Analyzing Spread Spectrum Channel Sounder Performance using Static Channel Measurements

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ABSTRACT

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Performing channel sounding experiments is often considered a first step towards statistical channel modeling. The data obtained for a particular wireless channel environment is then analyzed and the channel is modeled by comparing it to known statistical models. Hence the authenticity of the experimental data is a most critical part in the process of channel modeling. We use channel sounding equipment called Sounder for statistical channel modeling. For synchronizing the transmitter and receiver local oscillators, the sounder manufacturer, Berkeley Varitronics Systems Inc., recommends calibrating the sounder, called “training”. But, there is an uncertainty about the time for which the sounder should be trained before taking measurements for a particular duration of time. After training, the allowed measurement time is estimated from parameters like current count and last count obtained during training, but it is not absolute. How the measurement data varies when the sounder is operated in measurement mode beyond the estimated measurement time is unknown. For this reason, the validity of data obtained during some field measurements is highly uncertain. This thesis work tries to understand the behavior of sounder when operated in measurement mode up to six times the allowed measurement time and succeeds at it. The conclusions drawn from this thesis work are very useful during field measurements to understand the measurement data logged for
longer measurement times when the allowed measurement time is very less. This thesis work will also serve as a guide to the operation and understanding of our sounder for future students.

Approved: _____________________________________________________________

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CHAPTER 1 INTRODUCTION

This chapter describes the underlying concepts on which this thesis work is based on. The nature of a wireless channel and the importance of studying a wireless channel environment to overcome the wireless channel impairments are discussed. Channel sounding using the channel impulse response technique is introduced, which forms the basic working principle of the channel sounding equipment used for measurements in this thesis work. Spread spectrum concepts required to understand the wideband channel and the spread spectrum correlator used in a communication receiver for wideband channel sounding are introduced. The channel characteristics that form the basis of any statistical channel modeling are discussed. Understanding these parameters is important to process the impulse response estimates (IRE) measured with the sounding equipment, and design a channel model from them. Finally the importance of wireless channel modeling is explained through a practical application, the rake receiver.

1.1. Wireless Channels and Channel Characterization

The channel for any type of communication can be wired or wireless. A wired channel, as the name says, is a physical connection between the transmitter and receiver using a cable, wires, waveguide, etc. This physical connection helps enable the transmitter and receiver clocks to be synchronized at all times during communication. The wireless channel has no such physical connection, and this can limit the performance of the overall system. A wired channel is almost always constant and predictable, whereas the wireless channel is often best modeled as random.
In a wireless channel the information is transmitted as an electro-magnetic (EM) wave. For many wireless channels, it is very hard to find a direct path without any obstructions, called the line of sight (LOS) path, between the transmitter and receiver due to the presence of tall buildings, mountains, etc. This problem is more prominent in urban environments where most of the times, the transmitter is not placed high enough compared to the physical obstructions. Due to electromagnetic wave propagation effects from the surface of obstructions, the received signal is both delayed and attenuated. The propagation of an electromagnetic wave between the transmitter and receiver is explained using different physical principles based on the parameters like distance of propagation and time interval considered. Propagation models that concentrate on predicting the received signal strength over large separation distances between the transmitter and receiver are called large-scale propagation models. Other types of propagation models, the small-scale or fading models are those which concentrate on the short time duration effects on the received signal over very small distances [1], corresponding to roughly one-half the signal wavelength. Small-scale fading will be the topic of interest in this thesis.

Small-scale fading effects like random frequency modulation, time dispersion and rapid changes in signal strength during short time intervals and propagation distances are mostly caused by the multipath phenomenon in the radio channel. The direct path, if available, and the different multipath components arriving at the mobile receiver have random phases, amplitudes and angle of arrivals due to reflection, diffraction and scattering caused by the obstructions present in the wireless channel. These multipath
components induce attenuation, delay and possibly inter symbol interference in the received signal. Other factors influencing the small-scale fading are the following: the relative motion between transmitter and receiver resulting in the random frequency modulation effect due to different Doppler effects for different multipath components; the motion of the surrounding objects in a radio channel resulting in random Doppler shifts in multipath components; and the relation between the bandwidth of the channel and bandwidth of the transmitted signal [1]

1.2. Channel Impulse Response (CIR) and Channel Sounding

At any instant of time, the signal received at the receiver is not just the direct line of sight (LOS) path. It is the vectorial sum of the LOS path and different multipath components, each having different phase, amplitude and delay. In a highly obstructive environment, it is also possible that no LOS path is present. Due to the random phases and amplitudes of the multipath components, the received signal is a filtered version of the transmitted signal. Hence, the mobile radio channel can be modeled as a linear filter with time varying impulse response, where the variation in time is due to the fact that there is relative motion between transmitter and receiver (or motion of the scattering objects). This analogy between the channel and the filter enables us to study all the small-scale fading effects from the channel impulse response (CIR - Figure 1.1). Channel characterization/modeling is the study of the channel impulse response and is essential for understanding the small-scale fading effects of a mobile radio channel, which in turn helps us to design a communication system that operates at full capacity [1, 2].
Channel characterization can be done via deterministic and statistical approaches. In a deterministic approach, all the physical properties of the objects present in the channel to be characterized are accurately measured and the electromagnetic propagation equations are used to determine the effects of the channel. In the statistical approach, the channel is approximated to a generalized model and either analysis or experimental data is used to model the channel effects. The statistical approach is often preferred over the deterministic approach because of its analytical simplicity and convenience, as most channel environments are continuously changing and databases required for deterministic analyses are not updated frequently. Channel sounding is an approach for measuring the channel characteristics in the time or frequency domains.

Channel sounding techniques can be broadly classified as narrow band techniques and wide-band techniques. A number of wide band channel sounding techniques have

$$y(t) = x(t) \otimes h(t, \tau) = \int_{-\infty}^{\infty} x(\tau)h(t, \tau)d\tau$$

**Figure 1.1. Channel Impulse Response**
been developed for estimating the fading behavior in a mobile radio channel. According to [1], there are three wide band channel sounding techniques: *direct pulse measurements*, *spread spectrum sliding correlator measurements*, and *swept frequency measurements*. In this thesis work, the equipment used for channel measurements uses a spread spectrum stepped correlator which is based on the spread spectrum sliding correlator.

1.3. Spread Spectrum Communications

Spread spectrum (SS) communications initially found their use in the military applications like anti-jamming, guidance systems and experimental anti-multipath systems [3]. Later on, due to their inherent interference suppression, multipath combative capabilities, along with other advantages, they found applications in commercial radio communications. A definition of spread-spectrum systems is quoted from [4] as follows:

“Spread spectrum is a means of transmission in which the signal occupies a bandwidth in excess of the minimum necessary to send the information; the band spread is accomplished by means of a code which is independent of the data, and a synchronized reception with the code at the receiver is used for despreading and subsequent data recovery.”

A SS signal occupies a bandwidth greater (typically much greater) than required to send the information. Even though the bandwidth availability itself may be a problem,
the spreading of the signal is justified from the fact that more users can be accommodated in the same bandwidth due to the inherent interference reduction capability of SS—this multi-user application of SS is called code division multiple access (CDMA).

There are different types of spread spectrum communications based on the way the spectrum is spread. They are [4]:

1. Direct-sequence spread spectrum (DSSS) – a high-rate pseudo-randomly generated sequence spreads the spectrum by creating phase transitions in the data carrier; this is typically accomplished by a direct multiplication of the spreading and data waveforms.

2. Frequency hopping spread spectrum (FHSS) – pseudorandom frequency shift in the carrier containing data causes the spectrum spread.

3. Time hopping spread spectrum (THSS) – the signal is transmitted in bursts at pseudorandom times.

1.3.1. Maximal Length Sequences (m-sequences) and their Autocorrelations

Most of the material discussed in this section is taken from references [2], [4], and [5]. The DSSS technique uses pseudorandom (PN) codes which spread the carrier containing data signal by causing rapid phase transitions. Figure 1.2 shows a communication system block diagram employing the DSSS technique with PN codes. Figure 1.3 shows a single data bit of duration T seconds and the DSSS “chip” sequence of frequency $f_c$ Hz.
Spreading sequences are desired to have the following properties:

- Each element (i.e., 0’s and 1’s, or -1’s and +1’s) of the sequence occurs with equal probability.

Figure 1.2. A spread spectrum communication system block diagram

Figure 1.3. Plots showing a single data bit of duration T sec (above) and a chip sequence of frequency $f_c$.
• The run length distribution should be similar to the coin-flipping case, i.e., a run length of \( n \) occurs approximately \( 1/2^n \) of the time

• The autocorrelation of the sequence has only one large peak and all other values are significantly below this peak, which leads to small self interference due to multipath

• Minimal multiple access interference is achieved by making the cross correlations between different spreading sequences small at all values of delay

There are different types of spreading sequences like m-sequences, Gold sequences, Kasami sequences, to name a few. Out of these, m-sequences are one of the well known spreading sequences. M-sequences are generated by using linear feedback shift registers (LFSRs). Like their name, the maximal length sequences are the longest sequences obtained using an LFSR of a given length. Using a shift register of length \( m \) will give a sequence of length \( 2^m - 1 \). Figure 1.4 shows an m-sequence generator employing the LFSRs.

