DESIGN AND DEVELOPMENT OF A LOW-COST HIGH RANGE RESOLUTION X-BAND RADAR

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DESIGN AND DEVELOPMENT OF A LOW-COST HIGH RANGE RESOLUTION X-BAND RADAR

by

Paul Cantu

A THESIS

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DESIGN AND DEVELOPMENT OF A LOW-COST HIGH RANGE RESOLUTION X-BAND RADAR

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University of Nebraska, 2003

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Synthetic Aperture Radar (SAR) is one of the main tools for microwave remote sensing because of its multi-dimensional high resolution characteristics and the capability to operate in nearly all weather conditions, day and night. The University of Nebraska-Lincoln (UNL) initiated the design and development of a low-cost airborne SAR in January 2001 to support our Airborne Remote Sensing Program. The objectives of this project are separated into various evolutionary stages. This thesis will focus on the initial phase of design and construction of an X-band high range resolution radar (HRR) using basic RF/microwave and digital components. The following stages will expand on the HRR design to achieve a functioning X-band airborne SAR for the remote sensing of underlying vegetation parameters (tree height, leaf area index, biomass content, etc.) from a low altitude platform from a range of 1000 meters. The SAR system is an X-band, stepped-chirp FM, single polarization radar system. One of its unique features is that the signal generation consists of a timing-controlled D/A converter and VCO arrangement to generate the step-chirp signal, thereby allowing for less design complexity and a much lower overall system cost.
The individual block-segments of the SAR include a stepped-chirp FM waveform synthesizer, transmission and receive paths, antennas, quadrature detection and image signal processing. The system underwent rigorous in house laboratory testing and subsequent outdoor field-testing from a van-mounted boom where preliminary HRR one-dimensional images were obtained. It is anticipated that the following progression of development for this HRR system will be to use this design as a basis towards fully coherent, data acquisition from an airborne platform.
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Chapter 1

Introduction

1.1 Motivations

Monitoring the status and condition of the earth's terrain, such as vegetation fields, forests and soil surfaces is of primary importance to the understanding and the protection of the environment, as well as for natural resource management. It has been recognized that the use of microwaves in remote sensing is a suitable approach as microwave signals are able to penetrate cloud and vegetation cover and can operate independent of solar illumination. Moreover, microwaves have the ability to penetrate more deeply into vegetation compared to optical signals. In general, longer wavelengths penetrate vegetation much better than shorter wavelengths. Currently, there exist operational C-band and L-band systems operating in the 5 GHz and 1 GHz frequency bands, respectively, that are routinely used for sea surface and soil moisture remote sensing. However, higher frequencies such as X-band (10 GHz) are used for radar systems where the size of the antenna constitutes a physical limitation; this becomes especially important when considering an airborne platform. The main thrust in developing the X-band Synthetic Aperture Radar (SAR) is to provide a light-weight, low-cost system to ultimately operate from an airborne platform. The fact that the system operates in the X-band microwave range will allow for smaller RF (radio frequency) components and antennas, relative to C-band and L-band, making
it ideal for airborne operations. The system design is one which is easy to maintain and operate, and capable of quick turn-around in data acquisition.

Ultimately, X-band SAR will provide remote sensing of underlying terrain parameters (tree height, soil moisture, biomass content, etc.) from an airborne platform; however, there are several evolutionary steps needed towards accomplishing this goal. This thesis will focus on the design and construction of the X-band high range resolution (HRR) radar, which will be the core hardware design to the airborne SAR system. The HRR radar will be tested and validated using ground-based tests from a mobile boom-mounted van.

1.2 Scope of the Thesis

In this thesis project, a ground-based HRR radar system has been designed, constructed and tested. This system is mounted on a mobile, van boom-arm during the testing phase and validation phases. There are three objectives identified for this project:

i. *To design, construct and test a HRR radar system*

   The system design will be simple and straightforward. Simple, lightweight, compact, and low-cost design are the overall top-level requirements.

ii. *To validate the HRR radar resolution*

   The HRR will produce a ground range resolution better than 4 meters.

iii. *To use this technology as a basis towards the next evolution stage, i.e., graduating this HRR radar system to an airborne SAR platform*

   The experience gained in this project will be used for future evolutions towards an airborne platform SAR system.
This thesis reports on the design, development and testing results of the HRR radar system. Chapter 2 will review radar fundamentals, especially as they relate towards the design of a synthetic aperture radar system using the HRR radar as the system hardware. In particular, the concept of the stepped-chirp signal generation is highlighted.

A detailed description of the system design and its operation is presented in Chapter 3. In general, the construction of the SAR can be subdivided into five parts. These include (a) the signal generation, (b) the radio frequency (RF) section, (c) the antennas, (d) the receiver section and (e) the data acquisition section. The HRR is a quasi-monostatic radar, where a dual antenna system, in close relation to each other, is employed in this design. This design configuration will eliminate the problems associated with transmit/receive isolation, to minimized design complexity and the potential to expand the system for interferometric applications. The RF section is constructed from individual RF components. These include the use of a power splitter, band pass filter and RF amplifier. The main function of the receiver section is to filter and amplify the intermediate frequency signals received from the mixer output. The pre-processed signals are then converted into digital data by the data acquisition section, which includes an analog-to-digital (A/D) computer card, computer software and the computer storage. Analysis and processing of the data can be performed virtually in real-time while still gathering data on the aircraft or in the field.

Several radar cross section measurements have been carried out to verify the operation of the SAR system. These include the measurements of standard targets such as trihedrals of differing sizes at various downrange distances. The analysis, results and observations are presented in Chapter 4.

Finally, Chapter 5 will conclude with a summary of work performed and future recommendations for this X-band high-resolution radar.
Chapter 2
Review of Radar Theory

2.1 Radar Fundamentals

The term "radar" stands for RAdio Dectection And Ranging. This term was coined in the early 1950s, at a time when detection of ships and determination of their ranges were its primary purpose. However, after the proliferation of high speed computing, digital processing and other advances, radar has graduated beyond simple detection and ranging of targets, and is used in a wide range of applications for the military, air traffic control, police departments and remote sensing, to mention a short list.

