{int}}) = \left| \frac{1}{16} \sum_{m=1}^{16} \exp(+j2\pi \frac{N_{\text{int}} + q_{\text{int}}}{2GRI} t_m) p_c(m) \right| \quad (3.16)$$

After substitution, (3.15) & (3.16) are resultant. These last equations are instrumental in the analysis section where values for N and q will be substituted for the different interference classifications and results will be given as from Last [2].

3.2.6 Cycle Selection

A short discourse by Last [2] on the cycle selection process is given.

Cycle selection is the process by which the receiver identifies the cycle on which the zero-crossing measurement is to be performed. Most receivers use the principle that the ratio of the amplitudes of the two half-cycle peaks straddling each zero crossing is unique. By measuring these 'half-cycle peak ratios' (HCPR) the correct zero crossing may be identified.

The error between the estimated and true sampling points is termed the 'measured ECD error', using ECD to mean the time discrepancy between the carrier of the Loran-C pulse and its envelope. Specifically, ECD may be thought of as the discrepancy between the zero crossing used for phase tracking and the corresponding point on the envelope. If the magnitude of this measured ECD error exceeds 5 μs , the wrong cycle will be chosen.

It should be noted again that CWI is not the only contributor to ECD errors. The channel itself will distort the Loran-C pulse. A detailed treatise on these problems is given by Williams [99]. A simplified analysis of the cycle selection process is given in Last [2]. A more complete statistical measure of the cycle selection block performance can be found in Last [4]. The half cycle peak ratio (HCPR) is defined:

$$\text{HCPR} = \frac{A_1 + \Delta_1}{A_2 + \Delta_2} \quad (3.17)$$

$$\Delta_1 = \Delta_0 \sin\left(+2\pi f_{\text{int}} 2.5 \times 10^{-6} + \phi_{\text{int}}\right) \quad (3.18)$$

$$\Delta_2 = \Delta_0 \sin\left(-2\pi f_{\text{int}} 2.5 \times 10^{-6} + \phi_{\text{int}}\right) \quad (3.19)$$

where A_1 and A_2 are the ideal peak values about the zero crossing being tracked for the Loran component. Δ_0 is taken as the peak interfering signal's amplitude. Δ_1 and Δ_2 are the CWI components taken at the times of Loran-C carrier peaks about the tracking point. Under the conditions that $\Delta_0 \ll A_1$ and $\Delta_0 \ll -A_2$ some simplifying assumptions may be made which will make expression of the HCPR error clearer. All terms are referred from the illustration in Figure 3.3.

$$\text{HCPR} = \frac{A_1}{A_2} \left[1 + \frac{\Delta_1}{A_1} - \frac{\Delta_2}{A_2} \right] \quad (3.20)$$

$$\text{HCPR}_e = \frac{A_1}{A_2} \left[\frac{\Delta_1}{A_1} - \frac{\Delta_2}{A_2} \right] \quad (3.21)$$

where (3.20) is (3.17) redefined under above assumptions and (3.21) is the error in the ratio. After further substitution and rearrangement, the maximum absolute error in half cycle peak ratio is defined:

$$\begin{aligned} |\text{HCPR}_e|_{\text{max}} = & \left[R_1(f) R_2(f) \Delta_0 \frac{A_1}{A_2} \times \right. \\ & \left. \left[\left(\frac{1}{A_1} + \frac{1}{A_2} \right)^2 \times \sin^2(5 \times 10^{-6} \times \pi f_{\text{int}}) + \left(\frac{1}{A_1} - \frac{1}{A_2} \right)^2 \times \cos^2(5 \times 10^{-6} \times \pi f_{\text{int}}) \right]^{-1/2} \right] \quad (3.22) \end{aligned}$$

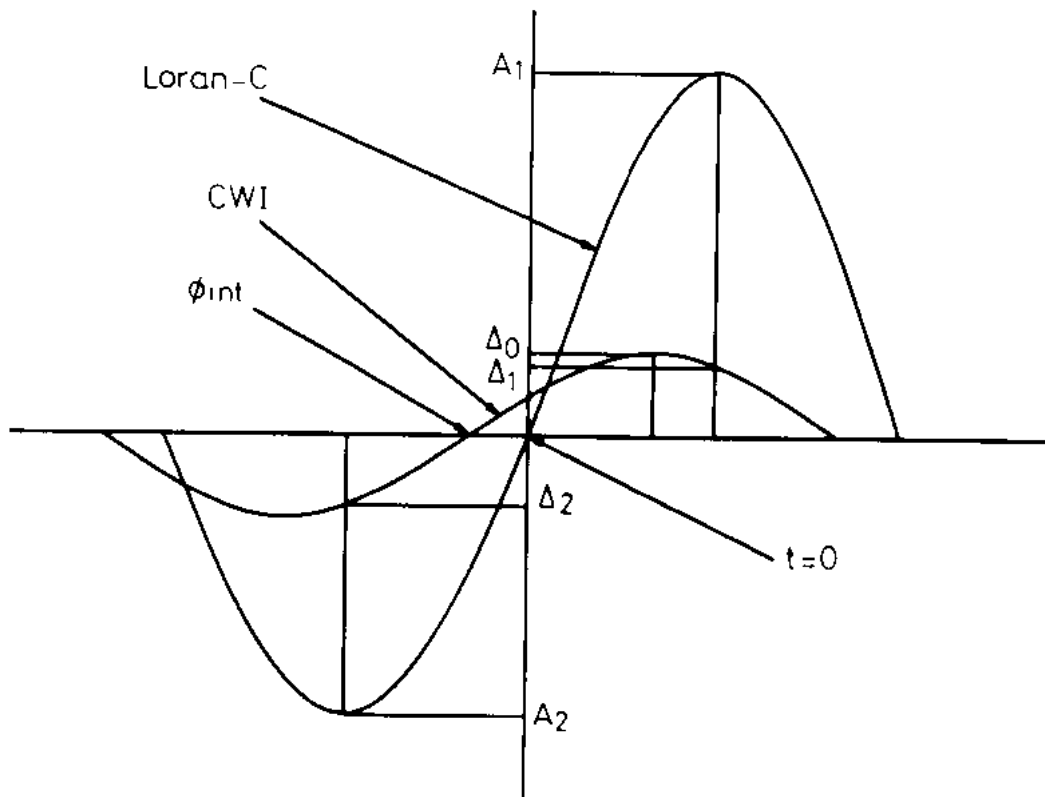


Figure 3.3 – Illustration of Half Cycle Peak Ratio Process – Source [4].

$$\tan(\phi_{\text{int}}) = \frac{\left(\frac{1}{A_1} - \frac{1}{A_2}\right)}{\left(\frac{1}{A_1} + \frac{1}{A_2}\right)} \cot(5 \times 10^{-6} \times \pi f_{\text{int}}) \quad (3.23)$$

where $R_1(f)$ and $R_2(f)$ are taken from (3.13) and (3.14) and maximum error occurs when (3.23) is satisfied. Thus, a frequency dependency also occurs. f_{int} controls R_1 , R_2 , as well as the remaining term in a complex manner [3]. Beckmann [3] shows that frequencies closer to the 100 kHz Loran-C carrier will tend to cause fewer errors.

3.2.7 Performance Analysis

We begin with analysis of synchronous CWI. As shown in (3.3), synchronous CWI has $q=0$. The carrier frequency is an integer multiple of $1/\text{PCI}$. During each phase code interval of 2GRI , the CWI's phase will then rotate $2N\pi$ radians, where N is a non-negative integer [1]. Thus, every large frame phasor will be identical, creating a bias in the reported carrier phase by the carrier tracking loop.

The size of the phase error which thus occurs arises from relative signal amplitudes, phase coding, interference frequency, GRI, receiver implementation and the phase difference between interference and the Loran-C signal. [1]

The bias in phase translates to a timing offset and a resultant position error. The ionospheric channel will also provide this interference to the receiver as well and the effects on the bias will differ depending upon atmospheric conditions [3]. Beckmann [3] proposed synchronous interference to be the most harmful type to Loran-C performance. Last [2] suggested that such performance degradation would tend to decrease the effective coverage of a Loran-C chain. The receiver cannot distinguish this type of CWI from a movement in actual user position. The signal envelope will also be distorted. The resultant bias in ECD can cause unacceptable cycle slips at even high SIR values. $R_2(f)$, (3.16), is a function of the GRI and f_{int} . For

synchronous frequencies the rejection varies from 10 to 18dB. Synchronous CWI is not affected by averaging. Substituting $q=0$ into $R_1(f)$, (3.15), verifies this. This illustrates quantitatively how synchronous CWI can be problematic for the phase tracking process. The attenuation of synchronous CWI in the cycle selection process is identical to that in the phase tracking process. The envelope distortion will also be constant after continued averaging. Large ECD errors, and consequently, greater likelihood for cycle slips will result for synchronous CWI.

The effect for near-synchronous interference is very similar. Instead of a constant bias, a slowly varying TOA error from phase tracking will result over the observation interval. To be more exact, it will rotate with a frequency of $q\pi/GRI$ rad/sec, where q is taken from (3.3). The reported ECD from the cycle selection block will vary similarly for near-synchronous CWI. The factors determining the range of near-synchronous frequencies about each spectral line, again, depends upon several factors. For mobile applications, the averaging period will generally be less than for stationary modes. Also, the cycle selection block usually has narrower bandwidth than phase tracking. Longer averaging periods will generally allow this. There is a limit on the averaging period due to the time constant of the channel. It is not truly stationary and further averaging will degrade performance. If one operates within this window, sufficient averaging may exist to minimize the range of near-synchronous frequencies.

Asynchronous interference is the least harmful type one can encounter. It can be attenuated significantly by phase decoding and averaging, (3.15) & (3.16). Lower potential availability hazards exist for asynchronous CWI. Envelope distortion can be averaged out and ECD errors can be minimized, which translate to fewer potential cycle slips.

3.3 Statistical Analysis of CWI

The previous section's purpose was to give understanding to the mechanisms of sampling and classification of CWI and also introduce basic analyses of receiver operations and how CWI was affected. Many simplifications were made in order to understand these processes. As a result, one is limited in the applicability of such

results. Much of the early work on CWI focused on such models and analysis was often limited to a single interferer before becoming tedious. In reality, CWI are numerous and are often modulated, which complicates analysis. To better understand the effects of the presence of many CWI, Last [4] explored the interference issue from a statistical perspective. This approach offers a more general view of CWI and its effects on Loran-C accuracy and availability.

Before deriving the necessary equations, some assumptions and clarifications need to be made. Since this model deals with the study of multiple interferers, the statistics can be quite complicated [4]:

The study of the statistics of the errors of timing measurements is complicated by a number of factors: 1) the number of interferers is not easy to define, 2) parameters such as the GRI, SIR, and ECD of the received Loran-C signals may occur in a range of combinations, and 3) complex schemes and various algorithms for phase measurement and cycle selection are employed in different receivers.

3.3.1 System Model

One must understand specifically what the statistics represent. Last [4] stated that statistics of time should be used for stationary receivers and both space and time joint statistics for mobile receivers. The main source of randomness here will be that of the relative carrier phase of each interferer to that of the Loran-C carrier phase, namely, ϕ_{int} is taken as $U [0, 2\pi]$. So, ϕ_{int} is now defined as a random variable and T_e is a function of it. Last [4] derives the probability distribution function. Its result is given:

$$F_{T_e}(t_e) = P\{T_e \leq t_e\} = \frac{1}{2\pi} \left\{ \pi + 2 \arcsin \left(\frac{C}{I} \sin \frac{2\pi}{10} t_e \right) \right\} \quad (3.24)$$

where the distribution function holds when $t_e \in \left(\frac{-10}{2\pi} \arcsin\left(\frac{I}{C}\right), \frac{10}{2\pi} \arcsin\left(\frac{I}{C}\right) \right)$ is satisfied. Differentiating (3.24) yields the pdf here:

$$f_{T_e}(t_e) = \begin{cases} \frac{2}{10} \frac{C}{I} \frac{\cos\left(\frac{2\pi}{10} t_e\right)}{\sqrt{1 - \left(\frac{C}{I} \sin\left(\frac{2\pi}{10} t_e\right)\right)^2}}, & t_e \in \left(\frac{-10}{2\pi} \arcsin\left(\frac{I}{C}\right), \frac{10}{2\pi} \arcsin\left(\frac{I}{C}\right) \right) \\ 0, & \text{elsewhere} \end{cases} \quad (3.25)$$

Figures 3.4 and 3.5 illustrate (3.24) and (3.25), respectively. Consider for a moment the fact that the wavelength at 100 kHz is 3km and the effective radius of a single station is many hundred kilometers. Given the total coverage area of one chain of several stations, the assumption on the distribution of ϕ_{int} seems quite reasonable for both ground and skywaves. The latter will also see marked variations over time [4]. This provides a clearer view of the errors, not just the maximum possible in magnitude. In addition to the distribution, $f_{T_e}(t_e)$, another measure of performance can prove useful. The Minimum Performance Standards (MPS) [X] define the maximum allowable timing error to be 100ns. Evaluating (3.24) at $\pm T_e$ yields a new random variable representing the confidence of $|T_e| \leq t_e$ [4]:

$$P\{|T_e| \leq t_e\} = \frac{2}{\pi} \arcsin\left(\frac{C}{I} \sin\left(\frac{2\pi}{10} t_e\right)\right) \quad (3.26)$$

This can give the designer a more useful tool in determining confidence bounds for performance. It is illustrated in Figure 3.6. Also, Figure 3.7 illustrates (3.4).