![Figure 1.4. PN sequence generator employing m stages of LFSRs](image)

The popularity of m-sequences is because of the following properties that mimic true randomness:
- Each $m$-sequence of length $N = 2^m - 1$ has $2^{m-1}$ ones and $2^{m-1} - 1$ zeros
- Adding an $m$-sequence with its delayed version term by term using modulo-2 addition gives another $m$-sequence
- The full period autocorrelation function of an $m$-sequence is

$$\phi(n) = \begin{cases} 
1, & n = lN \\
- \frac{1}{N}, & n \neq lN
\end{cases}$$

$$l = \frac{\text{delay(\tau)}}{\text{chip duration}(T_c)}$$

A typical auto-correlation function of a spreading waveform obtained from a rectangular chip shaping function $h_c(t) = u_{T_c}(t) = 1$ for $0 \leq t \leq T_c$ and zero otherwise, is shown in Figure 1.5.

![Figure 1.5. Plot showing the autocorrelation function of an m-sequence with a chip period of $T_c$](image-url)
The \( m \)-sequences also have some disadvantages, including that for a given LFSR length there are relatively few \( m \)-sequences, and even for a longer \( m \)-sequence, the full period cross-correlation values are very large for the worst case.

1.4. Spread Spectrum Sliding Correlator

The spread spectrum stepped correlator receiver was introduced first by Donald C. Cox in [6]. Figure 1.6 shows the block diagram of a communication system incorporating the basic idea of a sliding correlator receiver. We use the numerical values of [6], but the principle can be generalized to other values.

![Block diagram of a SS communication system with a sliding correlator receiver](image)

**Figure 1.6. Block diagram of a SS communication system with a sliding correlator receiver**

In the transmitter, a maximal length pseudorandom sequence \( s(t) \) is produced at a rate of 10 MHz, which is used for spreading the carrier with the data signal. At the receiver, for de-spreading, a similar spreading sequence \( s'(t) \) is produced but with a frequency of 2-KHz less than that of \( s(t) \). It is reset according to a standard reference.
signal every 75 msec. At the receiver, the signal received, \( \sum a_i s(t - \tau_i) \cos(2\pi ft - \phi_i) \) is a sum of different delayed versions of the transmitted signal varying in amplitude and phase. The received signal is split in a quadrature hybrid and each output is applied to a correlator.

The “slow” spreading signal is multiplied with a sinusoidal signal in each correlator and the wave form produced, \( s'(t) \cos(2\pi ft) \) is multiplied with the signals from the quadrature hybrid in the correlator producing \( \sum a_i s(t - \tau_i) s'(t) \cos(2\pi ft - \phi_i) \cos(2\pi ft) \) in the in-phase branch and \( \sum a_i s(t - \tau_i) s'(t) \sin(2\pi ft - \phi_i) \cos(2\pi ft) \) in the quadrature branch. It is assumed that the transmitter and receiver have the same phase reference. These signals are finally applied to integrating filters to produce \( I(t) = \sum a_i R_x(\tau_i) \cos \phi_i \) and \( Q(t) = \sum a_i R_y(\tau_i) \sin \phi_i \) which have a peak values depending on the phase at times equal to the delays of different multipath signals. Because the slow spreading signal is reset every time after a fixed time interval, it is possible to calculate the exact delay of different multipath signals.

1.5. Mobile Radio Channel Characteristics

To come up with a solution, we need to understand the problem first. This applies to any communication system. To improve the quality of any communication system, we should first understand the channel parameters and channel effects on the transmitted signal. Using this knowledge (which is often statistical), can we suggest a new
modulation technique or change the transmitted power, design remedial measures such as equalizers or deploy more antennas for diversity reception. Herein lies the importance of channel characterization. As mentioned, statistical modeling of the channel is one of the two ways to characterize a channel.

Link budget and time dispersion is considered two important wireless channel issues for any communication system [7]. The link budget provides a basic quantification of the required transmitter power, coverage, and etcetera to compute the received signal to noise ratio (SNR). The time dispersion of a transmitted signal is caused because of the multipath propagation of the channel. By understanding the time dispersion properties of a channel, the maximum data rate that can be achieved without employing equalizers can be calculated. The following section describes various parameters of a propagation channel.

1.5.1. Path loss (PL)

Interference from other RF sources has more effect on a mobile radio channel when compared to other noise effects. Hence, knowing about interference from other sources is as important as knowing the coverage to a particular mobile user [7]. PL gives a measure of the average attenuation suffered by a transmitted signal in traversing the channel to the receiver. The equation for PL is [8]:

$$PL(dB) = 10 \log \frac{P_r}{P_t}$$  \hspace{1cm} (1.2)

where $P_t$ is the transmitted power and according to Friis free space equation, the power received by a receiver at a distance $d$ from the transmitter is
\[ P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 dL}. \]  

(1.3)

\( G_t \) and \( G_r \) are the transmitting and receiving antenna gains, respectively, \( \lambda \) is the wavelength in meters, and \( L \) is the system loss factor.

For practical terrestrial cases, the mobile radio channel cannot usually be approximated as free space. A more generalized model using a parameter \( n \) is often used, and this also quantifies PL as a function of distance [7].

\[ PL(d) = PL(d_0) + 10n \log(d/d_0) + X_\sigma \]

(1.4)

For wireless terrestrial non-LOS channels, \( n>2 \) (free space), \( d_0 \) is a reference distance and \( PL(d_0) \) is a known value. The parameter \( X_\sigma \) is a usually a zero mean Gaussian random variable which explains the variation of average received power. The accuracy of the modeled PL is estimated from the standard deviation of \( X_\sigma \).

1.5.2. Multipath Time-Delay Spread

Most of the concepts discussed in this section are taken from [1], [7] and [9].

1.5.2.1. Power Delay Profile (PDP)

The power delay profiles are the primary output of wideband channel sounding techniques. They are used to measure many multipath parameters of a wireless channel. PDPs are the spatial average of \( |h(t)|^2 \) over a local area. Many local area PDPs are measured and the output is plotted against excess delay with a fixed time-delay reference. One such plot is shown in Figure 1.7.
1.5.2.2. **First-Arrival Delay (τₐ)**

This time delay is the delay after which the first transmitted signal arrives at the receiver. It serves as the reference point for calculation of other delays. Any delay greater than first-arrival delay is called excess delay.

1.5.2.3. **Mean Excess Delay (τₑ)**

Mean excess delay is the first central moment of a PDP (thus the PDP is treated as a “density” of power vs. delay). It gives a measure of the average excess delay produced by the channel. It is given by
\[ \tau_e = \int (\tau - \tau_A) P(\tau) d\tau = \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)} \]

(1.5)

where the summation formula applies when the multipath components are treated as discrete in delay.

1.5.2.4. \textit{RMS Delay Spread} (\(\tau_{RMS}\))

RMS delay spread is the standard deviation about the mean excess delay and the square root of the second central moment of a PDP. It is widely used as a parameter that can be used to estimate the effect of dispersion on the performance of the digital receiver. It can also provide an estimate of the maximum data rate achieved for transmission through the channel, given some complexity constraints. RMS delay spread is calculated using

\[ \tau_{RMS} = \left[ \frac{\sum_k P(\tau_k) \tau_k^2}{\sum_k P(\tau_k)} - \tau_e \right]^{1/2} \]

(1.6)

1.5.2.5. \textit{Maximum Excess Delay} (\(\tau_m\))

Maximum excess delay is a measure of the excess delay after which the signal power falls below a particular value and considered as noise. This threshold is relative to the maximum signal power and is often from 20 to 30 dB below the PDP maximum.

1.5.3. \textit{Coherence Bandwidth} (\(B_C\))

Coherence bandwidth is a quantity often expressed in relation to the RMS delay spread. If two frequency components of a transmitted signal have high correlation after
passing through the channel, the channels’ coherence bandwidth is said to be at least the absolute value of the frequency difference. When the separation between two frequencies is greater than $B_c$, the channel effects on those components is different compared to those inside the coherence bandwidth. Coherence bandwidth is the band of frequencies over which the channel effects are almost similar in terms of gain (and sometimes phase). If the allowed signal variation is 10%, i.e. the correlation function is above 0.9, and then the coherence bandwidth can be approximated by

$$B_c \approx \frac{1}{50 \tau_{\text{RMS}}}$$

(1.7)

The coherence bandwidth calculated with respect to RMS delay spread is only an approximation, and accurate values can only be obtained when measured as a function of specific channel impulse responses or transfer functions (the transfer function is the Fourier transform of the CIR).