Generally, remote sensing refers to the process of recording/observing/perceiving (sensing) objects or events at far away (remote) places. In remote sensing, the sensors are not in direct physical contact with the objects or events being observed. The technique requires an electromagnetic wave, of suitable characteristics, to travel from the objects/events to the sensors through the atmospheric medium. The choice of electromagnetic radiation wavelength (or frequency) depends upon the specific remote sensing application.

Synthetic aperture radar is basically a method of ground mapping and imaging that uses computer processing to sharpen the azimuthal resolution over what can be achieved by a conventional antenna. This technique first appeared in the early 1950s, but did not reach a high state of development for almost 30 more years. With the introduction of high-speed digital processing and other advances, SAR is rapidly emerging as a powerful remote sensing tool (Ulaby, et al., 1981).
2.2 The Radar Equation

In microwave remote sensing, the distinction between different ground targets being sensed is primarily caused by the difference in the signal strength received by the radar. Therefore, it is understood the received signal strength is the most important measurement in remote sensing. In the simplest case, the radar equation can be stated verbally as suggested by [Ulaby et.al, 1981]:

\[
\begin{align*}
\text{Power received} &= \left( \text{Power per unit area at target} \right) \times \left( \text{Effective scattering area of the target} \right) \times \\
& \quad \left( \text{Spreading loss of radiated signal} \right) \times \left( \text{Effective antenna receiving area} \right)
\end{align*}
\]

where the power at the receiver is the energy scattered back from the target (i.e. backscatter). The power per unit area at the target is the amount of energy intercepted by the ground target as a result of its physical characteristics. The effective scattering at the target represents the re-radiation of energy in all directions from the point source (ground target), to include that fraction of power received at the receiver. The spreading loss of radiated signal stems from the nature of the radiated power from the point source spreading out in all directions in the area of a sphere of radius \( R \). The further away from the point source, or the greater the radius \( R \), the less power received at the receiving antenna. The actual size of the receiving antenna will also affect the total power received.

The phonetic equation can be used to assist in deriving the fundamental radar equation as follows:

\[
P_r = \frac{P_t G_t A_r \sigma}{(4\pi R^2)^2}
\]

(2.1)
\[ P_r = \text{Power received} \]
\[ P_t = \text{Power transmitted} \]
\[ G_t = \text{Gain of the transmitting antenna} \]
\[ \sigma = \text{Radar cross section} \]
\[ G_t = \text{Gain of the transmitting antenna} \]
\[ G_t = \text{Gain of the transmitting antenna} \]

This equation can be broken down further for clarification:

\[ P_r = P_t \times G_t \times \frac{1}{4\pi R^2} \times \sigma \times \frac{1}{4\pi R_t^2} \times A_r \]

(2.2)

The power received (1) by the radar system \( P_r \), is a function of a linear frequency modulated chirp pulse of electromagnetic energy, \( P_t \) (2), that has been focused down by the transmitting antenna to a beam width, so that the energy level increases by a factor of \( G_t \) (3) over a spherically expanding wave (4). This focused energy illuminates an area on the ground that has a radar cross section (RCS) of (5). The backscattered power then radiates isotropically (or spherically) in all directions from the point scatter (6). The receiving antenna area, \( A_r \) (7), intercepts a fraction of this backscatter signal for processing.

Although the SAR sports a two-antenna design, both antennas are generally identical to each other in all characteristics, including gain, i.e. \( G = G_t = G_r \). As per [Ulaby, et al., 1983], the relationship between the antenna directive gain and receiving aperture area is:

\[ G = \frac{4\pi A_r}{\lambda^2} \]

(2.3)

or equivalently;

\[ A_r = \frac{G\lambda^2}{4\pi} \]

(2.4)
where \( \lambda \) is the wavelength of the radar system. Substituting Equation (2.4) into Equation (2.2) or (2.1) results in a modified radar equation:

\[
P_r = \frac{P_t G^2 \sigma \lambda^2}{(4\pi)^3 R^4}.
\]  

(2.5)

There are several forms of this radar equation; however, one common form equates the ratio of the returned signal power, \( S \) (or \( P_r \)), to the receiver noise power, \( N \). The noise figure of the receiver, \( F_N \), and the thermal noise power, \( kTB \) (\( k \) = Boltzmann constant, \( T \) = Temperature, \( B \) = Bandwidth), as well as the system and environmental losses, \( L \), are integrated into Equation (2.6) to become the radar equation of interest. The Signal-to-Noise (power) ratio at the receiver antenna port is

\[
SNR_{\text{antenna}} = \frac{P_t G^2 \sigma \lambda^2}{(4\pi)^3 R^4 F_N kTB L}.
\]  

(2.6)

2.3 SAR Principles

SAR systems can produce maps (or images) of radar reflectivity versus range and azimuth, typically in the form of a two-dimensional (2-D) image. One dimension in the image is called range and is a measure of the "line-of-sight" distance from the radar to the target. Range measurement and resolution are determined by precisely measuring the time from transmission of a pulse to receiving the echo from a target. Fine range resolution can be accomplished through pulse compression techniques, such as in this design: linear frequency modulation (LFM) chirp.

The other dimension is called azimuth (cross-range) and is perpendicular to range. It is the ability of the SAR to produce a relatively fine azimuth resolution which differentiates it from other radars. To obtain fine azimuth resolution, a physically large antenna is needed to focus the transmitted and received energy into a sharp beam. The sharpness of the beam defines the azimuth resolution. Fine azimuth
resolution is enhanced by taking advantage of the radar motion in order to synthesize a larger antenna aperture, which exploits the Doppler effect. The distance the radar platform travels in synthesizing the antenna length is known as the synthetic aperture. A narrow synthetic beamwidth results from the relatively long synthetic aperture, which yields finer resolution than is possible from a smaller physical antenna.