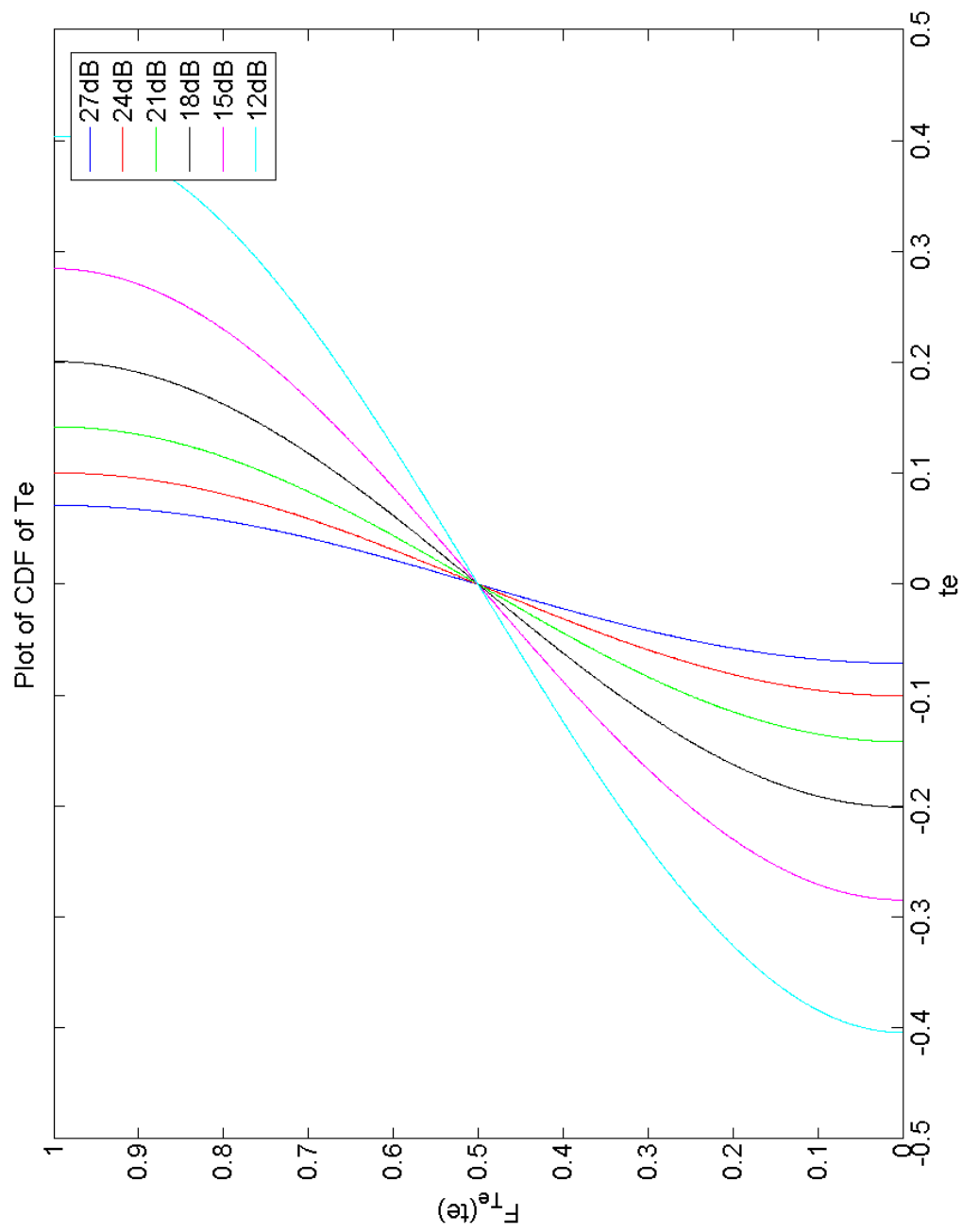


Figure 3.4 – Cumulative Distribution Function for T_e , $F_{T_e}(t_e)$

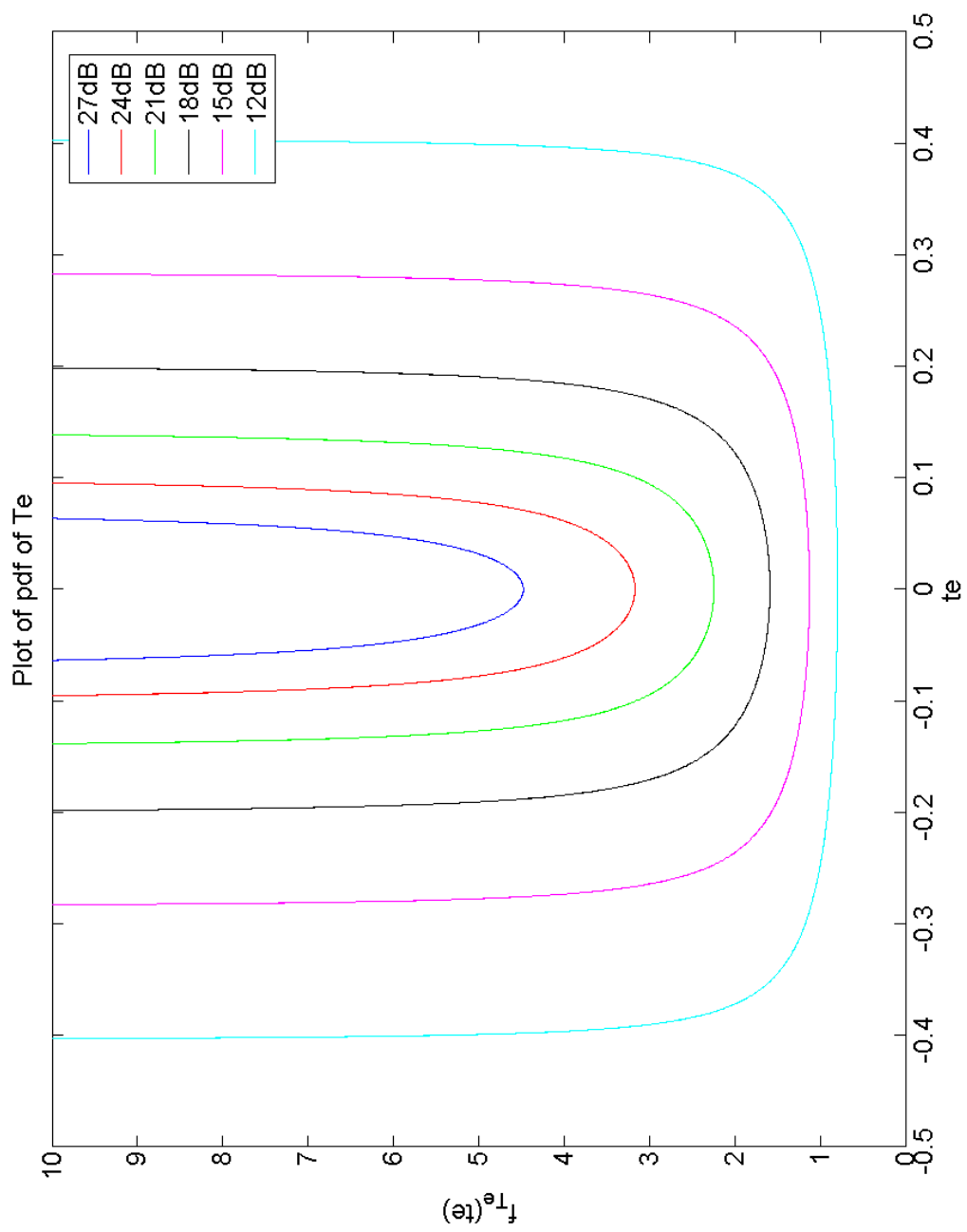


Figure 3.5 – Probability Density Function of T_e , $f_{T_e}(t_e)$.

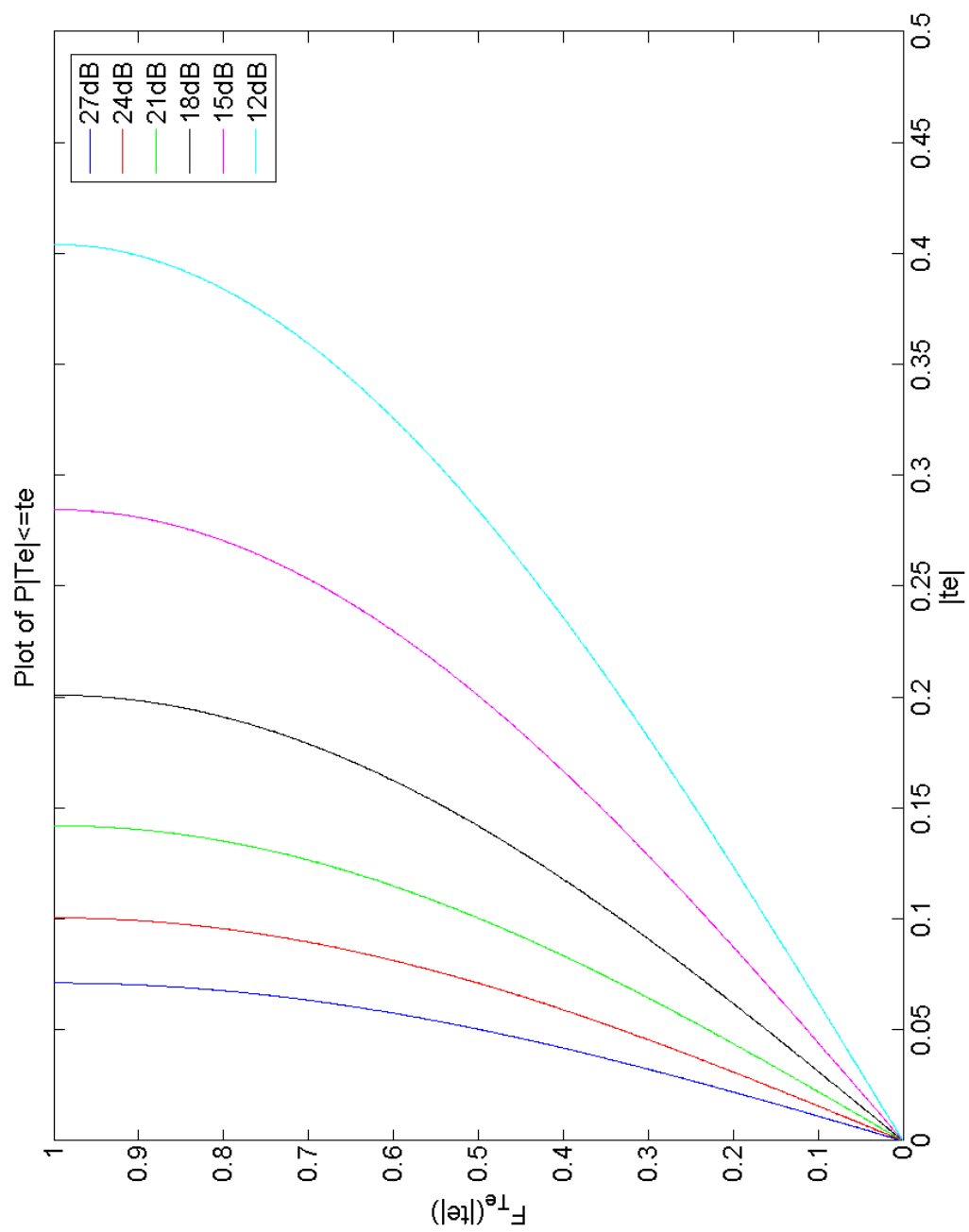


Figure 3.6 – Probability Density Function of $|T_e|$.

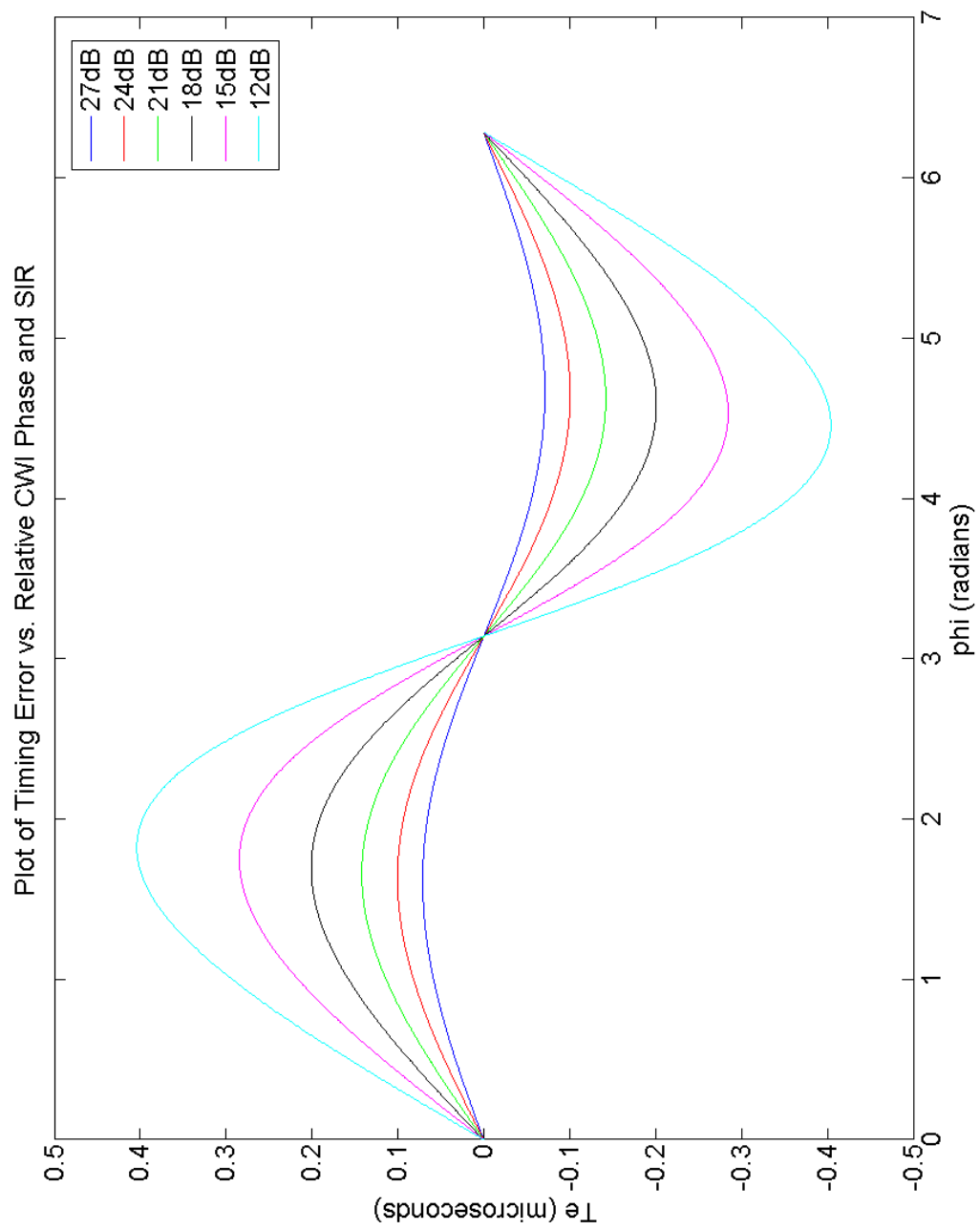


Figure 3.7 – Phase tracking error versus relative carrier phase.

3.3.2 Multiple Interferers

Extending the analysis to the case of many interferers requires a method of determining the interaction and resultant distribution of the timing error [4]. Last [2, 4] make the assumption that at high SIR, 24dB or better, the small angle approximation, $\sin(\phi_{\text{int}}) \approx \phi_{\text{int}}$, holds. (3.4) simplifies and the total timing error is then the sum of the individual errors:

$$T_e = \sum_{k=1}^K T_{ke} = \sum_{k=1}^K \frac{10}{2\pi} \frac{I_k}{S} \sin \phi_k \quad (3.27)$$

where I_k and ϕ_k are the respective amplitude and phase of each of the K interfering signals. This assumption is quite helpful; the individual errors can be summed directly to obtain the resultant. A new random variable, T_e , is defined as the sum of the K random variables of the individual contributors, T_{ke} . By definition, the resultant distribution is obtained by convolving the individual pdfs. To avoid this tedious computation for large K , [4] employed the use of characteristic functions. The statistics can be obtained by multiplication, instead of the many convolution operations otherwise required. For the case where 10 or more interferers are present, and are i.i.d., or lack dominant components, the Central Limit Theorem can be evoked and the probability density function can be approximated as Gaussian with truncated tails [4].

$$E(T_e) = E\left(\sum_{k=1}^K T_{ke}\right) = \sum_{k=1}^K E(T_{ke}) = 0 \quad (3.28)$$

$$\text{var}(T_e) = \sum_{k=1}^K \text{var}(T_{ke}) = \sum_{k=1}^K \left(\frac{25}{2\pi^2} \left(\frac{I_k}{S} \right)^2 \right) \quad (3.29)$$

With respect to tracking the correct crossing, a similar distribution is defined for HCPR:

$$R_e = \Delta_0 \frac{A_1}{A_2} \left[\left(\frac{1}{A_1} + \frac{1}{A_2} \right) \sin(5\pi f_{\text{int}}) \cos(\phi_{\text{int}}) + \left(\frac{1}{A_1} - \frac{1}{A_2} \right) \cos(5\pi f_{\text{int}}) \sin(\phi_{\text{int}}) \right] \quad (3.30)$$

where all parameters are as defined in (3.17-3.19). Thus, R_e is a function of ϕ_{int} , f_{int} , SIR, and the gradient of the particular zero crossing being tracked. After much derivation, Last [4], the respective cdf and pdf for the HCPR error under a single CWI are given:

$$F_{R_e}(r_e) = P\{R_e \leq r_e\}$$

$$F_{R_e}(r_e) = P\left\{ \Delta_0 \frac{A_1}{A_2} A \cdot \sin(\varphi + \phi_{\text{int}}) \leq r_e \right\} \quad (3.31)$$

$$F_{R_e}(r_e) = \frac{1}{2\pi} \left[\pi + 2 \arcsin\left(\frac{A_2}{\Delta_0 A_1 A} r_e \right) \right] \quad (3.32)$$

$$f_{R_e}(r_e) = \begin{cases} \frac{2A_2}{\pi \Delta_0 A_1 A \sqrt{1 - \left(\frac{A_2}{\Delta_0 A_1 A} r_e \right)^2}}, & \text{for } \frac{-\Delta_0 A_1 A}{A_2} \leq r_e \leq \frac{\Delta_0 A_1 A}{A_2} \\ 0, & \text{elsewhere} \end{cases} \quad (3.33)$$

where $\varphi = \tan^{-1} \left\{ \frac{\frac{1}{A_1} + \frac{1}{A_2}}{\frac{1}{A_1} - \frac{1}{A_2}} \tan(5\pi f_{\text{int}}) \right\}$ and (3.31) holds when

$$r_e \in \left(\frac{-\Delta_0 A_1 A}{A_2}, \frac{\Delta_0 A_1 A}{A_2} \right).$$

The HCPR error in (3.21) assumes ϕ_{int} is constant. Now, it is represented as a random variable $U \sim (0, 2\pi)$. In the case of multiple interferers, the cycle selection

error, as for phase tracking error, is taken as the sum of the individual error contributions:

$$R_e = \sum_{k=1}^K R_{ke} \quad (3.34)$$

$$R_e = \sum_{k=1}^K \Delta_{0k} \frac{A_1}{A_2} \left[\left(\frac{1}{A_1} + \frac{1}{A_2} \right) \sin(5\pi f_k) \cos(\phi_k) + \left(\frac{1}{A_1} - \frac{1}{A_2} \right) \cos(5\pi f_k) \sin(\phi_k) \right] \quad (3.35)$$

where f_k , Δ_{0k} and ϕ_k are the carrier frequency, amplitude, and relative phase (to Loran-C) for each of the K interferers. The same process is defined by Last [4] for simplifying analysis of error distributions by using CFs to eliminate the need to evaluate convolution integrals.

$$f_{R_{ke}}(r_e) = F \left\{ \frac{1}{2\pi} \Phi_{R_e}(\omega) \right\} \quad (3.36)$$

where (3.36) represent the probability density function of the total error. $\Phi_{R_e}(\omega)$ is the product of the individual characteristic functions given here:

$$\begin{aligned} \Phi_{R_{ke}}(\omega) &= \int_{-\infty}^{\infty} f_{R_{ke}}(r_e) e^{j\omega r_e} dr_e = F^{-1} \{ 2\pi f_{R_{ke}}(r_e) \} \\ \Phi_{R_e}(\omega) &= \Phi_{R_{1e}}(\omega) \cdot \Phi_{R_{2e}}(\omega) \cdot \dots \cdot \Phi_{R_{Ke}}(\omega) \end{aligned} \quad (3.37)$$

Invoking CLT, as for phase tracking, the joint distribution can be approximated as Gaussian with truncated tails. These extrema are set by the maximum theoretical being the sum of the individual component maxima. The mean and variance nearly completely describe the distribution for finite CWI:

$$E(R_e) = E\left(\sum_{k=1}^K R_{ke}\right) = \sum_{k=1}^K E(R_{ke}) = 0 \quad (3.38)$$

$$\sigma_{R_e}^2 = \text{var}(R_e) = \sum_{k=1}^K \text{var}(R_{ke}) = \sum_{k=1}^K \sigma_{R_{ke}}^2 = \sum_{k=1}^K \frac{1}{2} \left(\Delta_{0k} \frac{A_1}{A_2} A \right)^2 \quad (3.39)$$

A multiplicative constant to scale the curve such that the truncated area is unity is required. This will be more obvious for higher variances. This last note is true for both phase tracking and cycle selection error statistics.