### 1.6 Rake Receiver

In a CDMA based communication system, multiple users share the same band of frequency for communication. For this kind of system, there are broadly two types of detectors in the receivers [2]: conventional and multiuser detectors. A multiuser detector detects the signals from all the users simultaneously using co-channel demodulation principles, whereas the conventional detectors use the autocorrelation property of the spreading signals to detect only one signal, and treats all other signals as additional noise. (Note that nearly all multiuser receivers also use correlator, as their “front end.”) The rake receiver uses multiple conventional detectors.
One of the important statistical parameters obtained after channel characterization is the number of taps ($L$), roughly equal to the number of resolvable multipath components. This can be approximated as $L \approx \left\lceil \frac{T}{T_M} \right\rceil + 1$, with $T$ the symbol (or chip) duration and $T_M$ the duration of the CIR (often $\tau_{rms}$). The multipath phenomenon of the channel under analysis is explained using a tapped delay line model. A basic tapped delay line model of a frequency selective channel with a tap spacing of $1/\text{bandwidth of baseband signal}$ ($W \approx 1/T$) and tap weight coefficients $\{c_n(t)\}$ is shown in Figure 1.8 (10). The number of taps in a tapped delay line model and the weighting coefficients of each tap describe the number of multipath components present and their contribution of each multipath component to the overall received signal. The tapped delay line is a linear, finite impulse response (FIR) model for the channel.

\[
\eta(t) = \sum_{k=1}^{L} c_k(t) s(t-kW) + n(t)
\]

**Figure 1.8. A basic tapped delay line model showing the tap spacing and the tap weight coefficients**

The multipath components, along with the LOS path, contain useful information about how the transmitted signal is affected by the channel. Hence, by combining these time delayed versions of the transmitted signal at the receiver we can improve signal-to-
noise ratio (SNR). Also, this provides diversity at the receiver during decision making. The rake receiver is based on this principle. The received signal with all the multipath components is received by the rake receiver and all components falling under a particular delay window are provided with a separate correlation receiver programmed to that range of time delay. The outputs from each correlator are weighted to reproduce a better estimate of the transmitted signal at the receiver. When the multipath components delay separation is large, because of the autocorrelation properties of the spreading sequence, the components are nearly uncorrelated. Hence rake receiver’s performance is very good in outdoor environments with widely separated delays, whereas in to indoor channels since the multipath delay spreads are much smaller the performance may not be as good [1] (this can of course be remediated by bandwidth expansion to resolve the more closely-spaced delays, if permissible). A three branch rake receiver implementation is shown in Figure 1.9.

![Figure 1.9. A Rake receiver implemented using three correlators corresponding to three different ranges of multipath delays](image-url)
The weighting coefficients $\alpha_k$ are decided based on the correlator outputs $Z_k$. If the correlation output is high, it will be assigned high weighting factor. The correlation output is obtained over some number of samples, and when the correlator is aligned with the multipath component, the output is proportional to the SNR of that component. Also, the weighting coefficients are often normalized to make their summation unity.

\[
\alpha_k = \frac{Z_k^2}{\sum_{k=1}^{N} Z_k^2}
\]

(1.8)

where $N$ is the number of correlators.

1.7. Thesis Objective

Performing channel sounding experiments is often considered a first step towards statistical channel modeling. The data obtained for a particular wireless channel environment is then analyzed and the channel is modeled by comparing it to known statistical models. Hence the authenticity of the experimental data is a most critical part in the process of channel modeling. We conducted channel sounding experiments using equipment called Channel Sounder. The sounder consists of a transmitter and a receiver having independent local oscillators (LO) of 10 MHz each, which are used to generate all other clock frequencies required for operation of the sounder. Any small frequency error at the local oscillators will lead to a significant error in the synthesized high frequencies. Hence synchronizing the transmitter and receiver LOs is very important. The manufacturer of our sounder, Berkeley Varitronics Systems (BVS), calls the process of
synchronizing the transmitter and receiver oscillators “training,” and this is the same as a calibration before actual measurements.

During training, as will be discussed in later chapters, the time for which the LOs stay synchronized is measured in terms of two parameters: current count (CC) and last count (LC). When the training is started, the current count keeps track of the time during which the frequency error between the LOs is improved. Whenever the oscillators go out of synchronization (called a slip), the CC value is stored in LC and the CC starts from zero again. After training the sounder for a particular amount of time, the estimated allowed measurement time (or, the amount of time over which measurements can be considered reliable in terms of synchronization) is calculated as the maximum of CC and LC (which we call as Training Based Measurement time, $T_{TM}$). We assume that the data obtained during this measurement time is void of any errors.

The factors influencing the occurrence of slips during training are highly uncertain, and hence the slip itself is uncertain. This uncertainty is due to randomness in the output of the Rubidium clocks used as the master clocks in the transmitter and receiver. So, after training for any period of time, it is possible that we will get a $T_{TM}$ of, for example, less than 10 minutes. This has been a problem during many of our field measurements, when we train the sounder for more than a day and before taking the measurements, the count will still be a small value, often less than 10 minutes. A reasonable $T_{TM}$ for taking channel measurements is one hour. If we continue taking measurements beyond the assumed $T_{TM}$, we may not be certain of the error in the measured data, and how the error varies with measurement time.
This thesis work tries to address these problems during field measurements and the uncertainties with the allowed measurement time, by using real-time measurement data and simulations. We employ a static (time-invariant) multipath channel with known (and measured) characteristics to conduct experiments that will enable users to more accurately assess reliable sounder measurement time and performance.
CHAPTER 2 SPREAD SPECTRUM STEPPED CORRELATOR SIMULATION

2.1. Spread Spectrum Stepped Correlator Structure

In the spread spectrum sliding correlator, discussed in chapter 1, the spreading code for despreading in the receiver is identical to that used for spreading at the transmitter except for the fact that the frequency of the despreading code is lower than that of the spreading code by a small value, and it is reset periodically with respect to a standard frequency. This principle is used in finding the multipath components in the received signal. The channel sounding equipment used in this thesis (the channel sounder) also uses a spread spectrum correlator similar in principle to the sliding correlator, but with a few differences in operation. It is called a spread spectrum *stepped* correlator. Figure 2.1 shows a block diagram of a spread spectrum communication system using a stepped correlator in the receiver.

![Figure 2.1. Block diagram showing the functional blocks in a spread spectrum communication system using a stepped correlator receiver.](image-url)
As shown in Figure 2.1, the spreading sequence in the receiver is a time shifted version of the original spreading sequence used in the transmitter, running at an identical frequency to that of the transmitter. The time shift is an integral multiple of a single chip duration $T_c$. Every time a time shifted version of the spreading sequence is correlated with the received signal, the multipath component having delay equal to the time shift of the spreading code produces a peak output. Hence the minimum multipath delay resolvable with this type of correlator is $T_c$ and the maximum is $LT_c$, where $L$ is the length of the spreading sequence used. For example, in our channel sounder, the spreading code length is $L=255$ and the chip rate is 50 Mcps. Hence the minimum and maximum resolvable delays in this case are 20 ns and 5.1 μs, respectively.

2.2. Two Path Channel Model

To understand and simulate the working of a spread spectrum stepped correlator, the channel should produce at least one multipath component. Hence, in the experiments and simulations, the channel was designed to produce a LOS path with negligible delay $\tau_0$ and a single delayed path with fixed delay $\tau_1$. The construction of the two-path, time-invariant channel model is shown in Figure 2.2.
In this section, we investigate the basic operating principles and effects of a channel sounder using MATLAB simulations. In these simulations, we use the concept of the two path channel for our study. The transmitted data is a single bit, i.e. +1 of long duration (i.e., no data modulation). Before transmitting the data onto the channel, the data is binary phase shift keying (BPSK) modulated and spread using the $m$-sequence of length 255. The receiver with a stepped correlator is also simulated. The block diagram showing all the blocks in the communication system simulated is shown in Figure 2.3.

**Figure 2.2. Block diagram showing the two path channel with a direct path of very small delay and a delayed path of a known fixed delay.**
In Figure 2.3, the term $\phi_i$ represents the phase offset caused by the $i$th multipath component in the channel, and the term $\tau_i$ represents the delay induced in by the $i$th multipath component in the channel.

2.3.1. Effects of Channel Phase and Delay

When a +1 is to be transmitted over a channel, the signal transmitted from the transmitter antenna is given by:

$$S(t) = s(t) \cos(2\pi f_c t)$$

(2.1)
where $s(t)$ is the spreading signal generated from the $m$-sequence and the cosine term is the carrier waveform. The channel impulse response for the two path channel used in this simulation is given by

$$h(t) = 0.866\delta(t - \tau_0) + 0.5\delta(t - \tau_1)$$  \hspace{1cm} (2.2)$$

where $\tau_0$ and $\tau_1$ are the channel induced delays in the direct and delayed paths, respectively. In this case, $\tau_0$ is 50 chips duration and $\tau_1$ is 55 chips duration. Hence the relative delay between the two paths is 5 chips duration. The received signal at the receiver antenna is given by the convolution of the transmitted signal and the channel impulse response, given by

$$R(t) = [0.866s(t - \tau_0) + 0.5s(t - \tau_1)]\cos(2\pi f_c t + \phi)$$  \hspace{1cm} (2.3)$$

Here we assume that the phase offset caused by the channel in the direct path is identical to that in the delayed path. After performing the despreading and demodulation operations in the receiver, the in-phase ($I$) and quadrature ($Q$) branch outputs are given by

$$I(t) = 0.866R_{xx}(\tau_0)\cos \phi + 0.5R_{xx}(\tau_1)\cos \phi$$  \hspace{1cm} (2.4)$$

$$Q(t) = 0.866R_{xx}(\tau_0)\sin \phi + 0.5R_{xx}(\tau_1)\sin \phi$$  \hspace{1cm} (2.5)$$
where $R_{xx}(\tau) = \frac{1}{T} \int_0^T s(t)s(t-\tau)dt$ is the normalized periodic autocorrelation of the spreading signal and $T$ is the time period of the sequence. Figure 2.4 shows I and Q branch outputs as a function of the chip positions for the case when there is no phase offset in the received signal due to channel. From Figure 2.4, we can see that there are two peaks representing the direct path and the delayed path, and their positions on the abscissa represent the induced delay from the channel. Figure 2.5 shows I and Q branch outputs when there is a phase offset of 10 degrees in the received signal.