The quality of SAR images is heavily dependent on the size of the map resolution cell. The resolution cell is the area comprised of the range resolution by the azimuth resolution. In the following subsections, the range and azimuth resolutions are explained and calculated for this design.

Figure 2.1: Footprint Definition.

Figure 2.1 shows the geometry for the standard side looking SAR. As noted, the intersection of the antenna beam with the ground defines a footprint. As the platform moves, the footprint scans a swath on the ground, whose width is determined by the antenna beamwidth, range and incidence angle. The antenna 3 dB beamwidth is \( \theta_{3dB} \), the range from the platform to the target is \( R \), and the incidence angle is \( \theta_i \). Therefore, the size of the footprint painted on the ground is

\[
R \theta_{3dB} \sec \theta_i = \frac{R \theta_{3dB}}{\cos \theta_i} = \frac{R \lambda}{L \cos \theta_i} \text{ (range direction)}
\]
\[ R\theta_{3dB} = \frac{R\lambda}{L_h} \]  

(azimuth direction)

where \( L_v \) is the vertical length and \( L_h \) is the horizontal length of the physical antenna, and \( \lambda \) is the radar wavelength.

### 2.4 Range Resolution

Ground range resolution is defined as the minimum distance on the ground at which two object points can be imaged separately, shown in Figure 2.3, as the distance \( \delta R_g \).

\[
\delta R_g = \frac{ct}{2}
\]  

(2.7)
\[ \delta R_g = \frac{\delta R_s}{\sin \theta_i} = \frac{T}{2 \sin \theta_i} = \frac{c}{2B \sin \theta_i} \]  

(2.8)

where \( \delta R_s \) is the slant resolution, \( \theta_i \) is the incident angle, \( T \) is the pulse length, \( t \) is the pulse width, \( B \) is the frequency bandwidth of the transmitted radar pulse and \( c \) is the speed of light.

The ground range resolution is infinite for a vertical look angle (\( \theta_i = 0^\circ \)) and improves as the look angle is increased. Also, note that the ground range resolution is independent of the height of the radar platform. In order to achieve a fine range resolution, the pulse width must be minimized, which is accomplished via the frequency-modulated pulse (chirp) rather than resorting to using very short physical pulses. This, in effect, will reduce the average transmitted power and increase the operating bandwidth. Achieving fine range resolution while maintaining adequate average transmitted power can be accomplished by using a pulse compression (i.e. modulation) technique as described in Section 2.8.

### 2.5 Azimuth Resolution

The SAR system has a major advantage over the real-aperture side looking airborne radar (SLAR) in that the resolution in the azimuth direction is independent of the radar platform altitude or range distance to the target. SAR systems depend upon two conditions; (1) the motion of the radar platform relative to a stationary target, and (2) signal processing. Both of these conditions are used to synthesize an antenna (aperture) that is much longer in length than the actual antenna hardware. Condition (1) achieves this by taking advantage of the Doppler effect, and condition (2) by achieving a much narrower beamwidth in the along-track direction than that attainable with the real-aperture systems. Therefore, a larger synthetic aperture produces a finer azimuth resolution.
The SAR system transmits the microwave pulses to the imaging area and the echoes are received as “raw” SAR data, which are coherently recorded onto storage. It is these recorded echoes, which generate a coarse (raw) microwave reflectance ‘image’ of the illuminated ground area. The natural dimensions of this image are range and azimuth. Whereas the range resolution of the raw image is dictated by the pulse width (as described in Section 2.4), the ‘raw’ azimuth resolution is determined by the azimuth extent of the antenna pattern footprint on the ground, the antenna size, wavelength and the imaging geometry (SLAR properties). This ‘raw’ data in azimuth, would exhibit an unacceptable image for this dimension. Subsequent image processing of the ‘raw’ SAR data is required to dramatically improve the azimuth resolution by taking advantage of the coherence of successive echo signals as the antenna moves.

This unacceptable “raw” image is a product of the broad antenna pattern, i.e. a SLAR property, which is based only upon the magnitude of the received signal. It is the signal phase histories over azimuth that exhibits an extremely sensitive measure for the instantaneous distance between the radar receiver and the point target. After correlation of the transmitted chirp signal and the receivers echo ‘raw’ data in signal processing, the SAR data are focused to an azimuth resolution on the order of

\[ \delta R' = \frac{L_h}{2} \]  

(2.9)

where \( L_h \) is the length of the antenna. This is the finest potential synthetic azimuth resolution that can be theoretically achieved. However, Equation (2.9) is not the only parameter that has an impact on the SAR data. The coherent nature of the SAR signal will produce speckle in the image. To remove the speckle, the image is processed in digital signal processing (DSP) by averaging several looks. By increasing the looks, the interpretability of the SAR image is significantly enhanced; however this is at the
cost of the azimuth resolution. Therefore, the azimuth resolution Equation (2.9) must be adjusted to include the effects of averaging, i.e.,

\[ \delta R_a = n \times \frac{L_h}{2} \]  

(2.10)

where \( n \) is the number of looks to be averaged. Since it is desired to have a virtual square resolution cell, the nominal range resolution can be adjusted to meet this goal by selecting an appropriate value for \( n \).

### 2.6 Doppler Bandwidth

Tracking the Doppler shift history of a point target as it is illuminated by the radar provides the information necessary to resolve the azimuth location of the target. The Doppler shift history is obtained by comparing the reflected signals from a ground target with a reference signal that incorporates the same frequency of the transmitted pulse. The output is known as a phase history, and it contains a record of the Doppler frequency changes plus the amplitude of the returns from each ground target as it passed through the beamwidth of the antenna. Therefore, the Doppler bandwidth can be viewed as being constrained by the antenna beamwidth as

\[ \Delta f_D \approx \frac{2v_{rel} \theta_{az}}{\lambda} \]  

(2.11)

and since \( \theta = \frac{\lambda}{L_{az}} \), Equation (2.11) becomes

\[ \Delta f_D \approx \frac{2v_{rel}}{L_{az}} \]  

(2.12)

\( \Delta f_D = \) Doppler bandwidth

\( \lambda = \) Transmitted wavelength

\( \theta_{az} = \) Antenna beamwidth in the azimuth dimension

\( v_{rel} = \) Relative radial velocity of the radar platform to the target

\( L_{az} = \) Length of antenna along the azimuth dimension
A further detailed derivation for the Doppler bandwidth is illustrated with Figure 2.3.