3.3.4 Analysis

Now that these effects have all been presented, the next step will be to understand how receiver operations affect the statistics. As (3.13-3.16) and [2] show, CWI attenuation is heavily frequency dependent [4]. It has been determined that phase decoding and integration for a single CWI will not distort the Loran-C component. The same holds for multiple CWI [4]. Since we are operating under the assumption of unmodulated carriers as interference, the filtering processes need only be evaluated at each carrier frequency for amplitude and phase response. The latter will be absorbed into the phase uncertainty such that the amplitude need only be scaled for each interfering signal as for the case of a single CWI [4]. Below, the interfering signal is given as the sum of the individual components [4]:

$$x_{\text{int}}(t) = \sum_{k=1}^K x_k(t) = \sum_{k=1}^K I_k \sin(2\pi f_k t + \phi_k) \quad (3.40)$$

which was derived from (3.6) for a single CWI. After further substitution, the phase coded and averaged interference component is given here:

$$y_2(t) = \sum_{k=1}^K \left\{ \frac{1}{16N} \sum_{n=1}^N \left[\sum_{m=1}^{16} x_k(t) * \delta(t + t_m + t_n) \text{pc}(m) \right] \right\} \quad (3.41)$$

$$y_2(t) = \sum_{k=1}^K y_{2k}(t) \quad (3.42)$$

All analysis has yet to add attenuation factors of phase decoding, averaging, and filtering. The same method as presented in [1] can be applied to each CWI in terms of amplitude attenuation. The same holds for filtering operations. The latter will generally not hold for the attenuation of wideband CWI. Simulation will be required to produce the desirable distributions.

3.4 Conclusion

For a more general understanding, the statistical approach taken in 3.2 has proven quite helpful indeed. It has given the authors a platform on which to begin comparing these results with those of wideband overlay signals. Specifically, the cases of 2 CWI and 10 CWI are of great importance in comparing the results. Also, in chapter 5 these statistics have all been verified by simulation for unmodulated CWI prior to proceeding to the WB analysis. Thus, a greater confidence in the results of [1, 2, 4] is held by the authors.

Chapter 4 Lab and Simulation Work

4.1 Wideband Study

The overlay signals were designed as CW, not pulsed like Loran. The modulation method chosen by the designers was a modified version of the Feher-patented [24] QPSK, called Enhanced Feher QPSK (EFQPSK), the former of which is popular in modem designs for the same attractive qualities the latter was chosen in our work. Long range navigation from terrestrial sources requires tremendous power to overcome the required distances. Because of this, a very efficient amplifier is required to conserve power. To maximize this efficiency, the EFQPSK modulation incorporates a combination of controlled I & Q branch correlation and waveform mapping. The operations work together to ensure that no abrupt changes or slope discontinuities occur during symbol transitions. All this helps to tremendously reduce sidelobe energy, which is beneficial when riding extremely close to the band edges of Loran-C. One might then ask why not reduce sidelobe energy with a simpler OQPSK modulation, for example. It is much less complex and, with proper pulse shaping, can achieve comparable attenuation off the main spectral lobe. We then must return to the question of its envelope properties. Rectangular QPSK is constant in envelope, but the case changes when pulse shaping is required. If a linear amplifier is not used here, spectral regrowth from the nonlinear amplifier will negate any gains made in the pulse shaping. The EFQPSK method retains both capabilities, while sacrificing very slight performance losses and increased complexity. To further identify each signal, a pair of orthogonal (sufficiently) Gold codes, length 1023, was allotted to each signal. The orthogonality requirement is stringent between I and Q because the slight correlation introduced by the mapping process would cause the average spectrum to distort and become asymmetric. This is, obviously, undesirable. A full account of the design and structure of EFQPSK modulation can be found in Simon [24].

The reported laboratory results in 4.2 list several signals not mentioned elsewhere but here. Following the conclusion of these tests, it was decided to focus the remainder of the work on the two signals in adjacent bands to Loran, termed A and

B. They seemed to be the least harmful of the set. Much was attributed to the greater attenuation provided by the bandpass filter. It is important to note that, although adopted by the Northern European Loran System (NELS) as the standard analog BPF, better solutions exist. For instance, using an analog filter for anti-aliasing purposes and focusing noise and interference attenuation to the digital realm can be quite advantageous. First, more consistency will be noted. The wider bandwidth of the analog filter will mean less sensitivity to temperature drift for signals of interest. The use of linear phase digital filters further ensures the preservation of the critical Loran envelope. For now, though, we continue with the analog BPF assumption.

4.2 Laboratory Work

Laboratory tests have been conducted to uncover the susceptibility of Loran-C receivers to wideband CW interference. Our goal was primarily to show that existing receiver designs, including COTS (Commercial Off-the-Shelf) and laboratory grade units, would be negligibly affected by the overlay signals under certain conditions. If these overlays are implemented in the future, the continued use of legacy receivers will be necessary until newer designs can be introduced, which will actively separate Loran-C from the overlays. The experiments were not very elaborate in nature. They were designed as simple proof-of-principle type tests to ensure no obvious problems existed prior to the continuation of the overlay design and implementation.

4.2.1 Signal structure

Five distinct interfering signals were produced with two modulation methods. For all signals, each direct and quadrature branch was spread by a sufficiently orthogonal Gold code with length 1023. The first of the five signals was centered at 100 kHz with a chipping rate of 20 ksps. It was denoted as 'Fullband' because its main lobe occupied the entire 80-120 kHz passband for Loran-C. The next four each had chipping rates of 5 ksps and were centered at 85, 95, 105, and 115 kHz. They were termed A, C, D, and B respectively. First, standard rectangular OQPSK was used for modulation. The next modulation scheme employed was Enhanced Feher-Shaped

QPSK (EFQPSK). It introduced a nearly constant envelope and significantly reduced side lobe presence. The advantages of this type of QPSK modulation were that it allowed for a much more efficient power amplifier due to the envelope properties and the side lobe reduction allowed for closer proximity of all signals with much less adjacent channel interference. The chosen GRI for the Loran-C signal was 9940. Both A & B carriers are synchronous with this GRI. This GRI was chosen because it provided a worst-case scenario for the given overlay design.

4.2.2 Hardware description and test setup

As stated, two separate Loran-C receivers were used in testing. The laboratory grade unit was a SATMATE by Locus, inc. and the COTS receiver was a Furuno. The FPGA boards used were two Nallatech XtremeDSP Development Kits. The models were created in the Mathworks SIMULINK environment with the Xilinx SysGen library. This tool allows designers to build actual hardware models and port them to programmable bit files for the desired FPGA.

For most of the testing, each board would generate a separate set of signals, one for Loran-C and one for the CWI. Each board contained two ADCs and two DACs. When more than four total signals were required, a separate COTS Loran simulator was utilized to generate the Loran-C pulses. This was to ensure maximum dynamic range from the board DACs, minimizing quantization distortion. Although the DAC units were 16 bit, additional steps were taken to get as close to ideal as possible. The full dynamic range of each DAC, minus 1-2 dB to buffer against overflow, was utilized. When the COTS Loran simulator was used, additional filtering was required on its output. The digital circuitry used in its design was causing significant high frequency noise. A lowpass filter was used with a cutoff frequency of 1 MHz to ensure minimal effect on the phase of the Loran-C signal. An illustration of the basic testing setup is given in Figure 4.1, and a closeup diagram of the Loran transmitter is given in Figure 4.2.

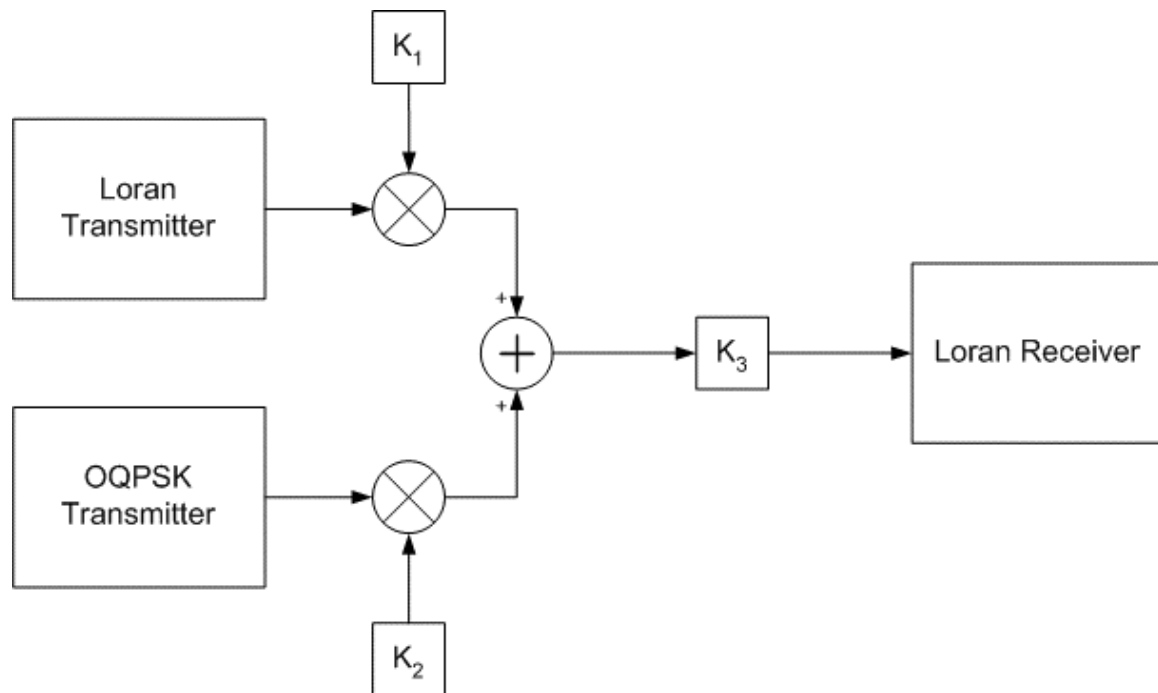


Figure 4.1 – Illustration of Basic Laboratory setup

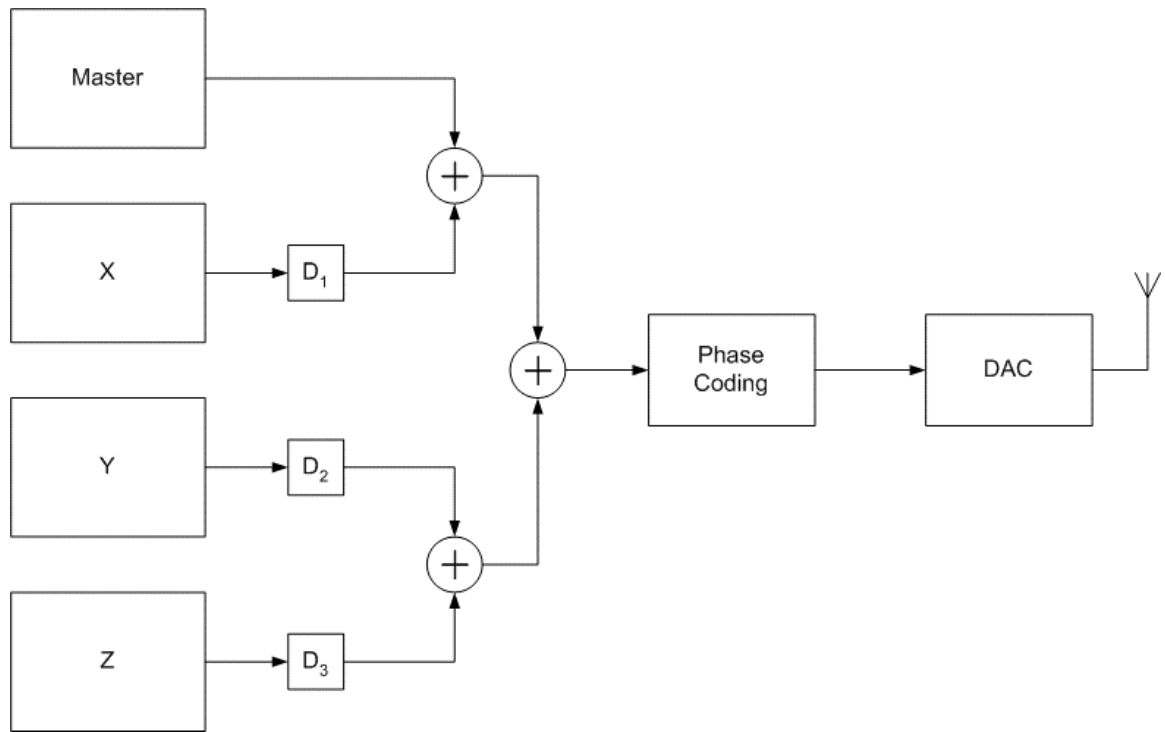


Figure 4.2 – Closeup Illustration of Loran Transmitter Block from figure 4.1.

The signal combinations of EFQPSK used are as follows: A, B, C, D, AB, CD, ABCD, and Fullband. A switching attenuator was used inline with the interfering signal to adjust for SIR. The absolute levels of each signal set were also adjusted to the input range of each receiver. Inline attenuators were added to ensure no receiver was overloaded and linear operation was preserved.

Because there is much non-linear processing, especially during the acquisition phase, in a Loran receiver, it was decided to only report results that could be consistently notated. Any other observations were left for later consideration. So, in the case of the Furuno receiver, the position locking time, T_{acq} , was the observable. In the higher grade Locus SATMATE receiver, lock capability was observed. This was due to the inconsistency in results in acquisition time attributed to greater non-linear processing and our suspicions that the SATMATE receiver was difficult to 'spooft'. It was suspected because with the same signals in which the less expensive Furuno receiver locked with little difficulty, the Locus unit would vary greatly in many observables such as lock capability, acquisition time, etc. We believe the Locus receiver may know more what to expect in terms of signal shape due to more accurate tables of channel ASFs. It is likely too intelligent a receiver to spoof unless great care was to be taken and, for the purpose of the experiments, too time-consuming.

An example of a test on the Furuno receiver is given. First, the SIR is determined and all levels are adjusted to their appropriate values. The receiver's manual recommends that upon initial startup, the system should be allowed approximately 20 minutes before recording results. This is to allow the adaptive notches to settle to their optimal values. So, when SIR is adjusted, this wait period is allowed before making any recordings. For each SIR, ten T_{acq} values were recorded. Their average was then recorded as the data point for that particular SIR. The results were recorded for all interference variations and desired SIR values.

The values of SIR refer to the average power of the Loran received signal over the average interferer power. At the time of the measurements there was no known universal method for treating the average power in Loran signals since they are pulsed and with varying fundamental periods depending on the GRI of the particular chains in question. Thus, the author decided to treat the average power as the total energy in a single pulse divided by a 300us period. This period was chosen because

99.9% of the energy is contained in this time frame and due to the low duty cycle of the pulses anything greater in time might yield figures that could become misleading to the reader. It was later discovered that for signal to noise ratio, as reported by Locus, Inc., the instantaneous power at the peak of the third crossing would yield, within 0.5dB, the same results assuming average power of the CWI was used for the noise measurement and other noise sources were neglected.

4.2.3 Results

As one can see from the plot of the Loran spectrum, 2.2, 99.9% of the energy in a single Loran pulse is contained in the band of 90-110 kHz. In fact, most filters for Loran receivers are set to reject energy outside this band. So, as seen in Table 4.1 and Figure 4.3, it is not surprising that signals A&B do not provide nearly as much interference as the in-band signals, C & D. However, without the presence of filtering one would see a greater amount of interference from the A&B versus C&D signals due to phase errors [1]. So, with increased filtering capabilities for those bands, one may see an even greater amount of rejection of the A&B signals.