**Figure 2.4.** I and Q branch outputs plotted versus the chip positions for a zero phase offset channel.
When the phase offset created by the channel is 0 degrees, the Q branch output is zero because of the orthogonality of the cosine and sine terms. Hence, in Figure 2.4, the Q branch output is zero, but in Figure 2.5 for the 20 degree phase shift case, we can notice small peaks corresponding to the two path positions.

2.3.2. **Effect of Analog Frequency Offset**

Assuming that there is no phase offset in the channel, Figure 2.6 shows the functional block diagram of the channel sounder incorporating an analog frequency offset at the carrier. The offset is given by $\delta f = f_c - f_r$. 

**Figure 2.5.** I and Q branch outputs plotted versus the chip positions for a 20 degree phase offset channel.
For the two-path channel, the signal at the receiver is given by

\[ r(t) = \left[ 0.866s(t - \tau_0) + 0.5s(t - \tau_1) \right] \cos(2\pi f_c t) \]  

(2.6)

If the carrier signal frequency at the transmitter is \( f_c \), and at the receiver it is \( f_r \), then the fractional frequency offset, \( \Delta f \) is given by:

\[ \Delta f = \frac{f_c - f_r}{f_c} = \frac{f_r - f_c}{f_c} \]  

(2.7)

The output of the correlator, \( I \), is given by:

\[
I(t) = 0.866R_{xx}(\tau_0) \frac{1}{T'} \int_0^{T'} \cos(2\pi f_c (\Delta f) t) \, dt + 0.5R_{xx}(\tau_1) \frac{1}{T'} \int_0^{T'} \cos(2\pi f_c (\Delta f) t) \, dt 
\]

(2.8)
where \( R_{xx}(\tau) = \frac{1}{T} \int_{0}^{T} s(t)s(t - \tau) \, d\tau \) is the normalized periodic autocorrelation of the spreading signal and \( T \) is the time period of the sequence. The correlator output consists of two peaks as shown in Figure 2.4, corresponding to the two paths of the modeled two-path channel, but with an amplitude modulation superimposed. This amplitude modulation results from the sinusoidal terms in (2.8). Figure 2.7 shows the variation of the correlator output with time for both the paths, for both simulation and analysis, for a frequency offset value of \( 10^{-5} \). The results from the simulation are compared with the calculated value using (2.8). As can be observed, agreement between simulations and analysis is essentially perfect.

**Figure 2.7.** Two-path channel correlator output variation with time – simulated vs. calculated, incorporating frequency offset.
In Figure 2.7, the carrier signal at the transmitter is generated at five carrier cycles per symbol transmitted, and the frequency offset is fixed at $\Delta f=10^{-5}$. Hence, the receiver carrier signal frequency, $f_r$, is $f_c(1+10^{-5})$. 
CHAPTER 3 CHANNEL SOUNDING EXPERIMENTS

The Berkeley Varitronics Systems (BVS) channel characterization equipment (the “sounder”), as the name suggests, is used to characterize the wireless channel for indoor and some outdoor environments. One significant challenge with any wireless communication system is to keep the transmitter and receiver synchronized all the time. Any loss of synchronization between the sounder’s transmitter and receiver will result in erroneous channel measurements, and hence channel effects cannot be studied accurately. The sounder has a transmitter and a receiver; both equipped with a rubidium (Rb) oscillator, which are in essence atomic clocks. The main drawback of Rb oscillators is that, they have short-term stability, which means that the oscillator’s frequency/phase characteristics vary over time. Hence, the manufacturer recommends operating the device in “Training Mode” before taking measurements. During this mode, the transmitter and receiver are connected back to back using a coaxial cable. This process is continued until the desired frequency stability of the oscillators is achieved.

3.1. Setup for Experiments

A linear time-invariant channel has been constructed, with a direct path and a fixed delayed path with a delay of approximately 80 ns. For the delayed path, we use a long coaxial cable (65 feet RG 8/u cable) and fixed attenuators (3 dB and 10 dB). To keep the power level of the delayed path signal at least 10 dB below that of the direct path, a 40 dB attenuator is used in the direct path. The experiment setup used for training the sounder is shown in Figure 3.1.
The transmitter and receiver should be connected back to back for training. Our experiment setup for training, shown in Figure 3.1, is different. The modified setup for training the sounder is used to minimize the number (and time) of connection changes between the training and measurement stages. The setup used for measurements is shown in Figure 3.2.

**Table 3.1. RF Equipment**

<table>
<thead>
<tr>
<th>Equipment</th>
<th>Details</th>
</tr>
</thead>
<tbody>
<tr>
<td>Delay cable</td>
<td>RG 8/u 65 feet</td>
</tr>
<tr>
<td>Splitter/Combiner</td>
<td>3 way 50 ohms</td>
</tr>
<tr>
<td>Attenuators</td>
<td>One 40 dB, One 10 dB, One 3 dB</td>
</tr>
<tr>
<td>Connecting cables</td>
<td>Three &lt;4 feet</td>
</tr>
</tbody>
</table>
The main drawback of using an open connection in the setup is that, due to impedance mismatch at the open end, power is reflected back from the open circuit, which will affect the direct path signal. Yet, because of losses in the cable, the delayed path attenuators, and the splitter and combiner, the reflected power will be below the direct path by a minimum of 100 dB. Hence we assume that this training setup is close to the recommended setup.

3.2. Terminology used in the Experiments

Before proceeding further into this thesis, it is important to understand the definitions of the terms used in the experiments.

1. Training time ($T_T$): Time duration for which the sounder is operated for training purpose before taking any measurements.

2. Current count (CC): When the sounder is operated in training mode, the time duration for which there has been no loss in synchronization (sync) between the
transmitter and receiver oscillators. This value can be read directly from the Raptor software interface used to control the operation and record the sounder measurements.

3. Last count (LC): When the sounder is operated in training mode, this is the value of the CC when there was a most recent loss in sync between the transmitter and receiver oscillators. This value can also be read from the Raptor software interface.

4. Training-based measurement time ($T_{TM}$): After the sounder has been trained for the required accuracy of frequency offset between the transmitter and receiver oscillators, this is an estimate of the possible measurement time achievable with that accuracy. It is calculated as the maximum of “current count” and “last count”.

5. Measurement Time ($T_M$): After the sounder has been trained, this is the time duration for which the sounder is used for taking measurements. This time can be measured using a stopwatch manually.

6. Data Logging Time ($T_{DL}$): During measurements, it is the time duration for which the measurement data from the sounder has been stored into a file using the Raptor software interface. $T_{DL}$ is always less than $T_M$. This time should also be measured using a stopwatch manually.

7. Transition Time ($T_{TR}$): After the sounder has been trained, the intervals of time during which the measurement data is not stored in a file are called transition time intervals. Transition time is the duration of each transition time interval. Such time intervals occur when:
a. changing the experiment setup between training and measurement modes of operation,

b. saving the current logged data file and creating a new file to continue logging data (This repeats for every 30 minutes of measurement time).

The total transition time duration can be measured as the difference between measurement time and data logging time.

3.3. Procedure for Experiments

This section gives a detailed description of the procedure followed for operating the sounder during training and measurements. The selection of modes of operation, observation of the correlator output and data logging is done using the “Raptor” software provided along with the sounder. The sounder communicates with the software on a PC or laptop through a serial port via an RS-232 interface.

3.3.1. Training Mode

The following steps are used during training of the sounder:

1. Setup the required equipment as in Figure 3.1 and power ON (XMIT OFF) the transmitter and receiver.

2. After the rubidium oscillators warm-up (indicated on the displays), set the following parameters on the transmitter,

   - Transmitter Power +5/ +10 dBm (as per requirement)
   - Center Frequency 5.12 GHz
   - Chip Code Rate 50 Mega chips per second
3. Start RF transmission on transmitter by pressing the “XMIT” button and then invoke the Raptor software on the laptop.

4. Select the “Setup” tab on the Raptor software and setup the center frequency and chip rate for the receiver, identical to that of the transmitter as shown in Figure3.3.

![Image](image.png)

**Figure 3.3.** Selecting the settings for the receiver before the training process is started

5. If the received power is more than -3 dB, the software shows, “Raptor Stable” and “Raptor Locked” as shown in Figure 3.4b. Now, begin training by clicking on the “Start Training” button in the software.
6. The Raptor software displays the values of last count, current count, accuracy, peak position and the time elapsed from the instant training started as shown in Figure 3.5a.

Figure 3.4. Figure showing the unstable (a) and stable (b) state of the raptor depending on the received power.

Figure 3.5. Figure showing the data displayed on the raptor software during training (a) and the final step when training is stopped (b).
3.3.2. Measurement Mode

After training the sounder for the required training based measurement time, which is calculated as the maximum of the current count and last count, the following steps are followed:

1. Stop the training process by selecting “Stop training” and “Accept Value” options on the Raptor software, shown in Figure 3.5b. At the same time, invoke a stopwatch to start measuring the measurement time for that trial.
2. Now select the measurement tab on the Raptor software and set the same operating frequency and chip rate as shown in Figure 3.6.