![Diagram](image)

Figure 2.3: Doppler spread geometry of a synthetic aperture

The Doppler frequencies received from ground-target return signals will increase as the radar platform moves from position 1, where there is a maximum Doppler shift of $-\delta f_d$, towards position 2, where the Doppler frequency shift will become zero. As the radar platform passes through position 2, the Doppler frequency shift will continue towards $+\delta f_d$ as the radar platform approaches position 3. It is the frequency range between positions 1 to 3, where the maximum Doppler shifts occur or when the target first enters and exits the radar beam width, respectively, which determines the Doppler bandwidth, $\Delta f_D$:

$$
\delta f_D = 2|\Delta f_d| = \frac{2v_s}{c} f_s = \frac{2v_{rel} sin\phi}{c} \frac{\lambda}{L_z} = \frac{2v_{rel} L_z}{R\lambda} \tag{2.13}
$$

Substituting $L_z = \frac{R\theta_{az}}{2}$ and $\theta_{az} = \frac{\lambda}{L_{az}}$ in Equation (2.13) leads to

$$
\Delta f_D = 2\frac{v_{rel}}{2L_{az}} \frac{R\lambda}{L_{az}} = \frac{v_{rel}}{L_{az}} \tag{2.14}
$$
where

\[ L_s = \text{Synthetic aperture length of beamwidth} \]
\[ R = \text{Distance from radar platform to target at bore sight} \]
\[ c = \text{Speed of wave propagation} \]
\[ f_s = \text{Operating frequency} \]

### 2.7 Ambiguities and Pulse Repetition Frequency (PRF)

Ambiguities can exist with pulse radars for both range and azimuth measurements. It is the pulse repetition frequency (PRF) that controls the extent of overlap of adjacent radar echos, accounting for the ambiguities. Generally, the radar maximum PRF must be low enough to avoid range ambiguity and the minimum PRF must be high enough to avoid Doppler ambiguity.

There is an upper limit on the PRF imposed by geometry (swath width). The PRF will need to be lower than the maximum bandwidth of the transmitted pulse (\( \Delta W \)) to prevent overlapping of the near-range and far-range return pulses, hence range ambiguity. The maximum unambiguous range \( (R_u) \) is

\[ R_u \leq \frac{cT}{2} \quad (2.15) \]

and rearranging Equation (2.14) leads to

\[ T \geq \frac{R_u2}{c} \quad (2.16) \]

and \( R_u \) is equivalent to the slant range swath extension, \( \Delta W \), as shown in Figure 2.4. Combining Equation (2.15) with the geometry in Figure 2.4, the upper limit for the PRF can be calculated as

\[ PRF = \frac{1}{T} \leq \frac{c}{2} \frac{1}{\Delta W} \approx \frac{c}{2} \frac{1}{\frac{R}{L_s} \tan \theta_v} \quad (2.17) \]
Figure 2.4: Slant range swath geometry to discern range ambiguity

where

\[ R = \text{Line of sight range from antenna to target} \]

\[ \Delta W = \text{Transmitted pulse bandwidth} \]

\[ L_r = \text{Length of antenna along the range dimension} \]

\[ \theta_v = \text{antenna vertical beamwidth} \]

\[ \theta_i = \text{incidence angle} \]

On the other hand, Equation (2.11) pointed out that the Doppler bandwidth is constrained [Ulaby, et al., 1981] by the antenna beamwidth as it is repeated here to be

\[ \Delta f_D \approx \frac{2}{\lambda} v_{rel} \theta_{az} \quad (2.18) \]

The radar PRF must be chosen to be greater than this as to limit aliasing and avoid Doppler (azimuth) ambiguity, thereby yielding an overall expression for the lower limit of the PRF given by

\[ \text{PRF} \geq \frac{2}{\lambda} v_{rel} \theta_{az} = \frac{2v_{rel}}{L_{az}} \quad (2.19) \]
2.8 Linear Frequency Modulation (LFM)

Most SAR systems use some form of linear frequency modulation (LFM) to achieve a high range resolution. Actually, this design is more interested in the chirp radar signal, which uses a combination of a pulse and FM signals. However, the area where the chirp and the LFM generally differentiates themselves in that the chirp requires a matched filter to de-chirp the echo signal for processing (thereby achieving maximum SNR), whereas the LFM simply uses the heterodyne process to prepare the echo for interpretation by the DSP. This thesis will use the terms; chirp and LFM, interchangeably since the design will use the process of implementing a pseudo-matched filter by shifting the echo signal.

The general principle of the transmitted chirp waveform is illustrated in Figure 2.5.