The results show that with a single A & B pair, one can have negative SIR at the receiver front end down to -18 dB before consistent failure in acquisition performance will occur. The proposed design was to have a front end SIR floor of 10-20dB at the station. Within a chain, it is suspected that no noticeable difference in performance (accuracy and reliability) is likely to be observed in either commercial or laboratory grade Loran receivers. However, it is possible that the SIR floor may limit the total number of available stations when operating differentially. This cannot be said with much confidence either way, of course, without significant studies into the advanced processing of more contemporary receivers. Also, these receivers are operating blind in the sense of their lack of awareness of the CWI signal nature. If implemented, one could add new capabilities to the receivers to better reduce the energy of the EFQPSK waveforms entering into the Loran receiver blocks. This added a priori knowledge could give several dB improvement in the total rejection and increase the overall Loran-C system performance plus have the additional overlay signals for increased navigational awareness. Ultimately, this study was to

Table 4.1 – Minimum performance requirements for QPSK over Loran in Furuno Rx

Furuno Testing	Minimum S/I (dB)
Fullband - Shaped	-1
Fullband - Unshaped	-1.6
AB	-18
CD	-3
ABCD (equal)	-7.6
ABCD (CD -10dB)	-13.4
ABCD (CD -20dB)	-14.7
SatMate Testing	
Fullband - Shaped	-5
AB	-19
CD	-1
ABCD (equal power)	-3.5

C.O.T.S. Loran Receiver Acquisition Time versus S/I

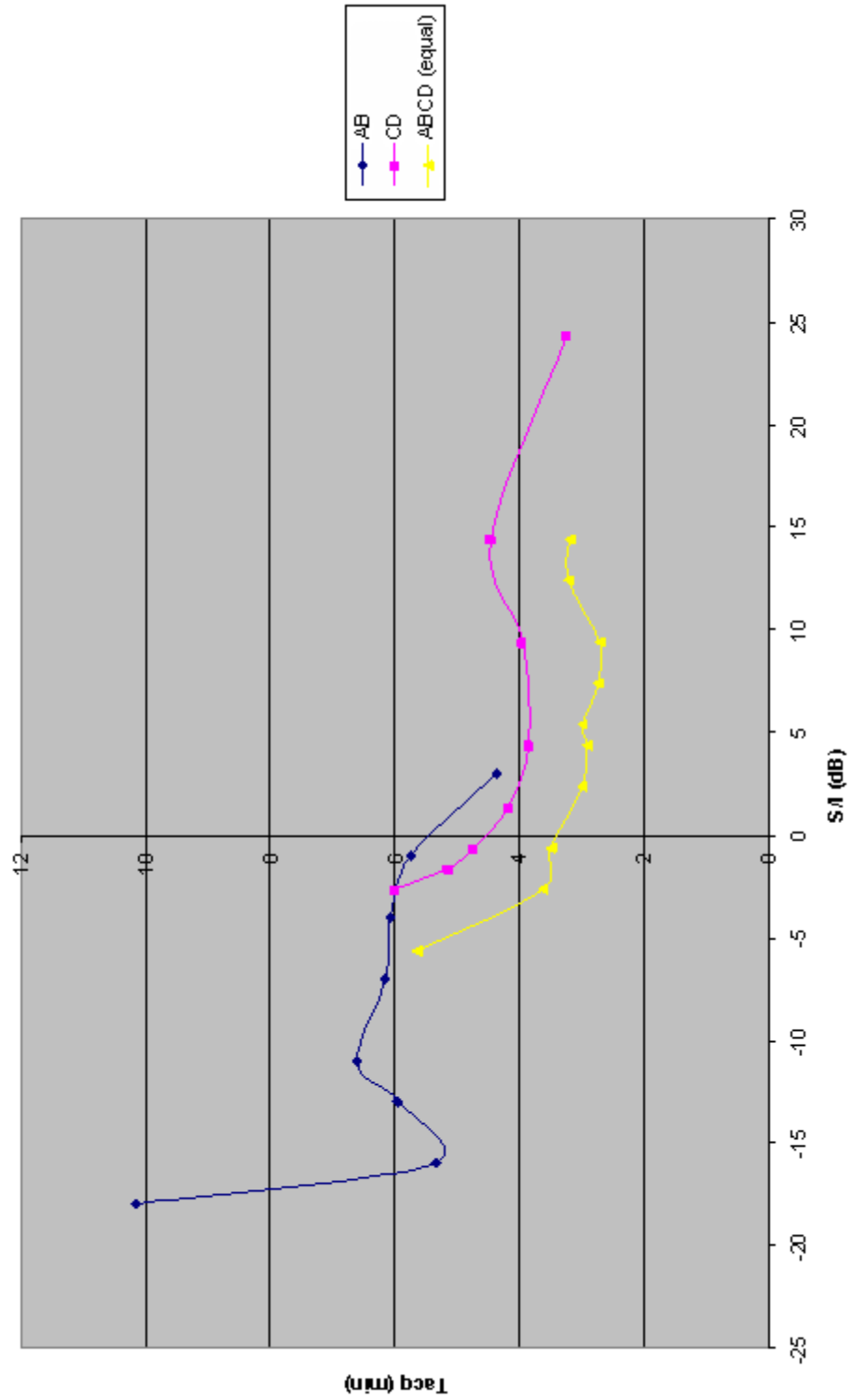


Figure 4.3 – Plot of Signal Acquisition Time versus SIR

show the effects on current and even legacy Loran-C receivers and it appears that compatibility may well exist within the bounds listed above.

Although time did not allow, it would be advantageous to perform a similar study using the alternate symbol structure of the interferers in a laboratory setting. It would be useful to be able to observe asynchronous WB CWI. Also, the averaging time would tend to possibly increase depending upon the location of the actual spectral lines, but the likelihood of acquisition and acceptable performance at even lower SIR values would be almost certain. Thus, with the contemporary receivers operating in differential mode, their allowable pool of stations would increase. Section 4.3 will discuss these possibilities in greater detail.

A more rigorous analytical analysis of just how the receivers are affected by the wideband interference is in order. The literature predominantly discusses narrowband CWI yet acknowledges the presence of spectral spreading due to phase and frequency shift keying. This could prove to be an analytically tedious or possibly intractable analysis, so a simulation model is likely in order to gauge the performance of Loran in the presence of wideband interference.

4.3 Simulations

The next step in understanding Loran-C susceptibility to wideband CWI is computer simulation. The approach taken in 3.3 is extended here to understand the effects statistically. We focus solely on the phase tracking process and leave simulation of the cycle selection for future work. Two separate models were developed. Each focused on different aspects of interference statistics. The first employed the initial interfering signal model, where the symbol rate was 5ksps. The second simulation model employed the modified symbol rate of 5.115ksps. In both models, the assumption of $\phi_{\text{int}} \sim U(0, 2\pi)$ for each CWI with respect to the Loran-C carrier phase.

The phase decoding operation was neglected so the incoming Loran-C signals are assumed all uncoded. Also, any bandpass filtering operations are assumed through a 5th order Butterworth bandpass filter, centered at 100 kHz, and with a passband bandwidth of 20 kHz. Lastly, all models assume a complex envelope representation with $f_0 = 100$ kHz.

4.3.1 Model One – Model

Both models were designed specifically to be compared against results from Last [4]. These simulations were intended to show the joint distribution over all space and time of the phase tracking errors in a Loran-C receiver. Only groundwaves are considered, although it is possible that distant skywaves might also be valid if the observation window is much less than the skywave channel's time constant. The group delay distortion of either channel, including antenna and other distortion effects were neglected because their consideration would have limited the applicability of the results due to variations among receivers and over time.

Two different sets of CWI were utilized in this model; a single A CWI and a pair of A&B CWI. The first set looked at the contribution from a single wideband signal versus its unmodulated counterpart in 3.3. The A&B pair was the most important because it observed the nominal self-interference for a single Loran station, given the assumption of co-located A&B pairs at each Loran station for the overlay design discussed in 4.1. The basic simulation model for the single A interferer is given. First, we generated 10,000 symbols to account for all possible symbol combinations under filtering. Instead of using the Gold codes, they were generated bitwise at the full chip rate using the MATLAB function, *randint*, and then converted to the proper EFQPSK baseband waveforms by the mapping process. The waveform samples were then complex modulated to the appropriate frequency, here -15 kHz. The random phase vector ϕ_{int} was generated as a column vector and the CWI vector multiplied it, creating a matrix of source CWI samples. Next, a complex constant was added to the matrix. This constant represented the magnitude and phase of the complex envelope at the standard crossing point. The phase argument was then taken of each element and then subtracted from the ideal Loran-C carrier phase. The matrix values were then passed to a histogram variable. This process was repeated until sufficient samples were employed such that the estimator variance was minimized and consistent estimates could be obtained. A separate parallel operation was also performed, which operated on the filtered CWI signals. Several simulation runs were made at different SIR values and the results for unfiltered and Butterworth filtered [1] phase errors resulted. The process was then extended to the

case of 2 interferers. For each CWI, a separate independent random phase, ϕ_{int} , was generated. The symbol timing was assumed identical for all CWI.

4.3.2 Results – Model One

The results for the first simulation model are given below in Figures 4.4-4.15. First, rectangular OQPSK was tested along with the desired modulation in the simulation with a single 'A' signal to illustrate how the reduction in sidelobe energy translated to timing errors. This can be seen in Figures 4.6, 4.7, 4.10, and 4.11. The peak phase tracking error is almost half between the two modulations. For all cases of unfiltered CWI statistics, the distributions were the same. The assumption of equally likely symbols allows this. The pdf estimates in [4] also verify the case for unfiltered CWI. It should be noted that these merely represent raw results without any other receiver operations. Phase coding, averaging, and notch filtering can greatly reduce these errors. In addition, the use of linear phase digital filtering would allow the critical Loran envelope.

4.3.3 Simulations – Part Two

The second simulation model employed slightly modified CWI. Their symbol rate was changed to 5.115ksps. The justification for this modification is given. The filtering process at the receiver front end allows higher frequency components from the CWI signals' information. The higher frequencies in messaging occur during symbol transitions. Thus, the greatest interfering energy will pass through periodically. Then energy takes its form in 'chirps' of varying shape and duration depending upon the actual symbols being transitioned, the SIR, and the relative carrier phases. By modifying their period slightly, we eliminated any periodicity such that the statistics were the same as for a mobile user in time and space. As a result, there is slightly more interfering energy allowed through the BPF. However, the statistics were more consistent, giving less likelihood of more extreme errors for stationary users. This model was very similar to the first. Since we can now say the symbol rate is effectively non-synchronous with the sampling scheme, we assume the error statistics to be equal for all relative 'initial' timing. Thus, a model such as in

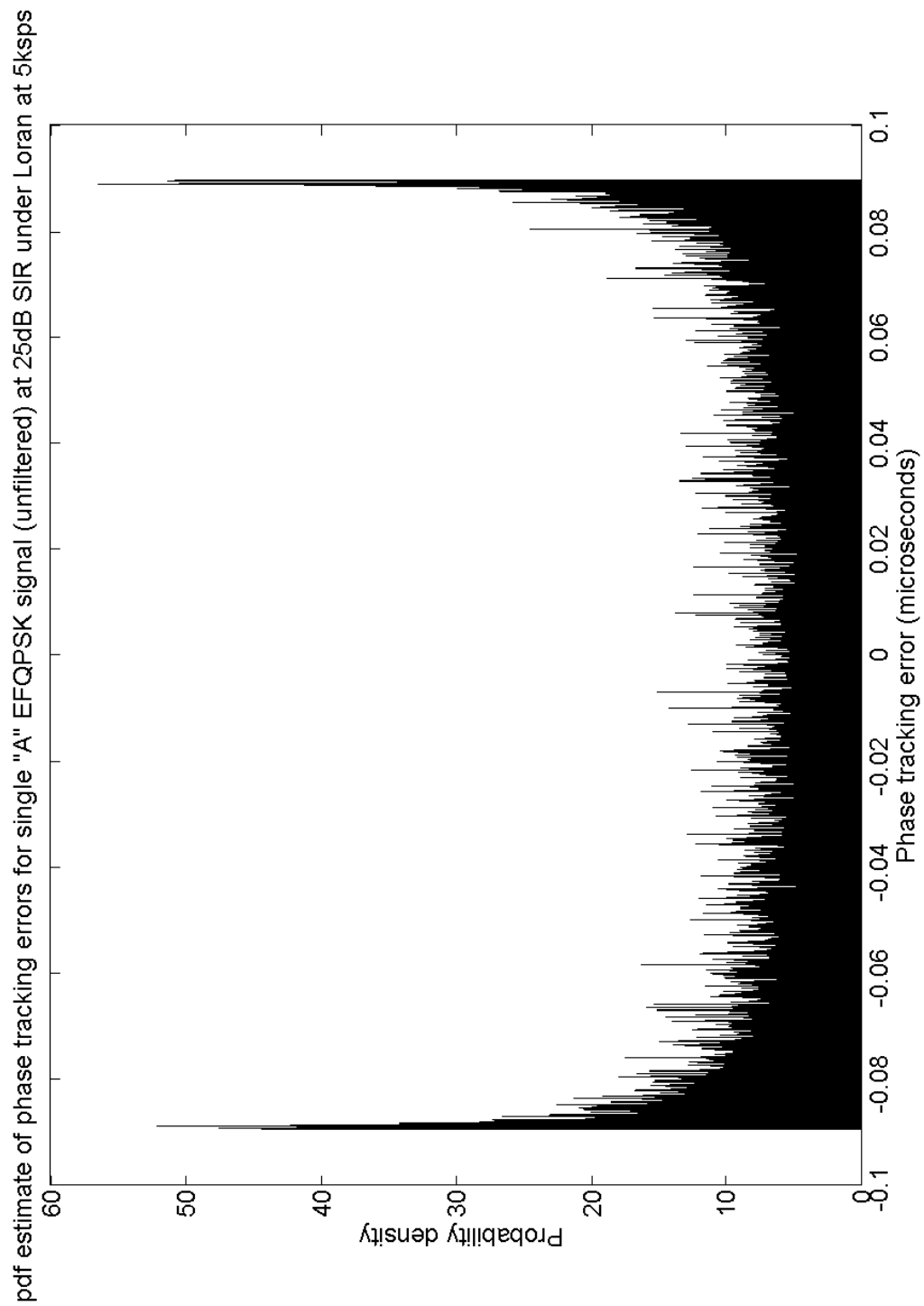


Figure 4.4 – Simulation #1: Unfiltered 'A' signal at SIR=25dB.

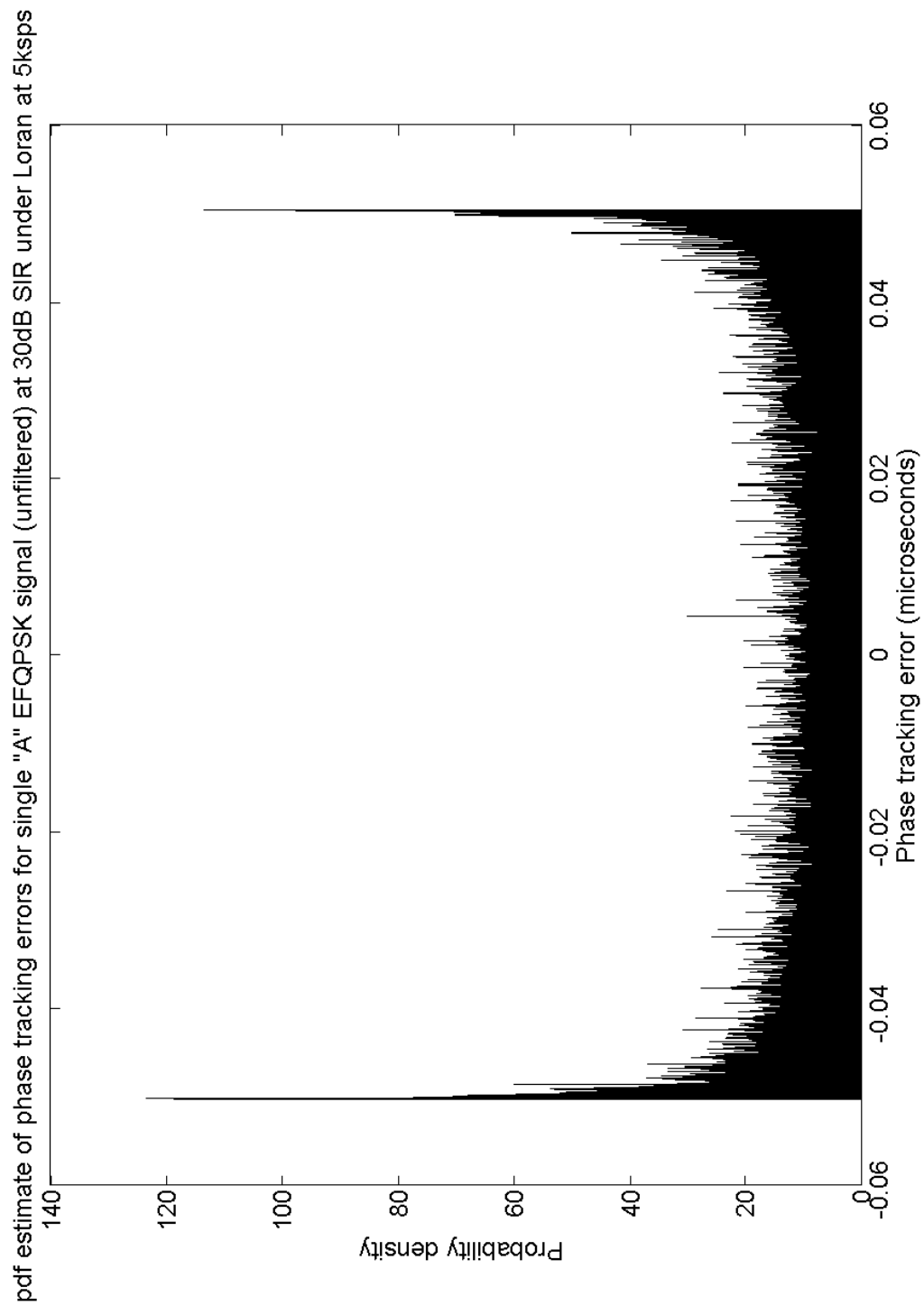


Figure 4.5 – Simulation #1: Unfiltered 'A' signal at SIR=30dB.