![Image](image.png)

**Figure 3.6.** Selecting the desired settings on the receiver before the process of taking measurements has started.
3. Before changing the experiment setup from training (Figure 3.1) to measurement mode (Figure 3.2), stop RF transmission on the transmitter by pressing the **XMIT** button. Once the measurement mode setup is finished, restart the RF transmission.

4. The receiver output is a power delay profile (PDP). Each set of PDPs can have a different number of samples as per the required delay span, which we set before logging data. By selecting the maximum delay span of 5.1 micro seconds using the **"FULL SPAN"** option, we get 510 samples per PDP. Because the transmitter and receiver are both stationary throughout the experiment, using the **"SET ZERO"** option, and selecting distance between transmitter and receiver as zero feet, we can get a maximum of 1020 samples per PDP, of which, the second 510 samples are a replica of the first 510 samples. Select Full Span or Set Zero as per the requirement.

Figure 3.7. Figure showing the power delay profiles generated for two settings available in raptor software – (a) Full Span (510 point PDP) (b) Set Zero (1020 point PDP).
5. Next press the record button and use a filename to save the logged data in a file. Once you press ok, the data from the receiver correlator is logged as a “.rap” file with the filename selected.

![Figure showing the receiver output data is being logged (Log), and the recorded file size is displayed on the screen.](image)

Figure 3.8. Figure showing the receiver output data is being logged (Log), and the recorded file size is displayed on the screen.

6. Once the training is stopped in step 1, the measurement time starts on the stop watch, but the data logging is only started in step 5. This time interval between the end of training and the beginning of data logging, which we regard as the transition time, is kept constant at approximately 90 seconds.

7. After the data logging is started, the data is logged into a new file after every successive 30 minutes is elapsed on the stopwatch. This requires stopping the data logging and restarting it with a new file name. This is also a kind of transition time and is kept constant at approximately 10 seconds.
This division of one complete measurement trial of duration $T_M$ into sub-trials of 30 minutes each and transition times is shown in Figure 3.9.

![Diagram](image)

**Figure 3.9. Example of a single trial of $T_M$=1 hour divided into 2 sub-trials of 30 minutes each, and transition times of 1min 30 sec and 10 sec.**

3.4. Processing of Measurement Data for Analysis

The logged data obtained from the sounder receiver is stored as a data file with a proprietary ‘.rap’ extension. This file is in compressed format, and needs to be converted into a more manageable file format like the ‘.out’ ASCII format. For this purpose, the sounder manufacturer, Berkeley Varitronics Systems, has provided software called ‘Chameleon (Raptor edition)’. In the Chameleon software, there is an option to choose which data fields to be extracted from the original ‘.rap’ file and placed into the ‘.out’ file. All the options available in the chameleon software are shown in Figure 3.10.

Even though the format of the logged data file is changed, a manageable output file size is also equally important for further analysis. Based on the required $T_{TM}$ (5 min to 30 min), the data logging times range from 30 minutes to more than 3 hours. The output file for such long measurement times will be very large. Importing such large files for analysis into mathematical software like MATLAB, which is the primary tool used for data analysis in this thesis, is not practical. Hence, the total measurement time is
divided into sub-trials of 30 minutes and each file is saved separately. These files are small enough for further analysis using MATLAB.

![Chameleon Data Converter](image)

**Figure 3.10. Screenshot of the chameleon software, for converting the logged receiver data into a convenient ASCII file.**

For a required $T_{TM}$ of 30 minutes, the measurements were taken for 3 hours (i.e. $6*T_{TM}$). If the whole measurement data during this trial were to be saved as a single file (`.rap`), the size of the file could be as large as 100 MB. Then, using the Chameleon software for extracting all the required parameters into a single file (`.out`) makes the file as large as several GB. Hence we make sure that the experiment data is not continuously logged for more than 30 minutes of measurement time. In this case, the single measurement trial of 3 hours measurement time is divided into 6 sub-trials of 30 minutes.
each for data logging. This is achieved by stopping the data logging after every 30 minutes of measurement time and saving the new log file with a different file name. Also, only one parameter (i.e. magnitude, phase) is extracted at a time from the ‘.rap’ file into the ‘.out’ file. This process gives ‘.rap’ files of less than 20MB and ‘.out’ file sizes of less than 25MB for each extracted parameter.

The magnitude and phase vectors extracted from the logged data may be of length 1020 or 510 depending on the setting (SET ZERO and FULL SPAN) used for that particular measurement trial. For the former case of 1020 samples, the second 510 samples are discarded from each PDP. The receiver sampling rate is 100 MHz, which is double the transmitter sampling rate. Hence the data output of the receiver is processed by performing vectorial addition of two adjacent samples in a power delay profile to match the transmitter sampling rate and chip duration resolution of 20 ns.

3.5 Observations Made from Experiments

Before starting our experiments, the main purpose was to analyze the output data recorded during measurements as a function of measurement time, while keeping the training based measurement time constant for a set of measurements (trials). We expected to see a degradation in performance (in the form of amplitude or frequency or phase changes) in the measurement data (power delay profiles) when the measurement time exceeded the training based measurement time. With the help of these experiments, we planned to develop a model for the error incorporated in the measurements when the
sounder was used in prolonged measurements for a time beyond the training based measurement time. Figure 3.11 explains the basic idea behind the initial measurements.

The experiments were conducted in different phases- a starting phase, an analyzing phase and a testing phase. Each phase, as the names suggest, was specifically designed for a particular set of goals. The observations made from each phase of the experiments were used to improve the test plan for the successive phases.
3.5.1. Starting Phase of Experiments

3.5.1.1. Objective

In the starting phase of the experiments, the main interest was to get an approximate understanding of the sounder output when operated beyond the recommended $T_{TM}$ and then make plans for future experiments. Hence this phase consisted of experiments with small $T_{TM}$ values of 5 minutes and 10 minutes. Each experiment trial consisted of training and measurement modes of operation as discussed in the previous sections. To maintain the $T_{TM}$ constant for each trial, we stopped the sounder training after the count had reached the required value of $T_{TM}$ for that trial. For example, during the 5 minutes $T_{TM}$ trials, the training of the sounder was performed until the $\text{max}(\text{current count, last count})$ equaled 300 (5*60 sec).

For a $T_{TM}$ of 5 min, five experiment trials were conducted and for a $T_{TM}$ of 10 min two experiment trials were conducted. During each trial, the measurement time was a minimum of five times the training based measurement time.

Note: The main difficulty faced in achieving the same $T_{TM}$ for different trials is that the total training time ($T_T$) will not be constant to obtain a constant $T_{TM}$. For example, during the 5 minutes $T_{TM}$ trials, the $T_T$ ranges from 5 minutes to as long as 5 hours for different trials; this is due to the random nature of the rubidium clock variations. Hence, the operator needs to be very cautious and patient to notice when the count reaches the required $T_{TM}$. The situation can even be tedious for the 30 minutes $T_{TM}$ trials. Hence we
decided that allowing a 10% difference in the count is reasonable, i.e., to obtain $T_{TM}$ of 30 minutes, the max(current count, last count) can range from 1600 to 2000 seconds.

3.5.1.2. Observations

For the analysis of the recorded PDPs, the magnitude of the PDPs was plotted against the measurement time. Some of the results for a $T_{TM}$ of 5 min are shown in Figure 3.12 and those for a $T_{TM}$ of 10 min are shown in Figure 3.13.

Figure 3.12. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for experiment trials with $T_{TM}$ of 5 minutes

Figure 3.13. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for experiment trials with $T_{TM}$ of 10 minutes
The results obtained in this phase of experiments were very interesting. In the PDP magnitude plots, the amplitude fluctuated between a minimum and maximum periodically because of the slip in the autocorrelation peak in the sounder output, and/or because of an offset in the frequency of the Tx and Rx clocks. The minimum values in the magnitude represent the slip in the position of the autocorrelation peak by one position (single chip duration). Also, we expect that there will be no slip in the autocorrelation peak for minimum duration of time equal to the training based measurement time. Yet that was not the case in most of the experiment trials. Also, the period between two consecutive minima in the analysis plots is not the same for all the trials for a particular $T_{TM}$, which is evident in Figure 3.12.

This unexpected behavior of the sounder was believed to be because of

1. $T_{TM}$ of 5 minutes was very small to be considered for any performance estimates and only two trials of $T_{TM} = 10$ min were conducted, i.e., this is insufficient data to come to any conclusions.

2. Most of the trials were conducted back to back keeping the sounder powered all the time. This might change the operating conditions for different trials of the same set, that is, conditions of the clocks may change after being “on” for a long time.
3.5.2. Analysis Phase of Experiments

3.5.2.1. Objective

Based on the observations made from the starting phase of data collections, new objectives were designed for the analysis phase of experiments. They were to observe the effect of conducting two experiment trials back to back without powering off the equipment and with powering off the equipment, and to observe the variations in the PDP parameters for different trials conducted with the same $T_{TM}$. In the analysis phase, the experiments were conducted and the data was collected for training based measurement times of 10 minutes and 30 minutes. For a $T_{TM}$ of 10 minutes, five experiment trials were conducted and for a $T_{TM}$ of 30 minutes, four trials were conducted.