The SAR design synthesizes a stepped-chirp FM pulsed wave as it is transmitted waveform. Part (a) of Figure 2.5 shows the instantaneous frequency transmitted, that is, the FM begins with an upward sweep for a period of $\frac{T}{2}$, followed by a downward sweep for the remainder of the pulse width. The transmitted waveform shown in Part (b) is modulated in frequency from a lower frequency to a higher frequency and return to the lower frequency in the time duration of the pulse width (T).
One of the remarkable aspects of this SAR design lies with the simple synthesis of the chirp waveform. A software programmable A/D card supplies a linear voltage signal to an RF VCO, which then produces the chirp FM waveform. A computer-timing card manages the overall timing and control of the system, to include the creation of the synthesized wave.
Chapter 3

System Design

3.1 Design Consideration

This chapter provides a simplified system design for a rudimentary, low-altitude, low-
velocity synthetic aperture radar. This thesis provides for a baseline, or rather, an
introductory step-evolution towards this goal. The goal of this project is to design
and develop a high range resolution (HRR) radar with system parameters that can
be expanded into an X-band SAR system in the future. The HRR radar is able to
detect various simultaneous targets (trihedrals) and also be able to distinguish these
targets in the slant-range configuration. Some important design issues that have been
considered include:

(a) Transmit Waveform

In radar remote sensing, there are two widely used configurations, namely the
pulse and FM-CW (frequency modulated continuous wave) schemes. The finite
duration of the pulse permits range discrimination for the pulse radar at an
increased average power than that of the CW signal. However, a third scheme
combines the pulse and the FM: this is the chirp waveform. The duration of the
chirp waveform is longer than would be required for the range discrimination,
but the energy in the pulse is the same as it would be for the equivalent short
pulse. This permits the same maximum range that could be achieved with a high
peak-power short pulse. Figure 3.1 shows the transmit waveform used in this system.

![Waveform Diagram](image)

Figure 3.1: *Synthesized transmitted chirp waveform characteristics.*

The pulse duration width, $T$ is $172.4 \, \mu s$; up-ramp of $86.4 \, \mu s$ ($216 \times 200 \, \text{ns} \times 2 = 86.4 \, \mu s$), plus the down-ramp of $215 \, \mu s$ ($215 \times 200 \, \text{ns} \times 2 = 86.0 \, \mu s$). The pulse repetition interval (PRI) is $2.5 \, \text{ms}$ (or equivalently, $\text{PRF} = 400 \, \text{Hz}$). The need to start the chirp $2.0 \, \text{ms}$ into the period is to ensure that the returns arrive only from the footprint illuminated on the ground at the specified aircraft altitude and incidence angle.

(b) Operating Frequency

The operating frequency of any radar system is primarily based on the function the system is to accomplish. At this preliminary stage of this SAR system, the functional goal is to simply distinguish between simultaneous slant-range targets at a determined resolution. However, ultimately this SAR will serve as an airborne-platform, remote-sensing radar for underlying terrain parameters. With this function in mind, the radar system is operated in the X-band range with a frequency band between $9.8 \, \text{GHz}$ to $10.2 \, \text{GHz}$ and a bandwidth of $400 \, \text{MHz}$. The extent of penetration into vegetation depends upon the moisture content and density of the vegetation as well as upon the wavelength of the signal. At X-band, the wavelength is approximately $3 \, \text{cm}$ in length and roughly on the
same scale of the crop stems or vegetation it will be measuring. Since X-band signals do not penetrate into the tree canopy, it will be able to yield information about the upper layers of the vegetation and superficial layers of certain types of ground cover. Parameters that are of interest include: leaf area index (LAI), above ground biomass, and tree height.

(c) Operating Platform

To simulate the low altitude (an indication of grazing angle), low-velocity (an indication of Doppler) effects of the SAR, the radar system is mounted on a telescopic boom-mounted van. The altitude, grazing angle and velocity parameters considered by the system are restricted to the limitations imposed by this platform. Figure 3.2 shows the ground testing geometry.

![Figure 3.2: Overview of radar platform in relation to target](image)

(d) Calibration

Calibration is needed to remove measurement errors due to inherent instrumentation measurements and measurement techniques. To remove the internal system variations such as the ambient temperature fluctuations, phase change due to cable flex/length mismatches and gain drifts in the amplifiers. The DSP section will address and correct for these calibration errors during processing. Other errors, such as distortions caused by active components and antennas, will be adjusted
for or removed when comparing the received echoes with a measured response of a known calibration target (i.e. a conducting sphere).

(e) Signal Processor

The processing of the return signal into an acceptable image is arguably one of the more complex aspects of the SAR design, following the signal generation. However, the DSP of the received echo is beyond the scope of this research and will not be considered in this thesis.

A summary of the system requirements listed in Table 3.1 are used as basic guidelines to select suitable system parameters.

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Design Considerations</th>
</tr>
</thead>
<tbody>
<tr>
<td>System configuration</td>
<td>Stepped-chirp FM synthesized waveform</td>
</tr>
<tr>
<td>Operating frequency</td>
<td>X-Band</td>
</tr>
<tr>
<td>Calibration</td>
<td>Internal and external measurement errors addressed in signal processing</td>
</tr>
<tr>
<td>Operating platform</td>
<td>A mobile boom-mounted van</td>
</tr>
<tr>
<td>Signal processor</td>
<td>PC-based system</td>
</tr>
</tbody>
</table>

3.1.1 Design Parameters

In Section 2.2, the radar equation derived shows how the received power is related to radar and target parameters of the system. The equation is duplicated here for convenience and is used to examine the individual parameters of the system.

\[
\text{SNR}_{\text{antenna}} = \frac{P_i G^2 \sigma \lambda^2}{(4\pi)^3 R^4 F_N k T B L} \tag{3.1}
\]
The waveform generation uses an X-band VCO that has a linear response in the required 10 GHz range and the average transmit power of 12.6 mW (+11 dBm). The operating frequency is at X-band with a wavelength ($\lambda$) of 3 cm for the reasons described in Section 3.1(b). The RF amplifier amplifies the transmitted signal ($P_t$) by +29 dBm to +40 dBm. The radar cross section (RCS) is proportionally related to the amount of reflected power diverted back towards the receiver from the illuminated area. A value of 10 dB will be used for the average RCS fluctuations ($\sigma$): this value is based on the several factors, such as terrain cover, angle of incidence, target composition, etc. [Skolnik, 2001]. The system uses two separate microstrip antenna arrays, each with virtually identical gains of 25 dBi. For a ground-based system operating at a boom-mounted van platform, the radar-to-target range ($R$) can vary between 20 meters to 100 meters.