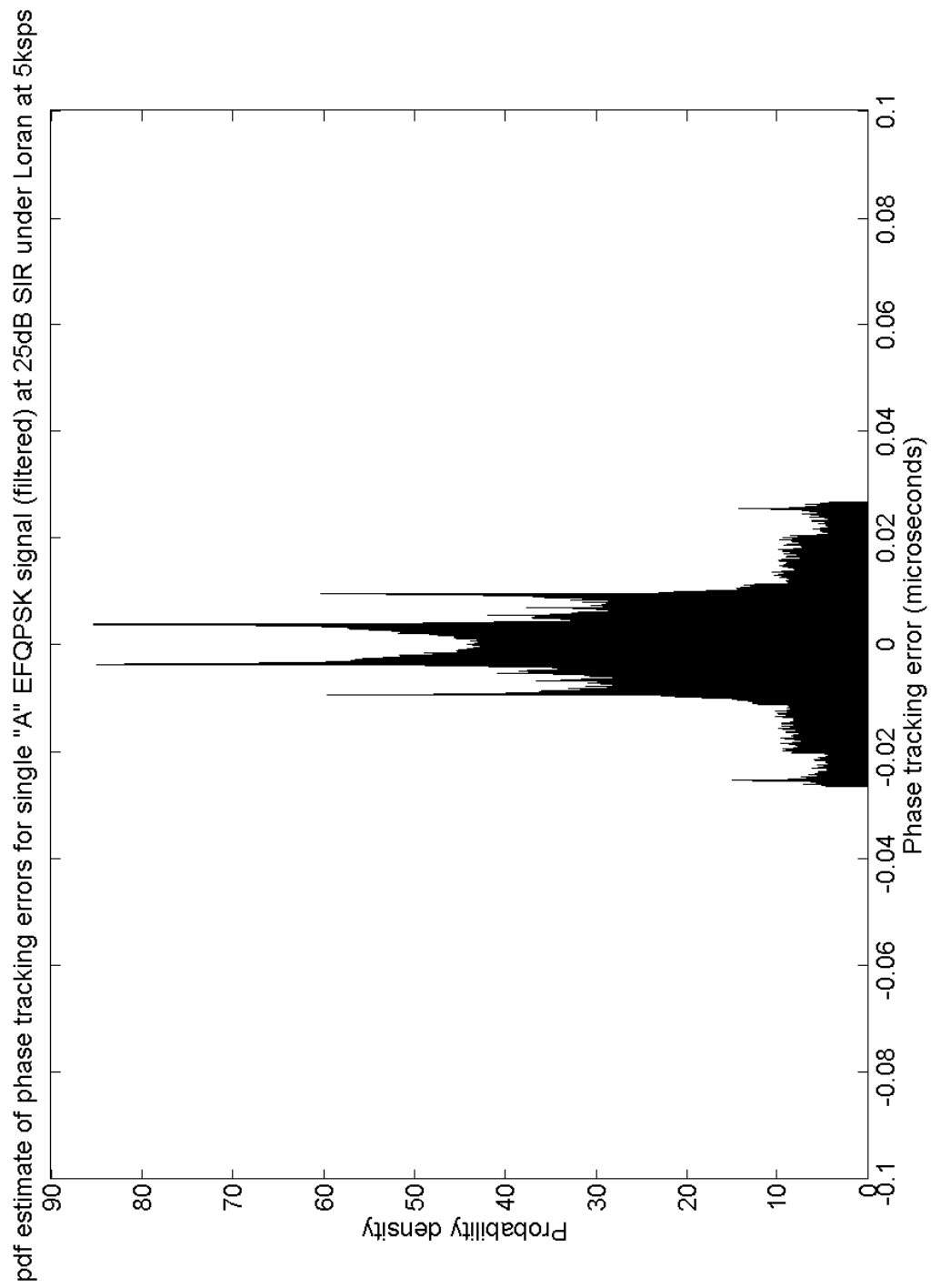


Figure 4.6 – Simulation #1: Filtered 'A' signal with SIR=25dB.

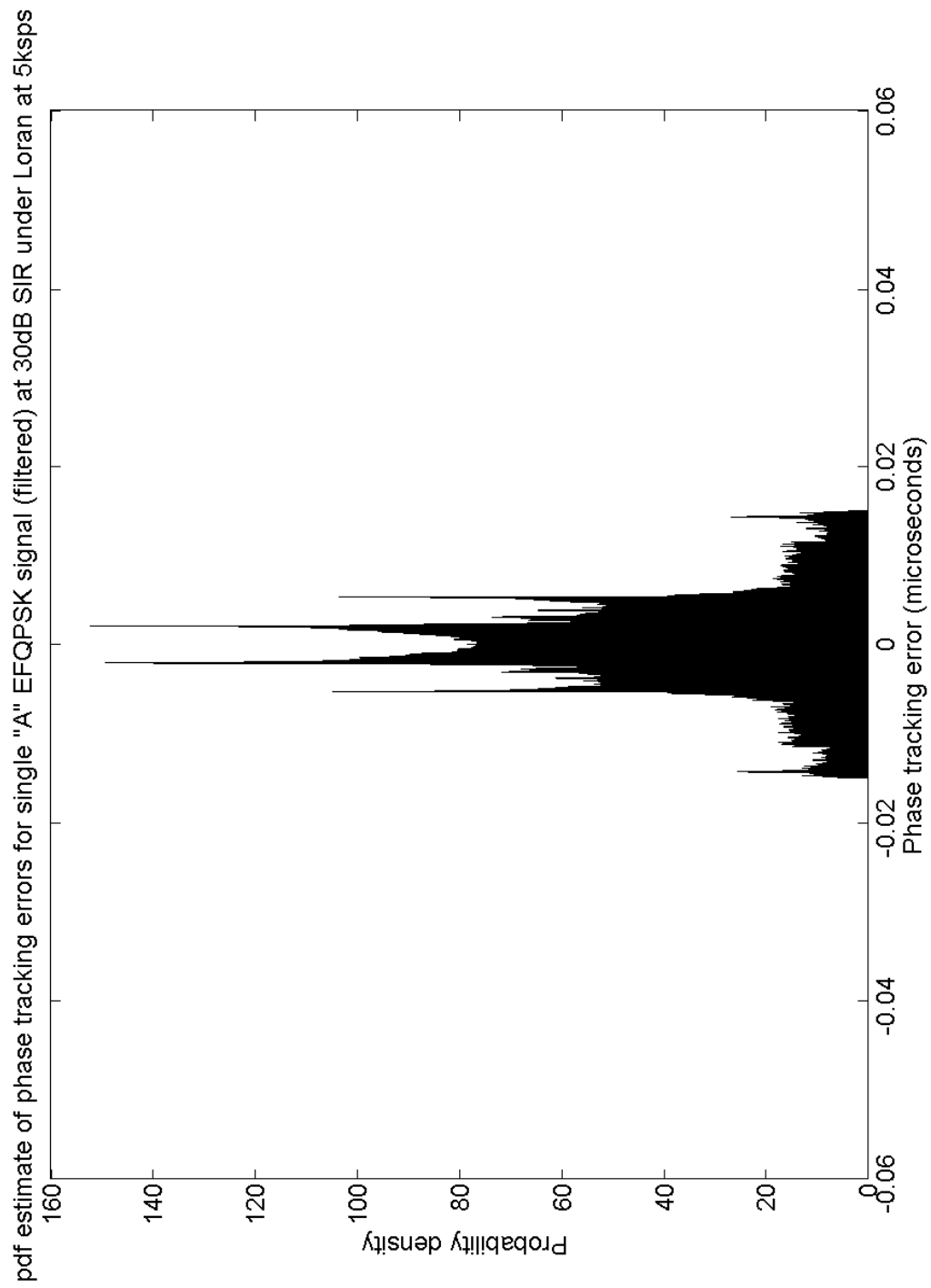


Figure 4.7 – Simulation #1: Filtered 'A' signal with SIR=30dB.

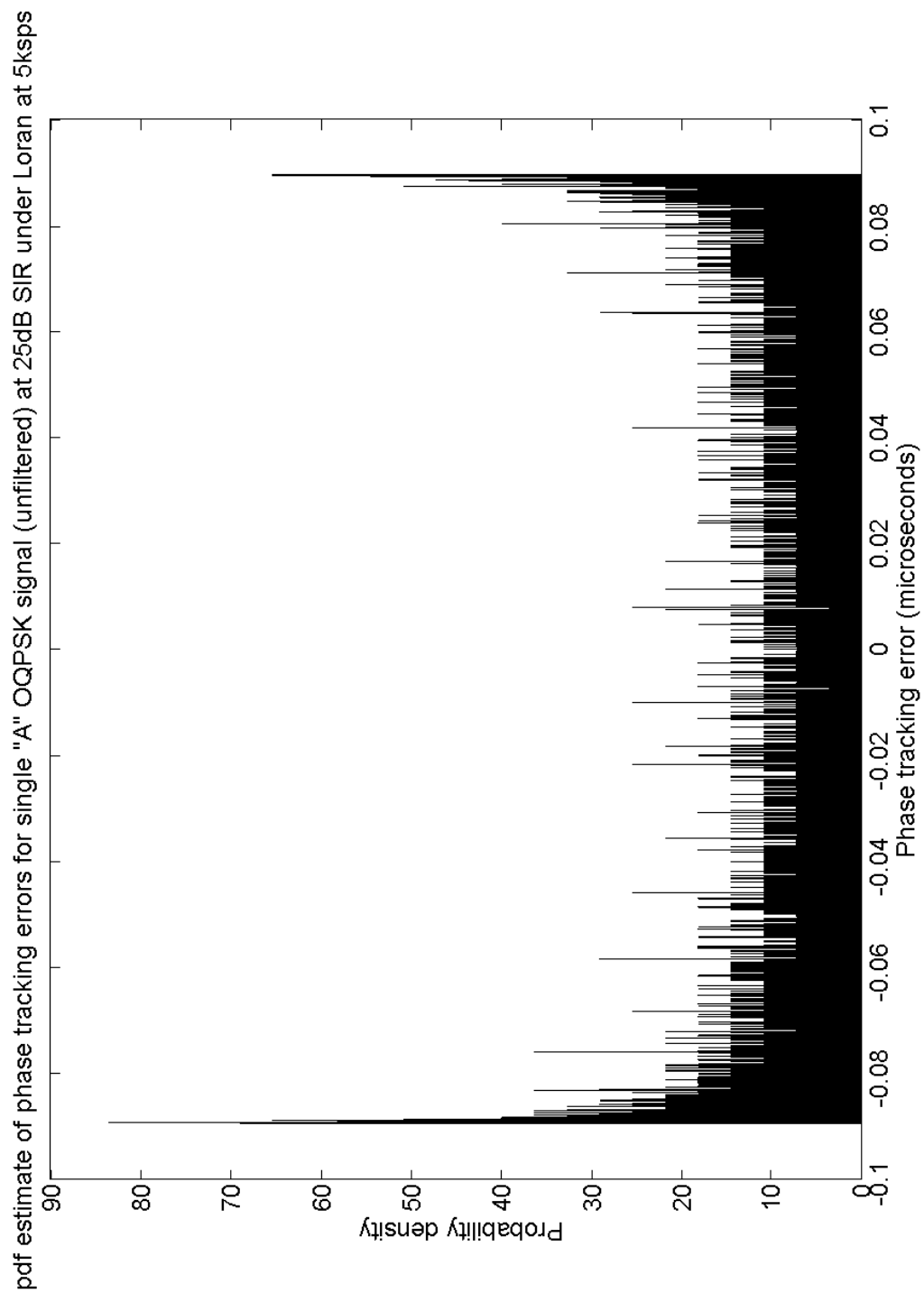


Figure 4.8 – Simulation #1: Rectangular QPSK (unfiltered) 'A' signal with SIR=25dB.

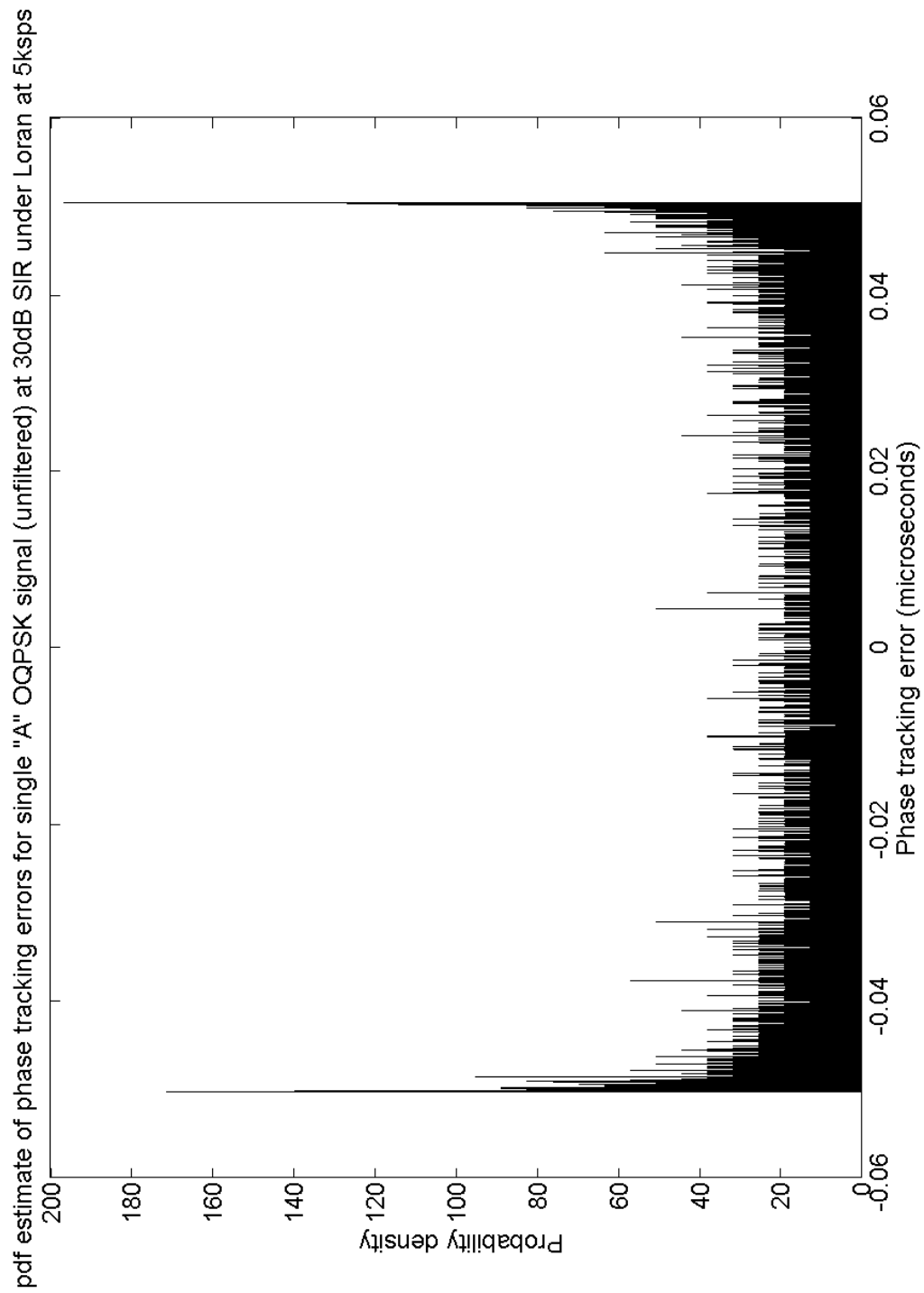


Figure 4.9 – Simulation #1: Rectangular QPSK (unfiltered) 'A' signal with SIR=30dB.

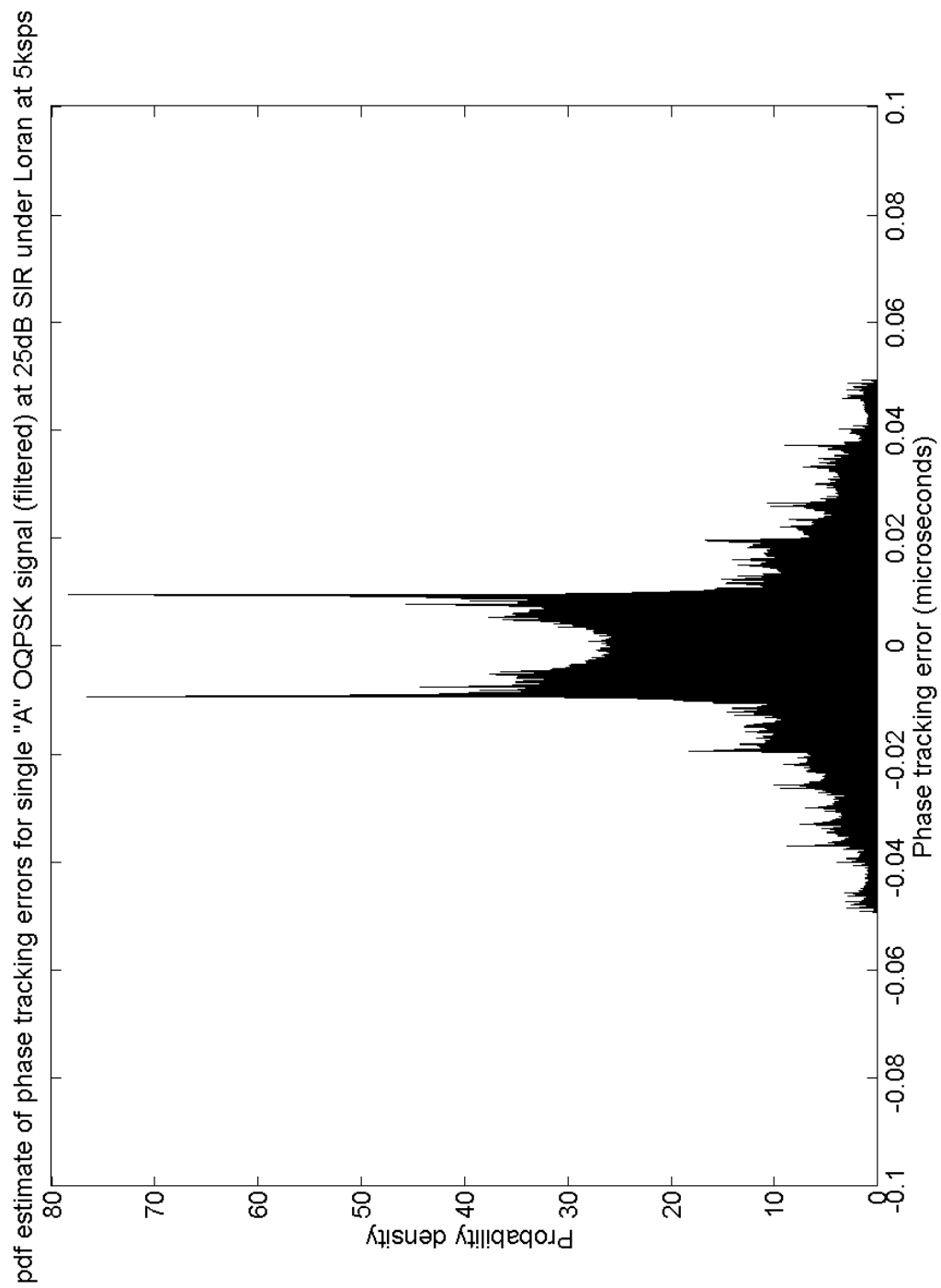


Figure 4.10 – Simulation #1: Rectangular QPSK (filtered) 'A' signal with SIR=25dB.