3.5.2.2. Observations

The procedure used in the starting phase was repeated with the data collected in the analysis phase to analyze the data. Example plots of the PDP magnitudes versus measurement time for 10 minutes and 30 minutes training based measurement times are shown in Figures 3.14 and 3.15 respectively.
As it is evident from the above figures, even with the extra care taken not to conduct the experiments back to back, there is not much change in the pattern of the output with time, except for the periodic correlation peak slips. Yet in addition, this time, two problems were found with the measured data. They are as follows:

**Figure 3.14. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for experiment trials with $T_{TM}$ of 10 minutes**

**Figure 3.15. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for experiment trials with $T_{TM}$ of 30 minutes**
1. From Figure 3.14b, we can see that there are sudden power gain “spikes” of almost 60 dB in the output, even though the shape of the output is retained.

2. From Figure 3.14 and Figure 3.15b, we can see that the plot of PDP magnitudes are not continuous i.e., some PDPs are not recorded in the output.

Additional experiments were conducted mainly to quantify these problems and report them to the sounder manufacturer, Berkeley Varitronics Systems, for starting the repair process.

3.5.3. Testing Phase of Experiments

3.5.3.1. Objective

By this time, we were confident that even for relatively large training based measurement times of 30 minutes, there was not a significant change in the sounder output for modeling the errors. The main objective for this phase of experiments was to test the sounder for the observed errors in the analysis phase. Hence, maintaining a constant $T_{TM}$ was not judged to be an appropriate approach, so the sounder was trained for more than a day’s time for each trial irrespective of the training based measurement time and the subsequent measurements were taken for more than 3 hours for each trial. In this way, we believed that we could be confident that the errors observed were not just because the sounder or laptop was not trained long enough. Also, for verifying the error regarding the missing PDPs was not the problem of the software or laptop used for data logging, some experiments in this phase were conducted using a different laptop.
3.5.3.2. **Observations**

The outputs recorded in these trials of experiments were analyzed using the same procedure as in the previous trials and some of the plots from this trial are shown in Figure 3.16 and Figure 3.17.

![Figure 3.16. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for one of the experiment trials in testing phase. Also illustrated is the sudden power gain problem](image)

Figure 3.16. Analysis plots showing the plot of magnitude of PDPs recorded versus measurement time for one of the experiment trials in testing phase. Also illustrated is the sudden power gain problem.
As it is evident from the above figures, the errors discussed in the above section were still present even for very long training times. These results were discussed in a detailed report sent to the sounder manufacturer and the sounder itself was sent back to the manufacturer for repair. The manufacturer finally concluded that these errors were caused by some problems in the Raptor software interface used to record the data. For a detailed explanation of the errors observed, see Appendix A and B, where the reports sent to BVS are reproduced.

Table 3.1 summarizes the classifications of the three phases of experiments based on the settings used and the way in which they were conducted.
### Table 3.2. Classification of different data collection phases based on the settings used

<table>
<thead>
<tr>
<th>No.</th>
<th>Setting</th>
<th>Phase I</th>
<th>Phase II</th>
<th>Phase III</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Warm-up time before first trial of the day</td>
<td>No</td>
<td>&gt; 1 hour</td>
<td>No</td>
</tr>
<tr>
<td>2</td>
<td>Back to back trials of same $T_{TM}$</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>3</td>
<td>Full Span (510 point) or Set Zero (1020 point)</td>
<td>Set Zero</td>
<td>Full Span</td>
<td>Set Zero/ Full Span</td>
</tr>
<tr>
<td>4</td>
<td>Total training time ($T_T$) recorded</td>
<td>No</td>
<td>Yes</td>
<td>Yes</td>
</tr>
<tr>
<td>5</td>
<td>Power OFF between two trials</td>
<td>Yes</td>
<td>No</td>
<td>Yes</td>
</tr>
<tr>
<td>6</td>
<td>Transmitter Power output</td>
<td>+5dBm</td>
<td>+10dBm</td>
<td>+10dBm</td>
</tr>
<tr>
<td>7</td>
<td>Last count recorded</td>
<td>No</td>
<td>Yes</td>
<td>Yes</td>
</tr>
</tbody>
</table>
CHAPTER 4  ANALYSIS OF OPERATION OF THE SOUNDER FROM EXPERIMENTS

In this chapter, we will focus on explaining the term accuracy and the cause for the correlation peak slips in the sounder output.

4.1. Quantifying the accuracy of the measured data

The term “accuracy” is defined as [11]

\[
\text{Accuracy}(A) = \frac{\text{one chip duration}(T_c)}{\max(\text{LC, CC})}
\]  

(4.1)

For our experiments, a single chip duration equals 20 ns, since the chip rate is 50 Mcps. The local oscillators (LOs) in the transmitter and receiver are the rubidium clocks operating at approximately 10 MHz each. All the other frequencies required for generating the spreading code and the carrier waveforms are synthesized from this LO. Hence, any small error in frequency of the LOs in transmitter and receiver leads to potentially large errors in the final outputs. The process of training the channel sounder before taking the measurements, as explained in chapter 3, is done to make the frequencies of the LOs in the transmitter and receiver as close as possible to each other to avoid such errors in the outputs.

The term accuracy, as defined in (4.1), gives a measure of the offset in the frequencies of the transmitter and receiver LOs. For example, if \( A = 6.67 \times 10^{-11} \), then \( \max(\text{CC,LC}) = 300 \) seconds. This means that, if the sounder is trained for an accuracy of \( 6.67 \times 10^{-11} \), there is a high probability that there will be a slip (which we define as a drift in the correlation peak position on the delay scale) of \( T_c \) (i.e., 20 ns) in the correlator.
output peak position after 300 seconds. However, this measure of accuracy is only an estimate and the actual value can only be obtained using the measured data. Consider a trial from dataset 1 with a training based measurement time of approximately 10 min. The plot of the correlation peak magnitude versus measurement time is shown in Figure 4.1. For this trial, the estimated accuracy was calculated as $3.33 \times 10^{-11}$, and the actual accuracy is calculated from:

$$A_{cal} = \frac{20ns}{T_D}$$

(4.2)

where $T_D$ is the duration between two consecutive direct path correlation peaks. Calculation of $T_D$ is shown in Figure 4.2.

Figure 4.1. Plot of magnitude of PDPS recorded versus measurement time for one of the experiment trials in testing phase.
Hence, for this case, the actual accuracy is $2.50 \times 10^{-11}$ (i.e., 20ns/800). Table 4.1 shows the estimated accuracy measured using equation 4.1 and the actual accuracy measured from the measured data for all the measurement trials.

Table 4.1. Estimated and measured accuracy values for all the measurement trials classified based on the training based measurement times

<table>
<thead>
<tr>
<th>Dataset</th>
<th>Trial</th>
<th>$T_T$</th>
<th>$T_D$</th>
<th>$A_{th}$</th>
<th>$A_{meas}$</th>
<th>$%e = 100 \times (A_{th} - A_{meas})/A_{th}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>NA</td>
<td>300</td>
<td>6.67E-11</td>
<td>6.67E-11</td>
<td>0.00</td>
</tr>
<tr>
<td>1</td>
<td>2</td>
<td>NA</td>
<td>1600</td>
<td>6.67E-11</td>
<td>1.25E-11</td>
<td>81.26</td>
</tr>
<tr>
<td>1</td>
<td>3</td>
<td>NA</td>
<td>1600</td>
<td>6.67E-11</td>
<td>1.25E-11</td>
<td>81.26</td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>NA</td>
<td>1700</td>
<td>6.67E-11</td>
<td>1.18E-11</td>
<td>82.31</td>
</tr>
<tr>
<td>1</td>
<td>5</td>
<td>NA</td>
<td>400</td>
<td>6.67E-11</td>
<td>5.00E-11</td>
<td>25.04</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>6</td>
<td>NA</td>
<td>1000</td>
<td>3.33E-11</td>
<td>2.00E-11</td>
<td>39.94</td>
</tr>
<tr>
<td>1</td>
<td>7</td>
<td>NA</td>
<td>800</td>
<td>3.33E-11</td>
<td>2.50E-11</td>
<td>24.92</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>0:41:00</td>
<td>1500</td>
<td>2.78E-11</td>
<td>1.33E-11</td>
<td>52.16</td>
</tr>
<tr>
<td>2</td>
<td>2</td>
<td>0:10:00</td>
<td>600</td>
<td>3.33E-11</td>
<td>3.33E-11</td>
<td>0.00</td>
</tr>
</tbody>
</table>

$T_{TM} = \sim 5 \text{ min}$

$T_{TM} = \sim 10 \text{ min}$

Figure 4.2. Magnified view of the plot in Figure 4.1 showing duration between two correlation peak minima.
From Table 4.1, we can see that the actual measured value of accuracy is often smaller than the estimated value. This is because the total training time is always greater than or equal to the training based measurement time for all trials and the values of the count we use for calculating the estimated accuracy are only the value obtained when the training is stopped.