Noise is a major factor limiting overall system performance. The SAR system design has a low noise amplifier (LNA) as the front end of the receiver. A consequence based on how the overall noise figure affects the system performance is considered. A complete measure of the noise sensitivity of the receiver takes into account the noise figures and gains of the cascaded networks; i. e. the LNA, the mixer stage, the IF amplifier stage and any losses in the RF transmission line. The receiver noise is calculated as

$$F_N = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_2G_1} + \cdots$$  \hspace{1cm} (3.2)$$

with the $F_x$'s and the $G_x$'s representing the individual cascaded network losses or gains within the receiver, respectively. Finally, the remaining parameters in Equation (3.1) are considered to be constants and are system independent.

Table 3.2 summarizes the significant design parameters of the X-band SAR system.
Table 3.2: SAR System Design Parameters

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Selected Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>System configuration</td>
<td>Stepped-chirp FM synthesized waveform</td>
</tr>
<tr>
<td>Operating wavelength, $\lambda$</td>
<td>3 cm (X-band)</td>
</tr>
<tr>
<td>Transmit power, $P_t$</td>
<td>20 dBm - attenuated from original 10W (or 40 dBm)</td>
</tr>
<tr>
<td>Received power, $P_r$</td>
<td>-35.5 dBm (min range), -63.5 dBm (max range)</td>
</tr>
<tr>
<td>Measurement range, $R$</td>
<td>20 meters (min) to 100 meters (max)</td>
</tr>
<tr>
<td>Average RCS, $\sigma$</td>
<td>10 dB</td>
</tr>
<tr>
<td>LNA noise figure, $F_N$</td>
<td>3 dB</td>
</tr>
<tr>
<td>IF filter (center frequency)</td>
<td>30 MHz</td>
</tr>
<tr>
<td>IF bandwidth, $B$</td>
<td>7 MHz</td>
</tr>
<tr>
<td>Antenna gain, $G$</td>
<td>25 dBi</td>
</tr>
<tr>
<td>Thermal noise</td>
<td>-133.8 dBm</td>
</tr>
</tbody>
</table>

Separating Equation (3.1) into the noise and the received power considerations to realize the approximate signal to noise ratio results in:

Noise power: \[ P_n = K T F_N B = -103 \text{ dBm} \]

Received power: \[ P_r = \frac{P_r G^2 \sigma \lambda^2}{(4\pi)^3 R^4} = \begin{cases} -35.5 \text{ dBm} & R = 20 \text{ m} \\ -63.5 \text{ dBm} & R = 100 \text{ m} \end{cases} \]

Therefore, the signal-to-noise ratio (SNR) is computed as

\[ \text{SNR} = \frac{P_r}{P_n} = \begin{cases} 67.5 \text{ dB} & \text{for near reflectors} \\ 39.5 \text{ dBm} & \text{for far reflectors} \end{cases} \]
3.2 System Description

The X-band SAR design is based on a low-cost approach relying on recent advancement in digital and RF technology to develop, what is historically a large and expensive instrument, into a compact and relatively inexpensive testing instrument.

In addition to the previously stated system design parameters, the SAR design is subjected to the constraints of being built on a low budget and having a minimal impact on the delivery platform (airplane, boom-mount van, etc.). These constraints combined, dictate many of the design decisions employed into the SAR.

The schematic of the integrated X-band SAR system is shown in Figure 3.3. In general, the design may be subdivided into four major sections:

- Signal Generation
- RF section
- Receiver section
- Data Acquisition

![Top-level block schematic of SAR design.](image)
3.2.1 Signal Generation Section

The signal waveform generation is perhaps the most vital and complex section within the entire system design. The SAR operates by timing the two-way delay for a short duration RF pulse, transmitted vertically downwards and to one side of the aircraft. The required level of accuracy in range measurements (better than 4 m) calls for a pulse length of a several nanoseconds; therefore, in order to reduce the RF power requirements, a pulse compression (chirp) technique is used. The waveform can be tied directly to the slant-range resolution of the system. Additionally, the PRF, pulse width, bandwidth and the linearity of the VCO output signal are vital to the SNR of system.

An ideal chirp waveform would have a perfect ramp function of frequency versus time as depicted in Figure 3.4 (a) where at any time along the duration of the transmitted pulse \( T \) lies a unique frequency \( f_{TX} - f_{RX} = f_{IF} \), called the intermediate frequency.

![Figure 3.4: (a) Ideal LFM Waveform. (b) Actual LFM Waveform.](image)

The ideal LFM ramp could be generated by a very expensive and sizeable X-band synthesizer. However, with the constraints imposed on the design of low-cost and having a small footprint, a solution of generating the waveform using a D/A computer card and an X-band VCO was implemented. In Figure 3.4 (b), the actual
waveform generated from the D/A-VCO combination illustrates a *stair-step* ramp. It is important that each vertical step length be reduced as to approach a linear slope. The *stair-step* emerges as a result of the limitations of the generation rate (Rg) and the output bit-resolution of the D/A card, and the inherent settling time of the VCO.

The range resolution in Figure 3.4(a) is calculated as

$$\delta_r = \frac{c}{2B}$$

(3.4)

when the LFM has a linear slope, or the alternative equation

$$\delta_r = \frac{ct}{2}$$

(3.5)

can be used when the solving for a stair-step slope as in Figure 3.4 (b). In Figure 3.4 (a), the pulse duration of chirp is equal to the reciprocal of the bandwidth, i.e. \( T = 1/B \) (or equivalently; \( \tau = 1/B \)). However, in Figure 3.4 (b), the pulse width \( \tau \) is no longer equal to the reciprocal of the bandwidth but is now equal to the width of horizontal step in the stair-stepped LFM \( (T \neq t \text{ and } \tau \neq 1/B) \).

Table 3.3 lists the range resolution results from both equations.