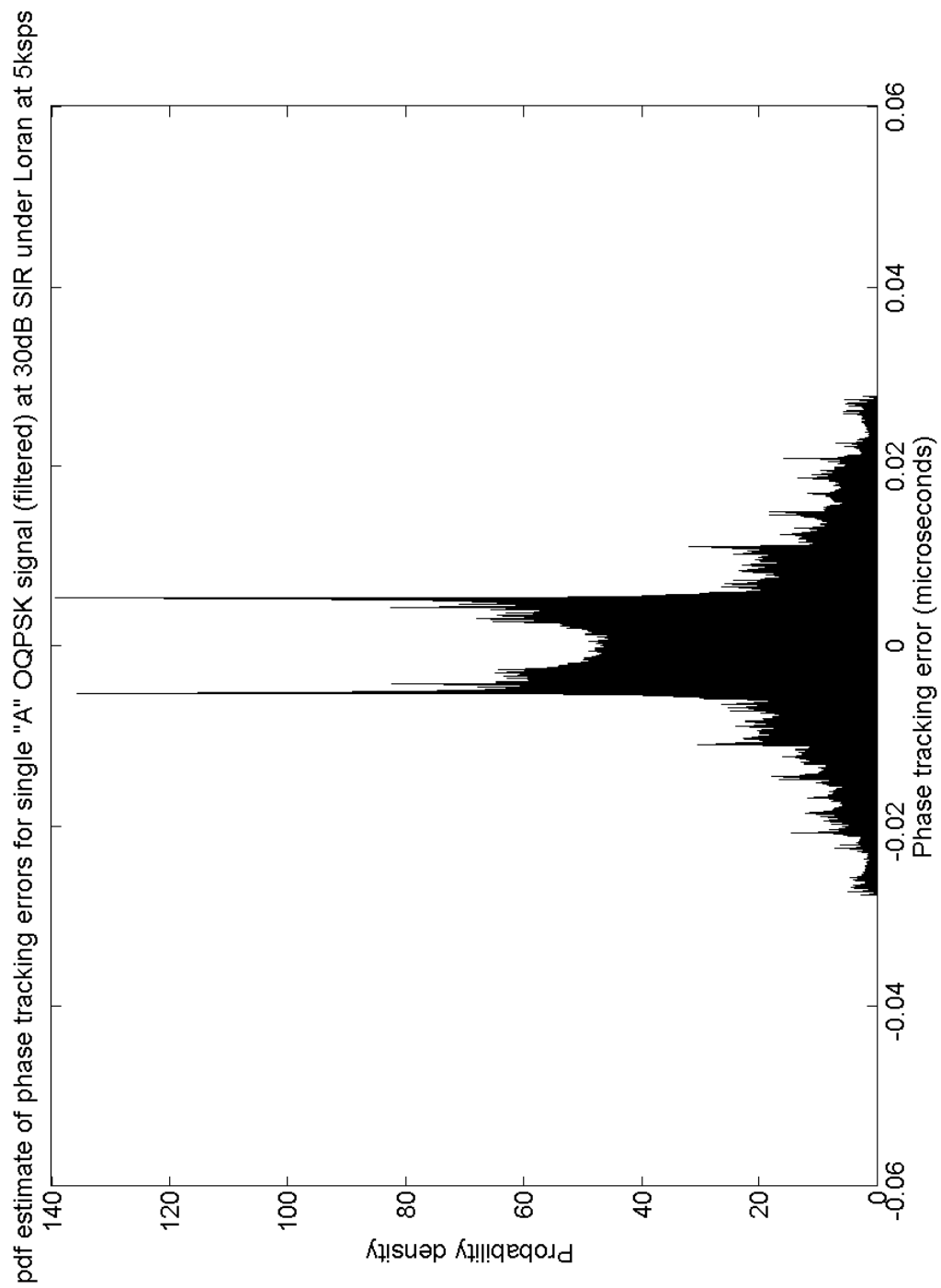


Figure 4.11 – Simulation #1: Rectangular QPSK (filtered) 'A' signal with SIR=30dB.

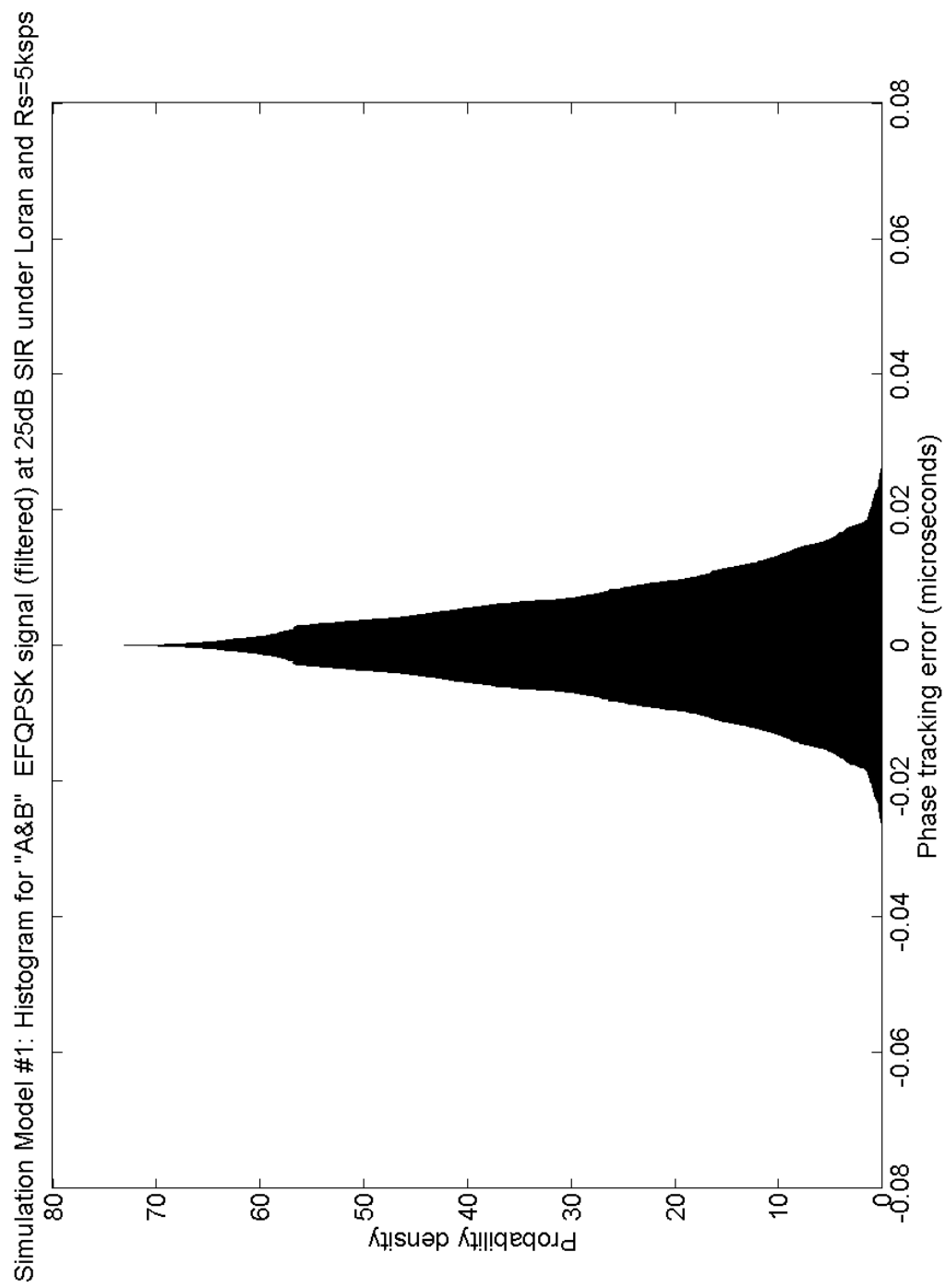


Figure 4.12 – Simulation #1: Filtered AB pair at SIR=25dB.

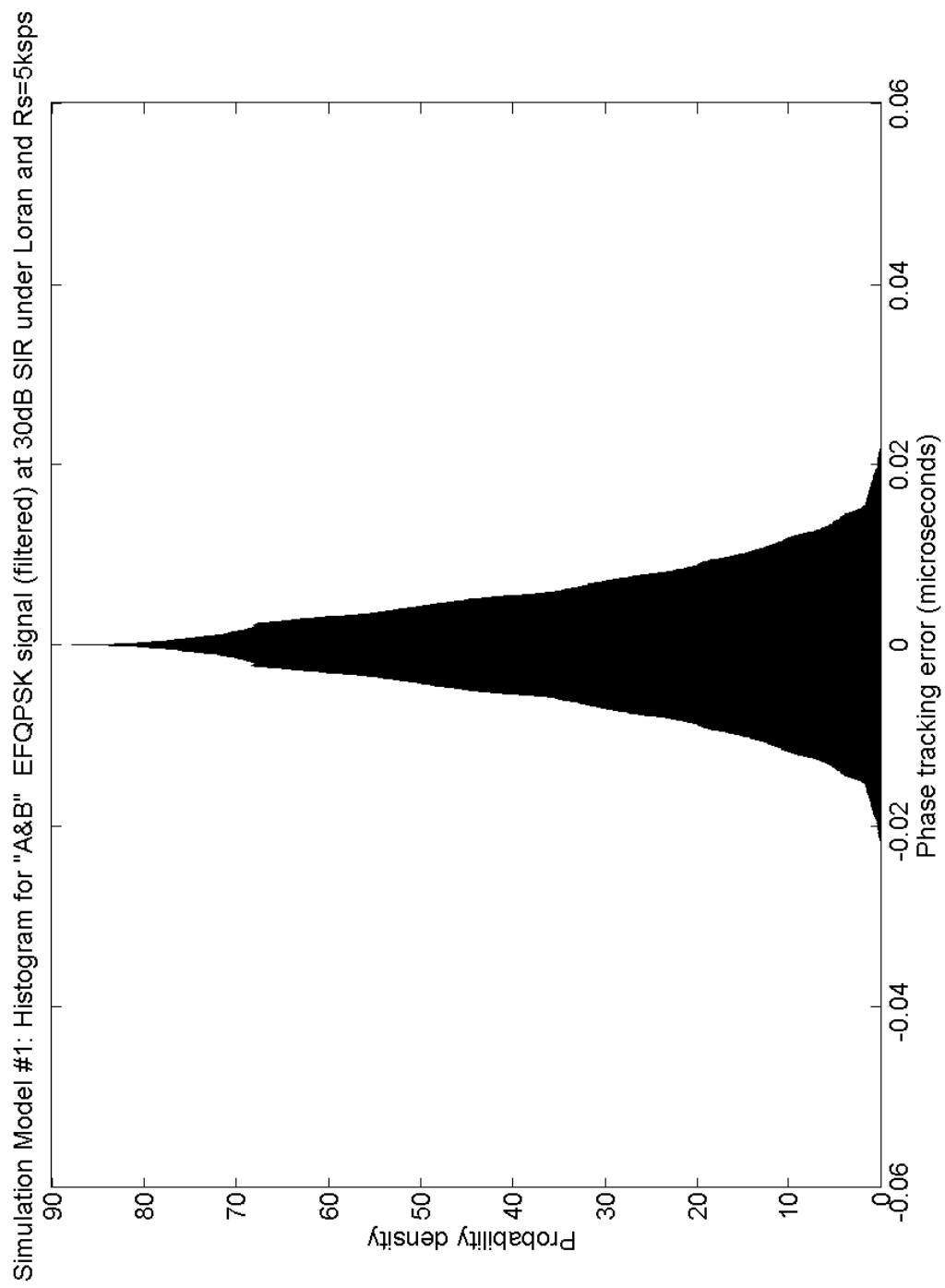


Figure 4.13 – Simulation #1: Filtered AB pair at SIR=30dB.

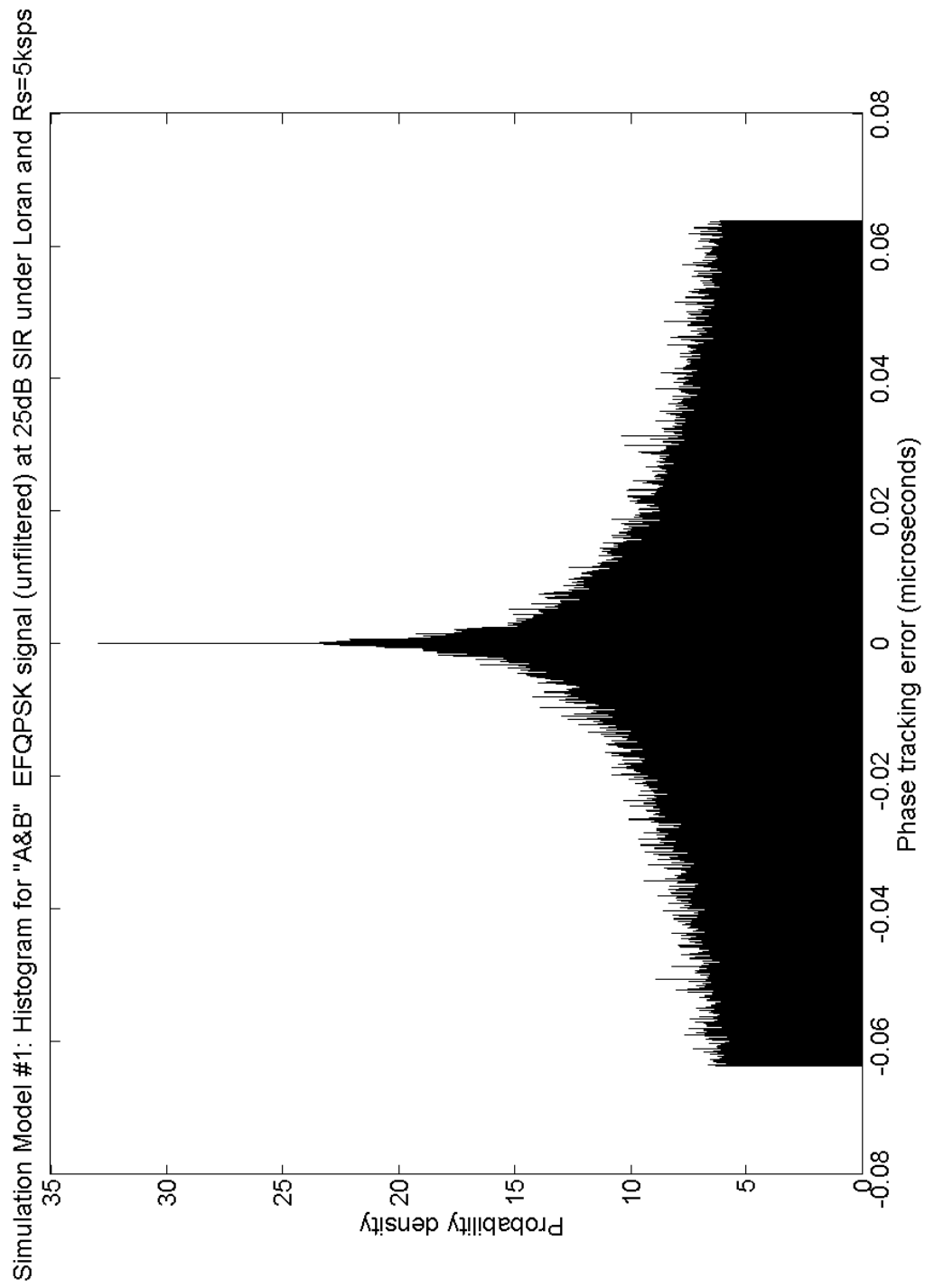


Figure 4.14 – Simulation #1: Unfiltered AB pair at 25dB.