4.2. Quantifying the cause for the correlation peak amplitude modulation

Although we are using a static channel with fixed delay and attenuation, the amplitude of the correlator output peaks corresponding to the direct and delayed path are not constant over measured time as can be seen in Figure 4.1 above (and all the analysis...
plots in Chapter 3). For both direct and delayed paths, the peak amplitude follows a sinusoidal pattern, with minimum values at regular intervals. The cause of this modulation is the frequency difference between the LOs in the transmitter and receiver, causing the spreading codes and the carrier signals to be generated at different rates. Hence the duration of a single chip in the spreading code and single carrier cycle is longer for a lower frequency oscillator compared to that for a higher frequency oscillator.

In the two path channel case, we get two peaks in each PDP corresponding to the direct path and delayed path, positioned with respect to their individual delays on the time scale. Figure 4.3 shows the variation of the amplitude and time delay of correlation peaks in a power delay profile at different time instances. From the figure, we can say that there is amplitude modulation in the correlation peak amplitude. This can be explained by the well known effect of carrier frequency offset. This effect was explained earlier in the simulated results. The other explanation will be with respect to the different spreading code generation rates.
In Figure 4.3, at time $t_1$, $\tau_1 - \tau_0 (\Delta t_1) = \Delta t_1$ and at time $t_2$, $\Delta t_2 = \Delta t_2$. The minimum delay that can be calculated using the BVS sounder is 20 ns, which is the chip duration. Hence the change in the delay ($\Delta t$) can be characterized only for integral multiples of 20 ns. This change in the delay by 20 ns is called a slip in the correlation peak. The correlation peak magnitude will be highest when the transmitter and receiver spreading codes are exactly aligned with each other and minimum otherwise. Hence, the maxima in the analysis plots denote that the transmitter and receiver spreading sequences are mostly aligned with each other and the minima represent the output when the spreading
sequences are least aligned. Figure 4.4 provides a visual understanding of amplitude modulation observed during a particular trial of $T_{TM} = 5$ minutes.

![Figure 4.4](image)

**Figure 4.4. Variation of the direct path amplitude over measurement time for a measurement trial.**

measurement time along with the amplitude modulation. The maxima and minima from Figure 4.4 are used to plot Figure 4.5. The surface plot in the figure shows the drifting of the direct and delayed paths over measurement time, and the mesh plot shows the amplitude variation.
Figure 4.5. Surface plot (a) and mesh plot (b) of the PDPs corresponding to the maxima and minima in Figure 4.4.
CHAPTER 5  SUMMARY, CONCLUSIONS AND FUTURE WORK

5.1. Summary and Conclusions

The main objective of this thesis work was to quantify the uncertainties regarding the allowed measurement time and the resulting error in the measured data when operated beyond the previously conceived measurement time based on the LC and CC obtained from training. From the analysis in chapter 4, we can say that we were able to achieve that objective. This thesis serves as a guide to the operation and understanding of our sounder for future students.

During each measurement trial, we made sure that the measurement time was five to six times the estimated measurement time (Training based measurement time, $T_{TM}$) obtained during training. Hence, we can say that the allowable measurement time is “well beyond” the $T_{TM}$ for without significant distortion of the logged data. The channel used for the experiments was a static two-path channel; hence any observed effects in the logged data can only be attributed to the channel sounding equipment, i.e., the sounder. A total of 19 measurement trials (71 sub-trials) were conducted with a total training time of 9622 minutes (or ~160 hours) and a total measurement time of 2130 minutes (or ~35 hours). Such large amounts of measurement data add to the confidence of our analysis results and conclusions.

As discussed previously (section 4.2), the position of the correlation peaks corresponding to the direct path and multipath components drifts periodically on a time delay scale by a single chip duration, i.e., 20 nsec in this case. For the time delay analysis of multipath components, we always consider the relative delay between the direct path
and the multipath components as their delays. For example, if the direct path correlation peak is positioned at 50 nsec on the delay scale and a multipath component is at 100 nsec, the direct path is taken as the reference and the multipath is said to be delayed by 50 nsec from the direct path. From our analysis, during our measurement time, the correlation peak slips are all synchronized, i.e. the two correlation peaks corresponding to the direct and delayed paths drift simultaneously. So, although there will be a change in the absolute delays, the relative delays remain constant throughout the measurement time (i.e., up to six times the training based measurement time). Hence, if amplitude modulation is synchronized for all multipath components, relative delays and relative amplitudes are constant, which means that delay spread measures are not affected by this. This is even true for time-varying channels, and if rate of change of channel is rapid compared with sounder “slipping,” measurement data is still accurate.

The term accuracy, as defined in section 4.1, gives an estimate of the fractional frequency offset ($\Delta f$) between the transmitter and receiver local oscillators based on the maximum of current and last counts after training. From the analysis of our experimental data, we can safely assume that, for a minimum training time of 12 hours, the actual accuracy measured from the measurement data is always better than the estimated accuracy. Hence, the estimated accuracy can be considered as the worst case of the actual accuracy obtained. The reason for this can be given from the fact that the training time of the sounder is always greater than or equal to the maximum of the current and last counts.

Consider the case when we get a very low value for training based measurement time even after training for long training times, because of the uncertainty in the loss in
synchronization between the oscillators during training. During some field measurements (in airports, government buildings, private properties, etc.), we will be obligated to take measurements on a particular day, no matter how small our training based measurement time. In such a case, we would ask, how accurate are the measurements? This can be answered by observing the amplitude modulation phenomenon discussed in section 4.2. For all of the measurement trials conducted, the correlation peak amplitude for direct and delayed path varies within 3 to 5 dB. Hence, we can be sure that for a measurement time of at least 6 times the allowed measurement time, the signal amplitude will only vary within 3 to 5 dB between maximum and minimum. So, depending on the application, if this amount of error margin is allowed, one can be confident about the measurement data correctness.

5.2. Future Work

As per the information provided by the sounder manufacturer Berkeley Varitronics Systems (BVS), the primary oscillators in the transmitter and receiver of the sounder are of 10 MHz each. This is multiplied up to 100 MHz to provide the sampling frequency in the transmitter and receiver units. Hence a small difference in frequencies between the transmitter and receiver primary oscillators will be magnified by 10 times for the sampling clocks. Hence, the carrier frequency and the spreading code sampling rate can differ between the transmitter and receiver. This thesis work addressed the effect of the frequency offset in the receiver carrier signal using simulations and measurement trials, assuming that the transmitter and receiver generated spreading codes are sampled
at the same frequency. For future work, simulating the effect of a difference in spreading code sampling rates in addition to the frequency offset in the receiver carrier frequency would be of interest. Additional simulated results can then be compared with the measurement data to explain more accurately the amplitude modulation effect produced on the correlator output.
REFERENCES


systematic measurement errors of correlative mobile radio channel sounders."


APPENDIX A ANALYSIS OF ERRORS DETECTED IN OUR SOUNDER

In this section, we try to include most of the material documented for the purpose of explaining to our sounder manufacturer, Berkeley Varitronics Systems, the abnormal behavior of the BVS channel characterization equipment (the “sounder”), observed during experiments conducted in the Multi-user Mobile Communications Lab at Ohio University. The main observations made were the following:

1. Variation in the update period of Power Delay Profiles (PDPs), and
2. Spurious power level “gain spikes” of more than 60 dB in PDPs at arbitrary times during measurements.

Several sounder parameter values we employ during our experiments are as follows:

Transmitter Power output +10 dBm
Center Frequency 5.12 GHz
Received Power -36 dBm (as measured on the Raptor software during measurements)
Chip Code Rate 50 Mcps
Output PDP length 1020 samples (We obtain this using the ‘Set Zero’ feature. Note that only 510 samples are unique, and the second set of 510 samples is a repetition of the first, which we discard when processing. If we used the “full span” button on the software, we would obtain only 510 samples, and it is the “set zero” feature that actually yields this 1020 sample output.)
Update Period Problem

For the results shown in Table A.1, the sounder was operated in the training mode for 25 hours and 3 minutes ($T_T$). At the end of this training time, the Last Count (LC) was 4,914 seconds and the Current Count was 12,245 seconds. From sounder documentation and prior discussions with BVS, the maximum of the LC and CC can be taken as a rough measure of “reliable measurement time” for channel measurements using the sounder. Hence, in this case, we assumed that the sounder should operate satisfactorily in the measurement mode for 12,245 seconds ($T_{TM}$).