<table>
<thead>
<tr>
<th>Equation 3.5 (stair-step slope)</th>
<th>Rg (Megasample/sec)</th>
<th>T (( \tau = 1/Rg))</th>
<th>Range Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>80 Ms/s</td>
<td>12.5 ns</td>
<td>1.88 meters</td>
</tr>
<tr>
<td></td>
<td>40 Ms/s</td>
<td>25.0 ns</td>
<td>3.75 meters</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Equation 3.4 (LFM slope)</th>
<th>Bandwidth (B)</th>
<th>Pulse Width (T)</th>
<th>Range Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>400 MHz</td>
<td>2.5 ns</td>
<td>0.375 meter</td>
</tr>
</tbody>
</table>
Since the HRR system design will serve as the hardware platform for the X-band airborne SAR, certain compromises must be made at this stage between incorporating the airborne-platform SAR parameters (e.g. increased range-to-target) versus attempting to design for the greatest range resolution of a ground-based system. For an airborne-platform, the designated range-to-target is 750 meters, which has the optimal pulse duration of the chirp of 102.3875 µs (as explained in the Receiver Section of this document). The following table establishes the relationship between the output bit resolution of the D/A, the generation rate of the D/A card and the associated pulse duration.

### Table 3.4: D/A Bit-Resolution vs. Pulse Duration

<table>
<thead>
<tr>
<th></th>
<th>D/A Bit Resolution</th>
<th>Generation Rate (Rg)</th>
<th>Pulse Duration (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SAR design</td>
<td>216 bits</td>
<td>5 Ms/s</td>
<td>172.8 µs</td>
</tr>
<tr>
<td>Current design</td>
<td>216 bits</td>
<td>80 Ms/s</td>
<td>10.8 µs</td>
</tr>
<tr>
<td>With extender circuit</td>
<td>4096 bits</td>
<td>80 Ms/s</td>
<td>51.2 µs</td>
</tr>
</tbody>
</table>

The original design of the waveform generation was specifically designed to meet the parameter criterion for an airborne platform SAR. The design did not take advantage of the full output range of the D/A and was restricted to only 216 bits. This restriction of 216 bits required the pulse duration time to be significantly large to accommodate the greater range-to-target distance; therefore a low generation rate of 5 MHz is required. The 216-bit limitation results from difference in the bit representation of the output voltage from the D/A component (i.e. D/A bit 800 = 6.01 V = 10.2 GHz and D/A bit 584 = 6.99 V = 9.8 GHz; 800 - 584 = 216 bits). However, since the HRR radar is tested on the ground and at a considerable shorter range-to-target distance, a higher generation rate of 80 MHz can be considered. This higher rate will
provide a better range resolution for a shorter range. Again, this current design is still limited to the 216-bit limitation as explained with the SAR design. However, to gain a much finer range resolution with the existing hardware, an analog "extender" circuit (schematic in Appendix A.3) is included into this design and thereby taking advantage of the entire 4096-bit (216 bits = 4096) range from the output of the D/A card as illustrated in Figure 3.5.

![Figure 3.5: D/A Output Bit Resolution.](image)

The greater number of bits representing the same voltage range (6.01 V to 6.99 V) of the D/A output would equate to a smaller voltage increment feeding the varactor inside the VCO. The 4096-bit analog output of the D/A card is combined with the precision steady state analog voltage of the extender digital circuitry, to provide the required input to the VCO. The input analog feed to the VCO will provide a very precisely correlated voltage-input to frequency-output synthesizing the desired LFM waveform. The frequency on the VCO is tuned (modulated) by varying the applied voltage from the D/A-extender circuit combination. The tuning characteristics of the VCO are provided in Appendix A.1. The extender analog circuit addresses two separate problems: (a) the settling time parameter of the VCO is excessively large (~150 nsec), and (b) the output bit resolution of the D/A card was restricted to 216 bits. By extending the output bit resolution of the D/A card, the incremental analog
voltage input to the VCO was dramatically reduced, thereby preventing the internal varactor of the VCO from fully charging and discharging. This charging/discharging corresponds directly to the settling time of the VCO. Changing the analog input voltage to the VCO to smaller increments changes the charge level on the varactor only minimally and in turn, improves the recovery time of the VCO considerably. Increasing the recovery time of the varactor reduces the settling time of the VCO and provides for a more linear ramp to the waveform. If more time and resources were available to improve on the analog design, it is believed a greater gain would result in the range resolution from such a small contribution. It will be left up to the next generation of hardware to improve on this circuit.

Figure 3.6 summarizes the synthesized LFM-chirp waveform;

![Figure 3.6: Waveform parameters.](image)

### 3.2.2 Radio Frequency (RF) Section

The RF section consists of the aforementioned signal generated section, the power divider and the high power RF amplifier. Setting aside the signal generation, the RF section is fairly straightforward. The power divider receives the generated waveform,
where the signal is reduced by 3 dB at each exit port. The signal at one port is used as a local oscillator to be mixed with the received echo return. The signal at the remaining port is passed onto the RF amplifier where the power level is increased by 29 dB. Although the maximum RF amplifier gain is rated at +40 dB, an 11-dB attenuator at the output port limits the amplifier gain to +29 dBm. This prevents saturation of the receiver LNA for shorter-range operation. The full amplification of the RF amplifier will be required in the airborne SAR configuration, where the range to target is increased dramatically. After the power level of the signal has been amplified by the RF amplifier, the synthesized LFM chirp signal is then coupled into free space via the transmitting antenna.

### 3.2.3 Receiver Section

The main function of the receiver section (or IF section) is to filter and amplify the return echo signal. The receiver antenna receives the re-radiated echo returns from the target, which are at a considerable lower power level than when the signal left the transmitter. The LNA amplifies the return signal by 30 dB to elevate the weak signal to required SNR levels. At this stage, the signal is at RF and is rather difficult and expensive to manipulate. The high frequency signal is down-converted into an intermediate frequency (IF) signal by a mixing process. The single side band (SSB) mixer, a down-converter, outputs the difference in frequency between the copy of the transmitted signal and the received echo return: \( |f_{LO} - f_{RF}| = f_{IF} \).