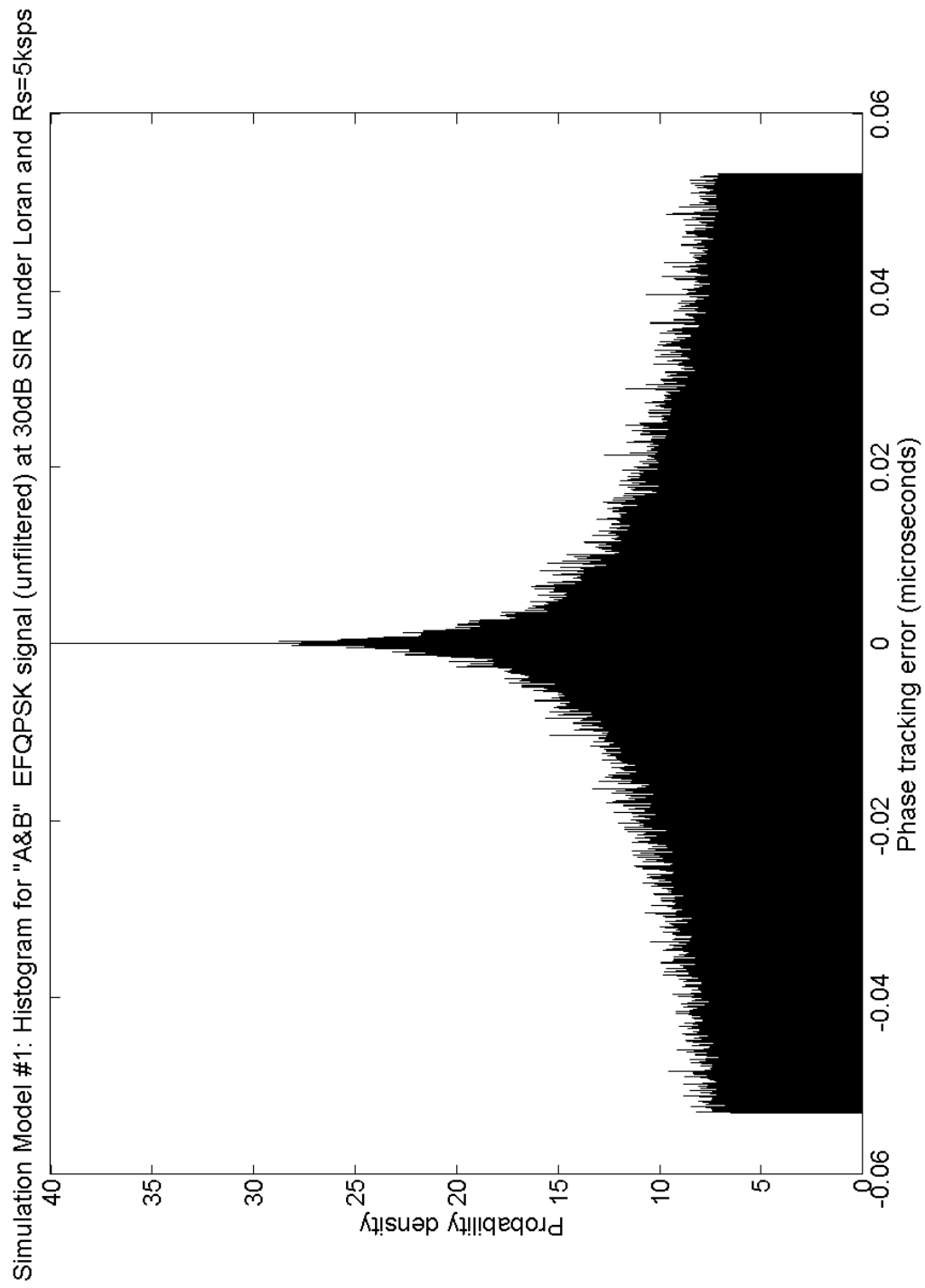


Figure 4.15 – Simulation #1: Unfiltered AB pair at 30dB.

part two was not necessary. We decided to focus simply on the ensemble over all time and space as in part one for 1 and 2 CWI self interferers.

The software model was different from part one in only one aspect, the symbol rate. Very little code needed modification to translate to the study of the new signal structure. The results here were compared against those in part one to illustrate the effective increase in interfering energy resulting from the higher symbol rate.

4.3.4 Results – Part Two

At 5.115ksps, the bandwidth expansion allows slightly more energy into the Loran receiver. We conducted these experiments to observe the difference in reported phase tracking error statistics and the results are given below in Figures 4.16-4.23. As one can see, this easily verifies that the new symbol rate negligibly affects the error statistics, while reducing hazards for stationary users. As for the first model, additional processing may reduce phase tracking errors tremendously. For example, laboratory test results in 4.1 show that a combined SIR of -18dB for an A&B signal pair in a commercial Loran receiver is possible before failure consistently occurs in position reporting. Thus, these simulations are overly pessimistic in their reporting.

4.4 Conclusions and Future Work

The carriers and symbol rates are synchronous in the original design, which suggests [3, 4] that early cycle slips may occur. For future work, we suggest repeating laboratory experiments with the new symbol rate. Also, with the updated software, multiple boards can be run reliable from a common clock. Thus, control over the relative symbol timing in the laboratory studies is possible. Long term drift can also be minimized, yielding more consistent results. The addition of more advanced prototype filters would be advantageous to the designers. Currently, the funding for this particular part of the project has been moved elsewhere. However, designers are in talks with other funding sources that may well desire to continue work in this specific direction. It is hoped that the work we performed here helped to further this project and was commensurate with the expectations of successful graduate research.

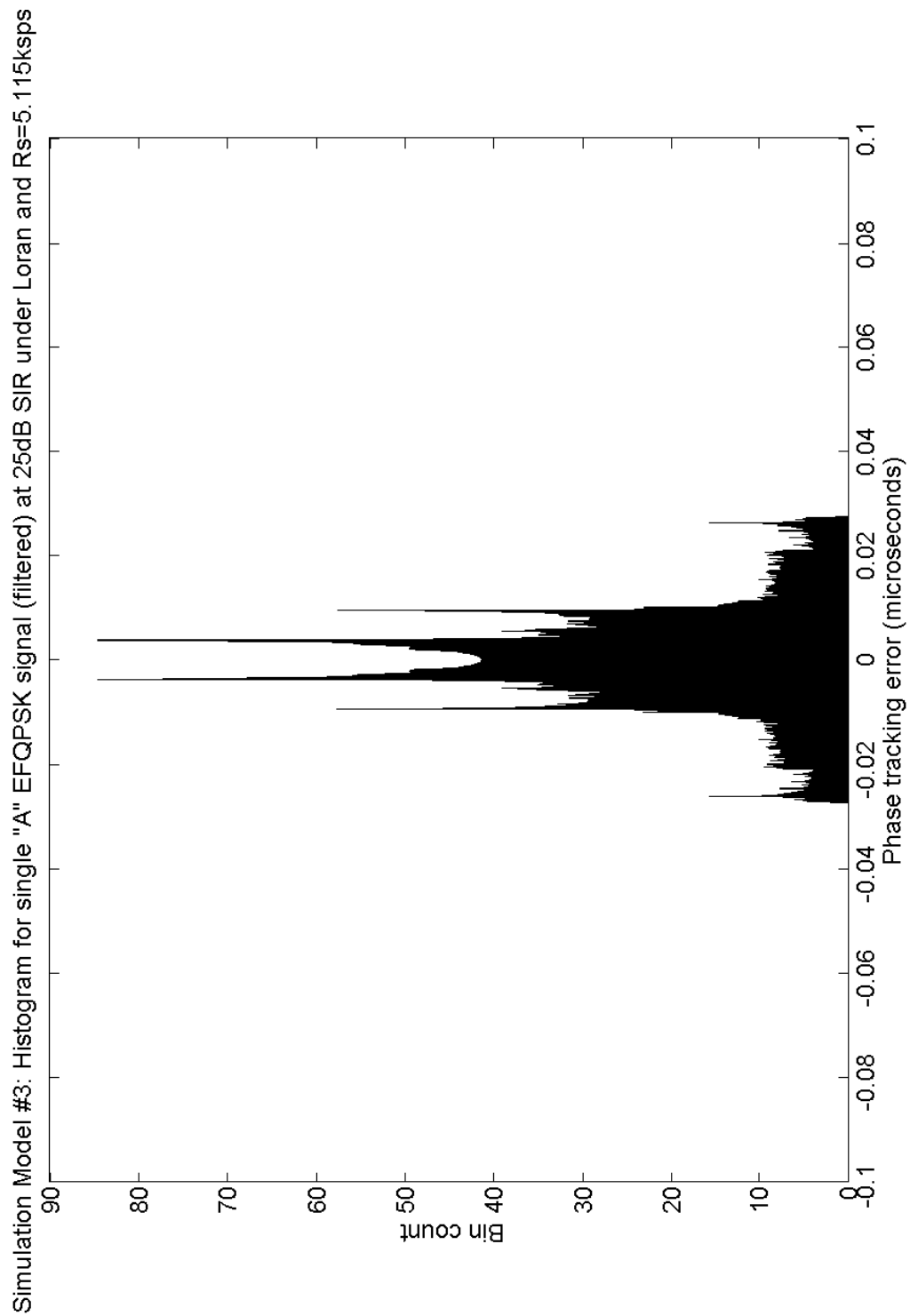


Figure 4.16 – Simulation #2: Filtered 'A' signal at SIR=25dB.

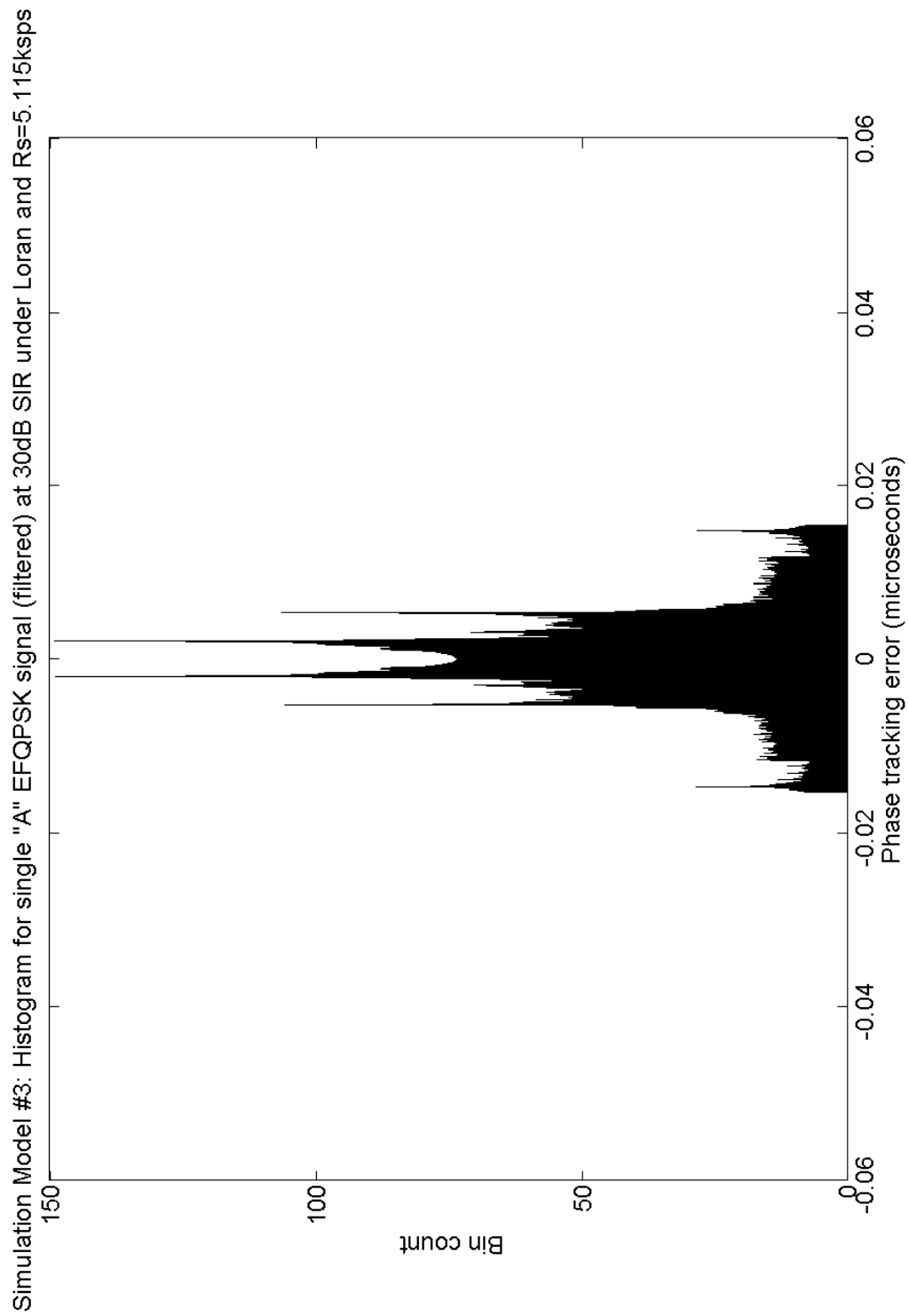


Figure 4.17 – Simulation #2: Filtered 'A' signal at SIR=30dB.

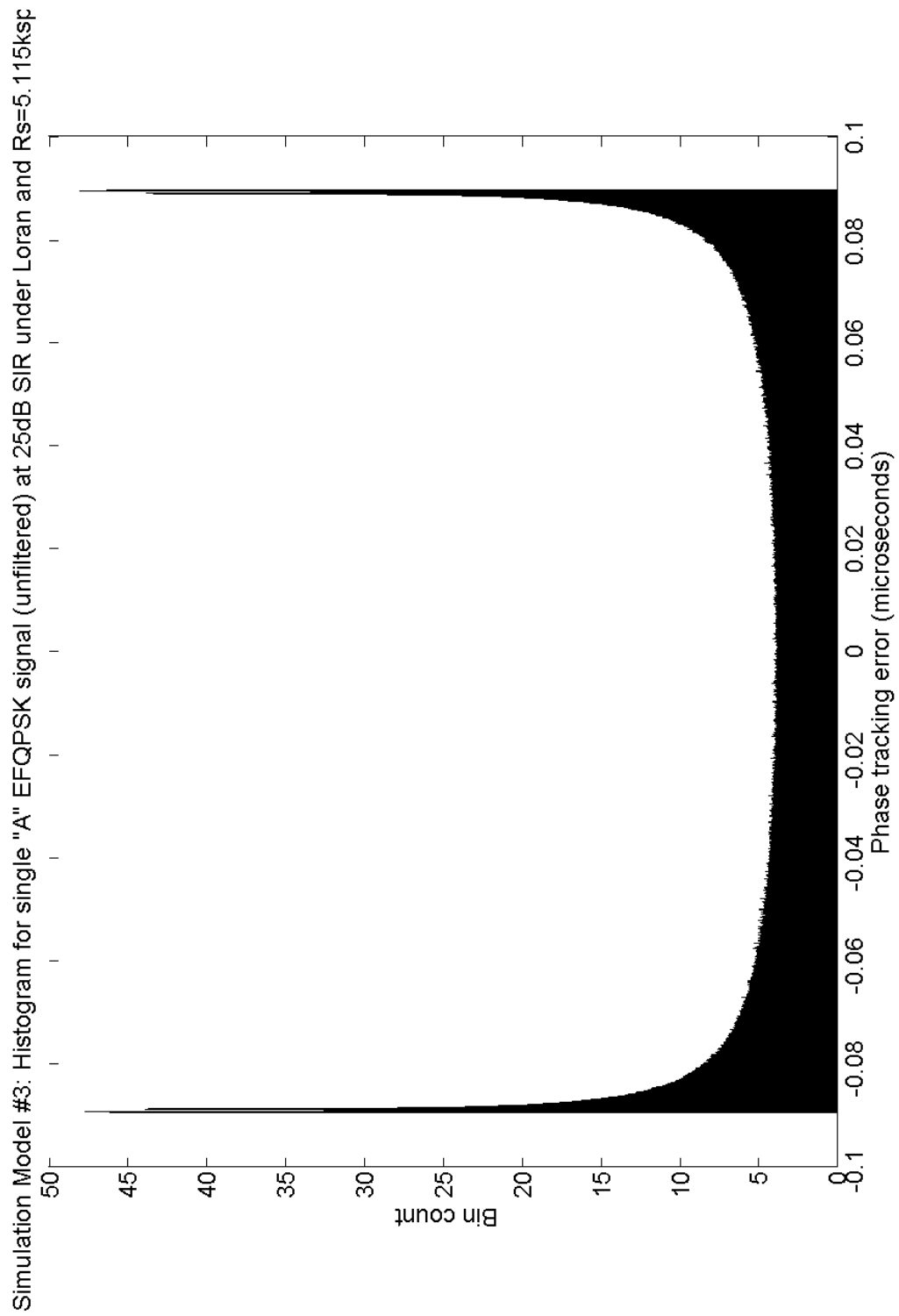


Figure 4.18 – Simulation #2: Unfiltered 'A' signal at SIR=25dB.