<table>
<thead>
<tr>
<th>Sub Trial</th>
<th>$T_{start}$ (H:M:S)</th>
<th>$T_{stop}$ (H:M:S)</th>
<th>$T_M$ (min)</th>
<th>RTC $T_{start}$ (H:M:S)</th>
<th>RTC $T_{stop}$ (H:M:S)</th>
<th>$T_M$ (min)</th>
<th># Measured PDPs N</th>
<th># Expected PDPs $N'$</th>
<th>$N'-N$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1a</td>
<td>0:02:30</td>
<td>0:35:00</td>
<td>32</td>
<td>17:22:12</td>
<td>17:54:42</td>
<td>32.5</td>
<td>4644</td>
<td>4653.7</td>
<td>9.87</td>
</tr>
<tr>
<td>1b</td>
<td>0:35:10</td>
<td>1:01:00</td>
<td>24</td>
<td>17:54:51</td>
<td>18:20:41</td>
<td>25.8</td>
<td>2586</td>
<td>3694.57</td>
<td>1108 .457</td>
</tr>
<tr>
<td>1c</td>
<td>1:01:05</td>
<td>1:30:00</td>
<td>29</td>
<td>18:20:47</td>
<td>18:49:41</td>
<td>28.9</td>
<td>2896</td>
<td>4138.364</td>
<td>1242 .364</td>
</tr>
<tr>
<td>1d</td>
<td>1:30:10</td>
<td>2:00:00</td>
<td>30</td>
<td>18:49:52</td>
<td>19:19:41</td>
<td>29.8</td>
<td>4238</td>
<td>4267.241</td>
<td>29.2 408</td>
</tr>
<tr>
<td>1e</td>
<td>2:00:10</td>
<td>2:30:00</td>
<td>30</td>
<td>19:19:52</td>
<td>19:49:41</td>
<td>29.8</td>
<td>4294</td>
<td>4267.241</td>
<td>18.2 408</td>
</tr>
<tr>
<td>1f</td>
<td>2:30:10</td>
<td>3:00:00</td>
<td>30</td>
<td>19:49:52</td>
<td>20:19:41</td>
<td>29.8</td>
<td>2979</td>
<td>4267.241</td>
<td>1288 .241</td>
</tr>
<tr>
<td>1g</td>
<td>3:00:10</td>
<td>3:30:00</td>
<td>30</td>
<td>20:19:47</td>
<td>20:49:42</td>
<td>29.9</td>
<td>3011</td>
<td>4281.56</td>
<td>1270 .56</td>
</tr>
</tbody>
</table>

The total measurement time of approximately 3 hr 30 min was divided into 7 sub-trials of times in the column labeled $T_M$; these sub-trials were approximately 30 minutes each. Even though the data has been logged separately into 30 minute files, all the data belongs to a single continuous measurement. A stopwatch was started at the moment
when training was stopped and in the Raptor software we “accepted value.” Times $T_{\text{start}}$ and $T_{\text{stop}}$ are the starting and stopping times of the data logs for each sub-trial. RTC $T_{\text{start}}$ and RTC $T_{\text{stop}}$ are the starting and stopping times of the data logs obtained from the logged data during the sub-trials. The number of PDPs logged ($N$) during each sub-trial is significantly different from the expected number of PDPs ($N'$) which we can calculate using the update period formula, provided by Boris Sheyer, Chief Enginer of BVS (bsheyer@bvsystems.com).

Update Period $(T) = T_o + nT_e$  

(A.1)

where $T_o = 11$ ms, $T_e = 0.4$ ms and $n = 1020$ (#samples per PDP), so $T=0.419$ sec. For example from (A.1), for the first 32.5 minute measurement period in row 1 of Table 1, we would expect to obtain $N'=(32.5\text{minutes} \times 60\text{seconds/minute}) \div (T \text{ seconds})=4653.9$ PDPs, when we actually obtained $N=4644$ PDPs. Thus it appears that the rate at which PDPs are being output is time varying, when we expect it to be constant.

Power Level Gain Problem

Figure A.1 illustrates what we mean by the “gain spikes.” This plot shows two PDPs with power in dBm versus sample number. For our fixed 2-path channel, this figure shows two measured profiles taken very close together in time, plotted on the same graph. One PDP has a power level approximately 65 dB above the “normal” PDP. The gain spikes are also shown in Figure A.2, which plots the two path amplitudes vs. time.
In Figure A.1, we can clearly see the direct path and delayed path peaks approximately 100nanosec apart (the direct paths have the “x,y” cursor values listed). In Figure A.2, these path amplitudes from a sequence of output PDPs are plotted as a function of time. We can see many gain spikes are evident, yielding many “erroneous” PDPs. (Even though the “erroneous” PDP shapes appear fine, this gain spiking does appear to be a problem.) We have tested the sounder transmitter by itself using a spectrum analyzer, and we did NOT see any such gain spikes, so we believe that this problem lies in the receiver.
Note the blank spaces in Figure A.2 around 2000sec, 4000sec, 9000sec and 11000sec, which correspond to sub-trials b, c, f and g of Table A.1, respectively. These gaps do not mean that there are no PDPs measured during this time. When we plotted this graph, the update rate was assumed to be 2.4 PDPs per second, its normal value for full delay span operation. This value of 2.4 PDPs/sec is obtained using (A.1), i.e., the update rate is $1/T$. From Table A.1, the number of generated PDPs during these sub-trials (b, c, f and g) are significantly smaller than the expected number of PDPs calculated for the update rate of 2.4 PDPs per second. Since the time scale is assumed constant in Figure A.2, all the PDPs generated during these trials are compressed together. What is really happening of course is that PDPs are distributed over their respective time frames corresponding to varying PDP output (update) rates during those time periods. Figure
A.3 illustrates this variation in PDP output rate conceptually (this is not measured data, but only an illustration to help clarify the time variation of PDP output rate).

Figure A.3. Conceptual illustration of observed time variation in PDP output rate.
In Appendix A, the update period problem and the gain spike problem were discussed on the context of a single measurement trial. In this section, we analyze the update period problem in a general sense by considering all the measurement trials conducted. A total of 59 measurement trials were conducted before the sounder was repaired. During these measurements, maximum care was taken to record the time of the start and stop of each measurement trial, to accurately calculate the number of PDPs generated during that time period; using the Update Period \( T = T_0 + nT_e \). The confidence in measurement time accuracy is based on the fact that it is in agreement with the RTC time log obtained from the measured data.

From the total 59 measurement trials, 39 trials were conducted using the “FULL SPAN” setting, to generate 510 samples per PDP (Update rate = 4.6512 PDPs per sec) and 20 trials were conducted using the “SET ZERO” setting, to generate 1020 samples per PDP (Update rate = 2.3866 PDPs per sec). Data from all these trials are shown separately in Tables B.1 and B.2 for SETZERO and FULLSPAN settings, respectively. These tables provide an illustration of the magnitude of the error in the number of PDPs generated, based on the calculated number of PDPs using the \( T = T_0 + nT_e \). For example, assume a measurement trial of exactly 30 minutes of logged data performed using FULL SPAN setting and SET ZERO setting separately.

For FULL SPAN setting (510 samples per PDP), we have the following update period and rate:

\[
T = 11\text{msec} + 510 \times 0.4\text{msec} = 215\text{msec}
\]
Update rate \((R) = 1/T = 4.6512\) PDPs per sec

Then we obtain the following for the number of (expected) PDP’s generated

\[(30\times 60)\text{sec} \times R = 1800\text{sec} \times 4.6512\ \text{PDPs/sec} \approx 8372 \ \text{PDPs}\]

Similarly for SET ZERO setting (with 1020 samples per PDP), we obtain

\[T = 11\text{msec} + 1020\times 0.4\text{msec} = 419\text{msec}\]

\[R = 1/T = 2.3866 \ \text{PDPs per sec}\]

and the expected number of PDPs is \((30\times 60)\text{sec} \times R = 1800\text{sec} \times 2.3866\ \text{PDPs/sec} \approx 4296\]

**Table B.1**

*Table showing the data related to the 20 trials for SETZERO setting*

<table>
<thead>
<tr>
<th>Trial starts T_start</th>
<th>Trial stops T_stop</th>
<th>Total time T_M (sec)</th>
<th>Expected no. of PDPs</th>
<th>Measured no. of PDPs</th>
<th>Error in the no. of PDPs</th>
</tr>
</thead>
<tbody>
<tr>
<td>0:01:40</td>
<td>0:31:00</td>
<td>1760</td>
<td>4200.00</td>
<td>4207</td>
<td>-7.00</td>
</tr>
<tr>
<td>0:01:33</td>
<td>0:30:00</td>
<td>1707</td>
<td>4074.00</td>
<td>4078</td>
<td>-4.00</td>
</tr>
<tr>
<td>0:01:30</td>
<td>0:30:00</td>
<td>1710</td>
<td>4081.00</td>
<td>4062</td>
<td>19.00</td>
</tr>
<tr>
<td>0:01:30</td>
<td>0:30:00</td>
<td>1710</td>
<td>4081.00</td>
<td>4075</td>
<td>6.00</td>
</tr>
<tr>
<td>0:01:30</td>
<td>0:30:00</td>
<td>1710</td>
<td>4081.00</td>
<td>4076</td>
<td>5.00</td>
</tr>
<tr>
<td>0:01:30</td>
<td>0:30:00</td>
<td>1710</td>
<td>4081.00</td>
<td>4050</td>
<td>31.00</td>
</tr>
<tr>
<td>0:30:06</td>
<td>1:04:00</td>
<td>2034</td>
<td>4854.00</td>
<td>5139</td>
<td>-285.00</td>
</tr>
<tr>
<td>0:01:30</td>
<td>0:30:00</td>
<td>1710</td>
<td>4081.00</td>
<td>4063</td>
<td>18.00</td>
</tr>
<tr>
<td>0:30:06</td>
<td>1:06:00</td>
<td>2154</td>
<td>5141.00</td>
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From Figure B.1, it is clear that almost 50% of the trials have an error in the number of PDPs generated compared with the expected number of PDPs.

Figure B.1. Plot and Histogram of the error in the number of PDPs, which is the difference between the expected number of PDPs and the measured PDPs; for SETZERO setting.

Table B.2

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Figure B.2 also shows that almost 50% of the trials have an error in the number of PDPs generated.
Figure B.2. Plot and Histogram of the error in the number of PDPs, which is the difference between the expected number of PDPs and the measured PDPs; for FULLSPAN setting.