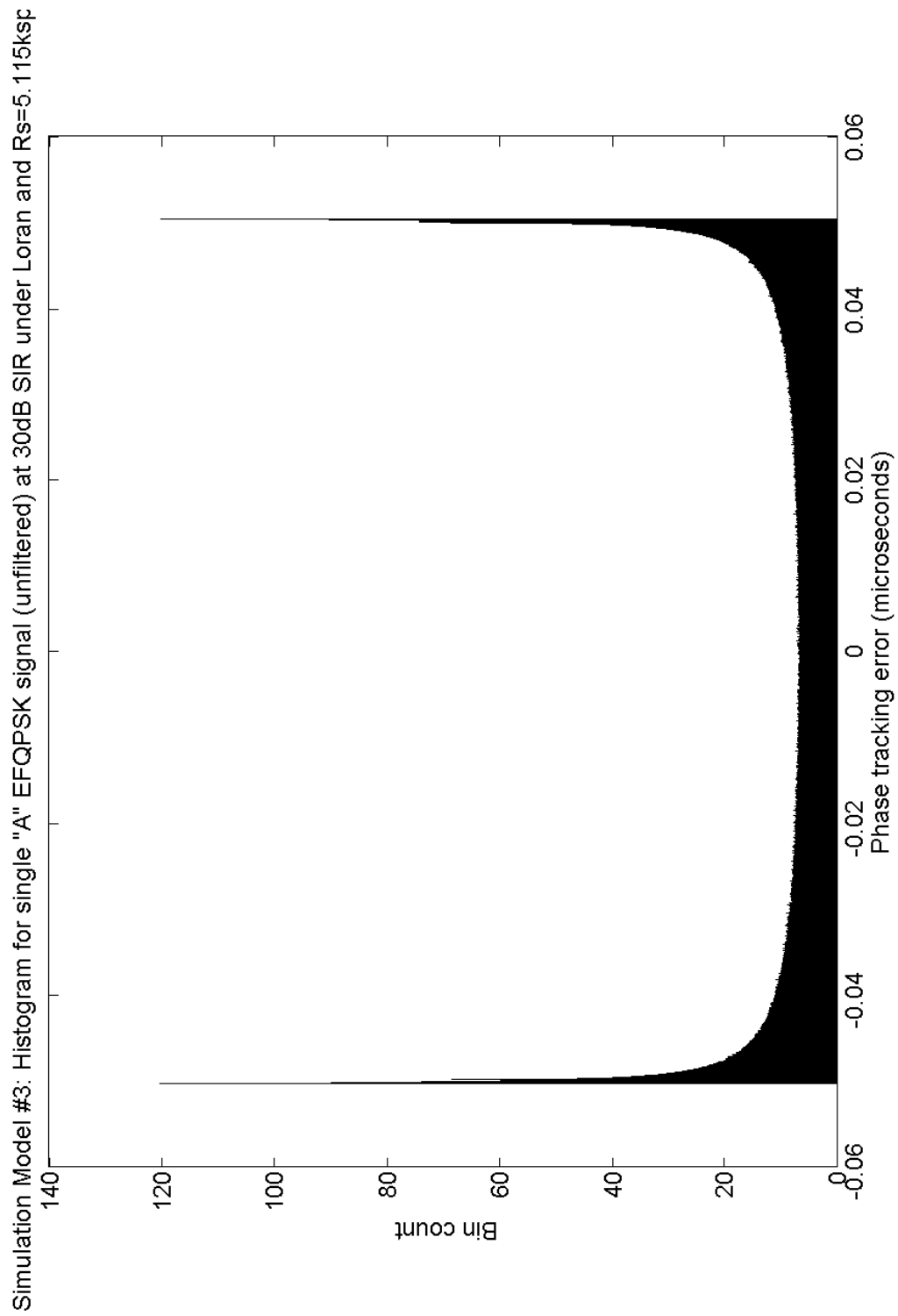


Figure 4.19 – Simulation #2: Unfiltered 'A' signal at SIR=30dB.

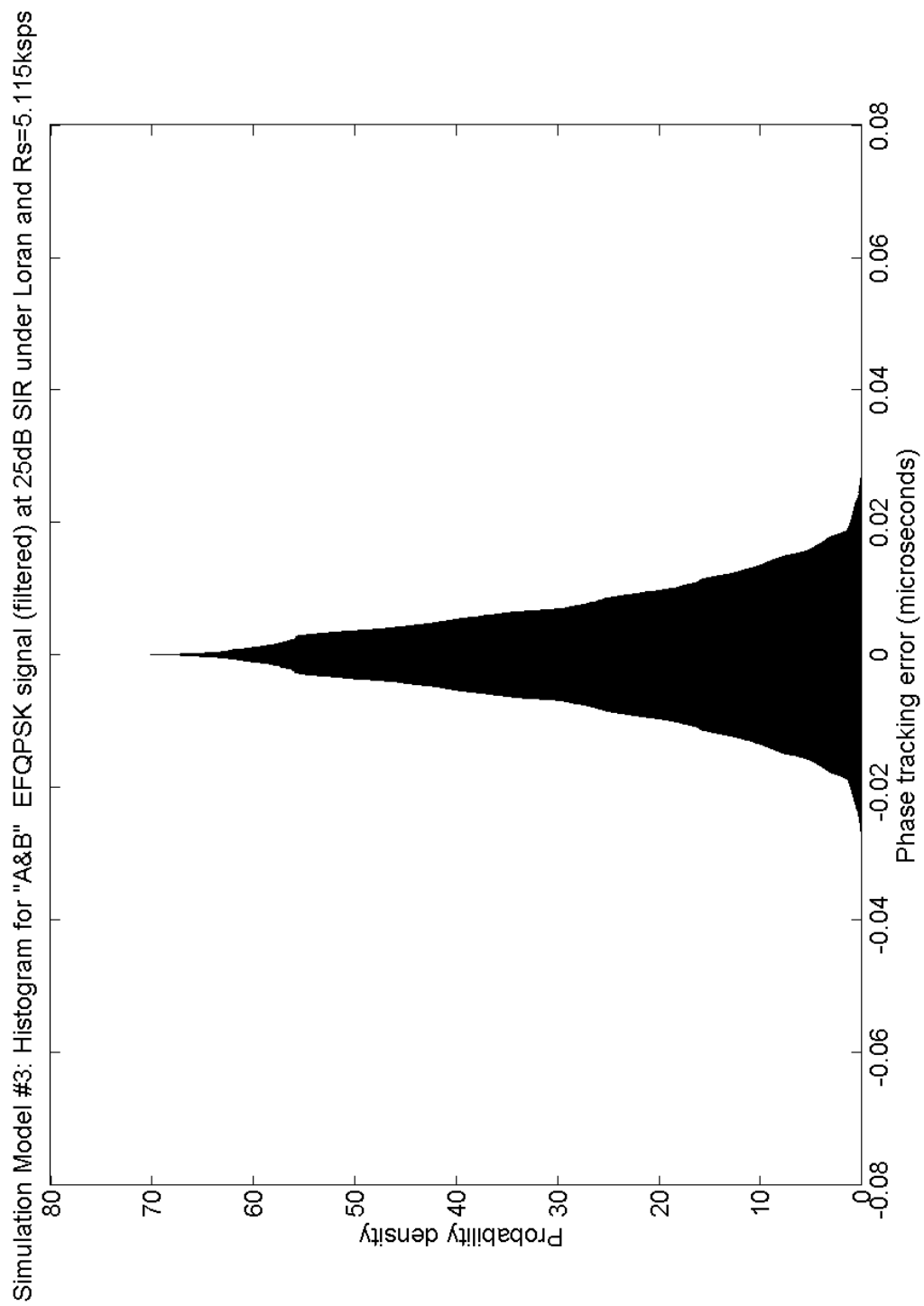


Figure 4.20 – Simulation #2: Filtered 'AB' signal at SIR=25dB.

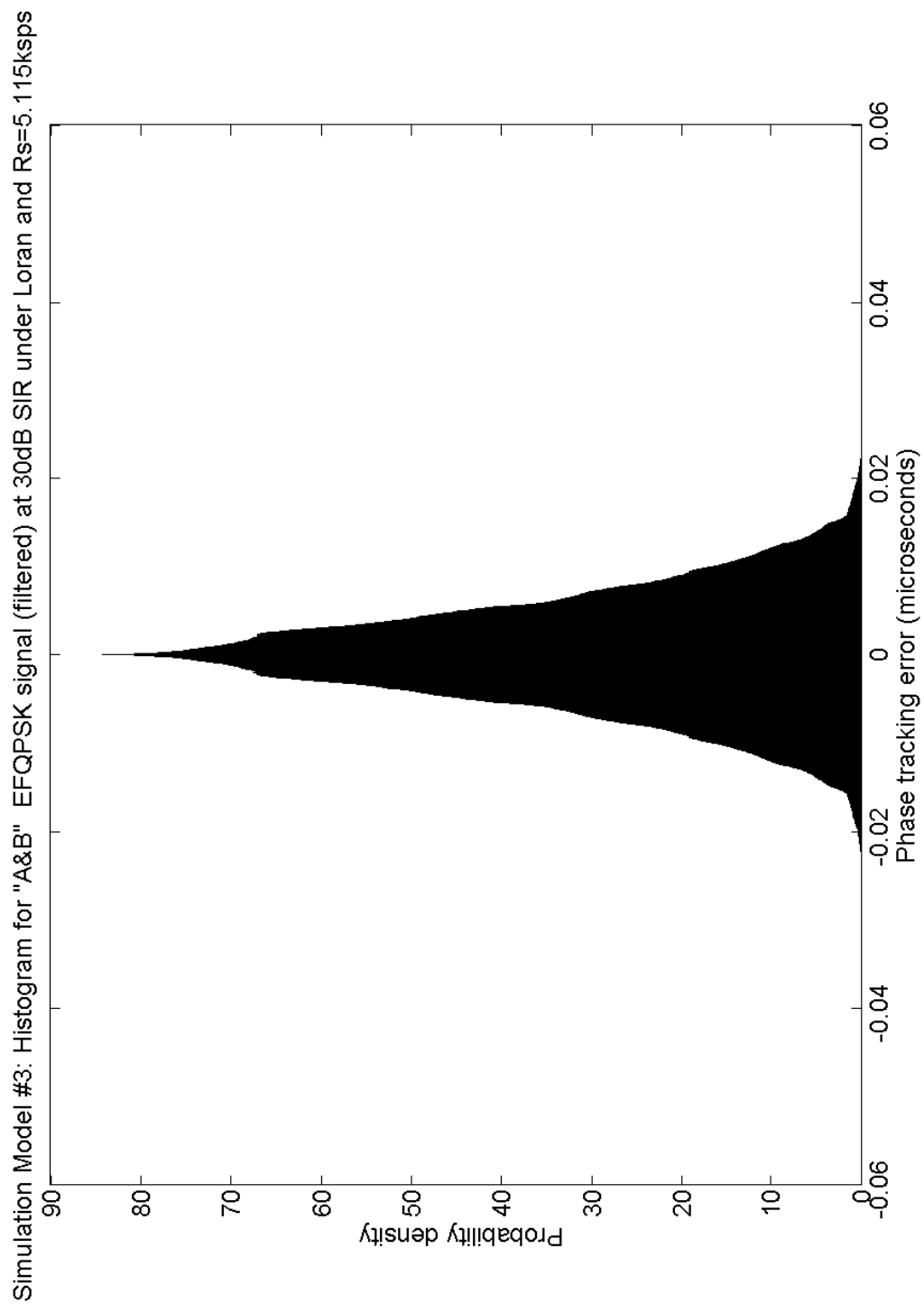


Figure 4.21 – Simulation #2: Filtered 'AB' signal at SIR=30dB.

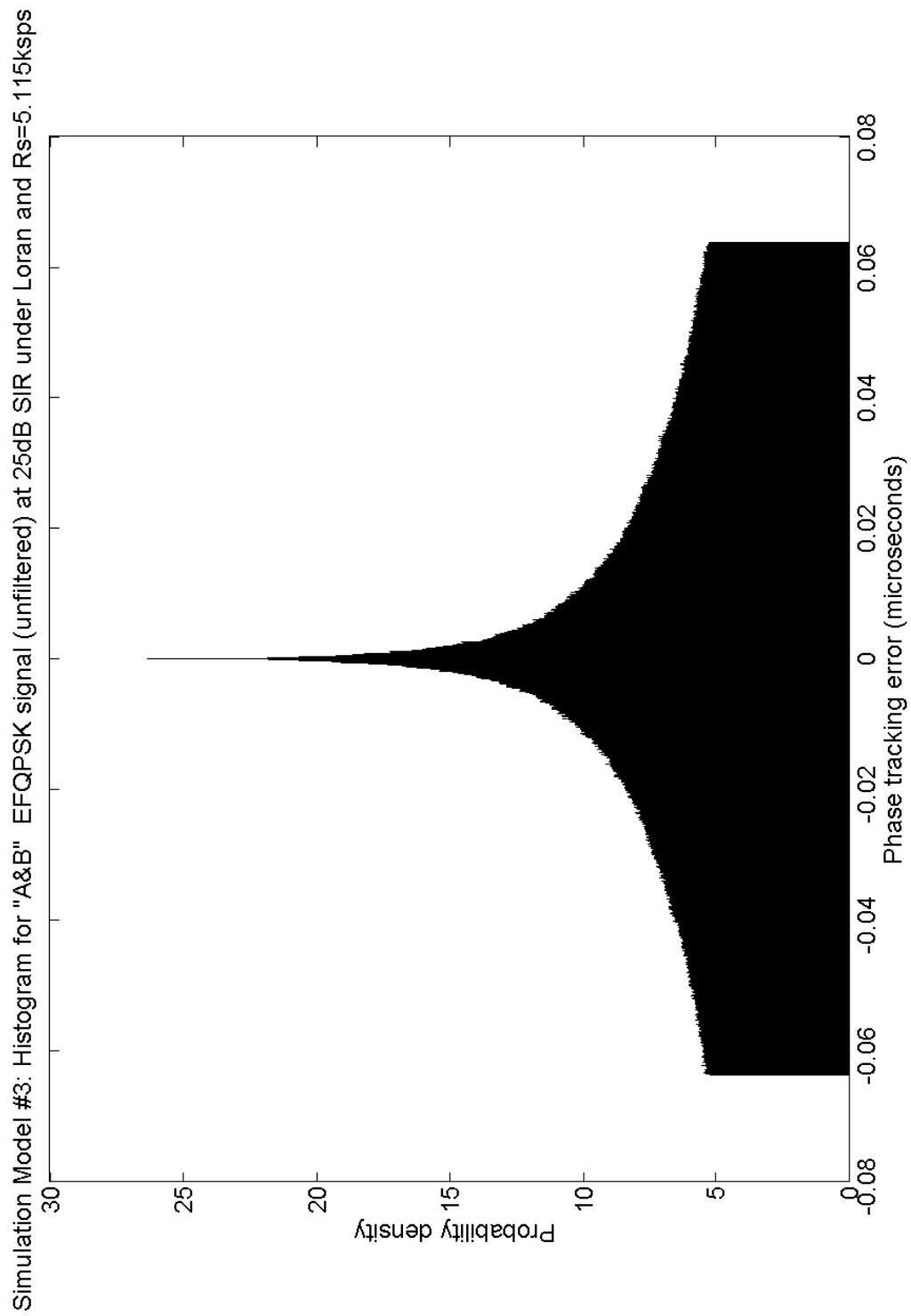


Figure 4.22 – Simulation #2: Unfiltered 'AB' signal at SIR=25dB.

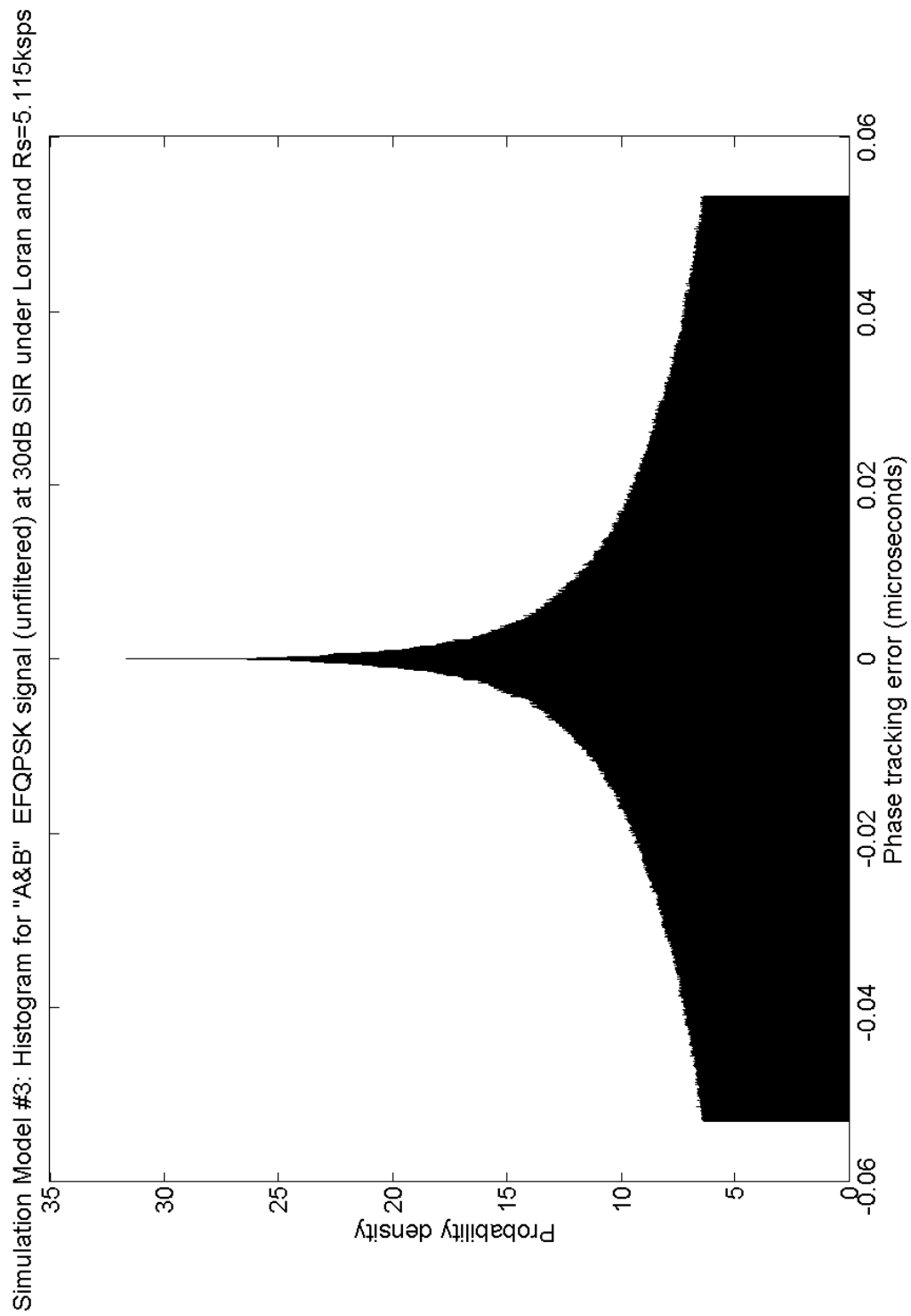


Figure 4.23 – Simulation #2: Unfiltered 'AB' signal at SIR=30dB.

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