Referring to the radar Equation (2.5), the received power decreases at a rate proportional to \( R^4 \), which means that if the range-to-target is doubled, the received power will decrease by a factor of 16. As the range increases, the maximum received power decreases but the minimum detectable power level remains the same. This is due to the fact that the thermal noise level is independent of range. Therefore,
the overall effect is the decrease in the system dynamic range. This decrease in the system dynamic range can be avoided by employing a variable gain amplifier in the IF stage. The AGC (automatic gain control) of the IF amplifier is adjusted via an external potentiometer, to accommodate the dynamic range the radar. The center frequency of the IF amplifier is at 30 MHz with a bandwidth of 10 MHz. The noise figure of this amplifier is 4 dB, which is still greater than that of the front-end LNA, and although the amplifier has the maximum gain of 70 dB, minimum gain will be used during ground testing. The last stage of the IF section is band pass filter (BPF). The BPF is inserted after the IF amplifier to reject any unwanted signal outside the passband (30 MHz ± 3.5 MHz). The IF frequency is directly proportional to the range, as can be seen the following equation

$$f_{IF} = \frac{2BR}{cT}$$  (3.6)

where $B$ is the system bandwidth, $T$ is the pulse duration time, $c$ is the speed of light and $R$ is the range-to-target. On the airborne SAR platform, the estimated range from radar to target is 1000 m, the pulse duration is 102.4 μs and the system bandwidth is 400 MHz. Substituting these values for the parameters in Equation 3.6 leads to the IF frequency of 30 MHz. To determine the bandwidth of the IF frequency, which directly determines the 3-dB cutoff requirement of the BPF, the following equation is used

$$\Delta f_{IF} = \frac{2B\Delta W}{cT}$$  (3.7)

where $\Delta W$ is the swath width on the ground as determined in Equation (2.21). In the SAR configuration, the swath width is 100 meters, which would bring the IF bandwidth to 7 MHz. The output of the band pass filter is then fed to the data acquisition section for further processing.
DESIGN AND DEVELOPMENT OF A LOW-COST
HIGH RANGE RESOLUTION X-BAND RADAR

by

Paul Cantu

A THESIS

Presented to the Faculty of
The Graduate College at the University of Nebraska
In Partial Fulfillment of Requirements
For the Degree of Master of Science

Major: Electrical Engineering

Under the Supervision of Professor Ram M. Narayanan

Lincoln, Nebraska

August, 2003
DESIGN AND DEVELOPMENT OF A LOW-COST HIGH RANGE RESOLUTION X-BAND RADAR

Paul C. Cantu, M.S.
University of Nebraska, 2003

Advisor: Ram M. Narayanan

Synthetic Aperture Radar (SAR) is one of the main tools for microwave remote sensing because of its multi-dimensional high resolution characteristics and the capability to operate in nearly all weather conditions, day and night. The University of Nebraska-Lincoln (UNL) initiated the design and development of a low-cost airborne SAR in January 2001 to support our Airborne Remote Sensing Program. The objectives of this project are separated into various evolutionary stages. This thesis will focus on the initial phase of design and construction of an X-band high range resolution radar (HRR) using basic RF/microwave and digital components. The following stages will expand on the HRR design to achieve a functioning X-band airborne SAR for the remote sensing of underlying vegetation parameters (tree height, leaf area index, biomass content, etc.) from a low altitude platform from a range of 1000 meters. The SAR system is an X-band, stepped-chirp FM, single polarization radar system. One of its unique features is that the signal generation consists of a timing-controlled D/A converter and VCO arrangement to generate the step-chirp signal, thereby allowing for less design complexity and a much lower overall system cost.
The individual block-segments of the SAR include a stepped-chirp FM waveform synthesizer, transmission and receive paths, antennas, quadrature detection and image signal processing. The system underwent rigorous in house laboratory testing and subsequent outdoor field-testing from a van-mounted boom where preliminary HRR one-dimensional images were obtained. It is anticipated that the following progression of development for this HRR system will be to use this design as a basis towards fully coherent, data acquisition from an airborne platform.
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Chapter 1

Introduction

1.1 Motivations

Monitoring the status and condition of the earth's terrain, such as vegetation fields, forests and soil surfaces is of primary importance to the understanding and the protection of the environment, as well as for natural resource management. It has been recognized that the use of microwaves in remote sensing is a suitable approach as microwave signals are able to penetrate cloud and vegetation cover and can operate independent of solar illumination. Moreover, microwaves have the ability to penetrate more deeply into vegetation compared to optical signals. In general, longer wavelengths penetrate vegetation much better than shorter wavelengths. Currently, there exist operational C-band and L-band systems operating in the 5 GHz and 1 GHz frequency bands, respectively, that are routinely used for sea surface and soil moisture remote sensing. However, higher frequencies such as X-band (10 GHz) are used for radar systems where the size of the antenna constitutes a physical limitation; this becomes especially important when considering an airborne platform. The main thrust in developing the X-band Synthetic Aperture Radar (SAR) is to provide a light-weight, low-cost system to ultimately operate from an airborne platform. The fact that the system operates in the X-band microwave range will allow for smaller RF (radio frequency) components and antennas, relative to C-band and L-band, making
it ideal for airborne operations. The system design is one which is easy to maintain and operate, and capable of quick turn-around in data acquisition.

Ultimately, X-band SAR will provide remote sensing of underlying terrain parameters (tree height, soil moisture, biomass content, etc.) from an airborne platform; however, there are several evolutionary steps needed towards accomplishing this goal. This thesis will focus on the design and construction of the X-band high range resolution (HRR) radar, which will be the core hardware design to the airborne SAR system. The HRR radar will be tested and validated using ground-based tests from a mobile boom-mounted van.

1.2 Scope of the Thesis

In this thesis project, a ground-based HRR radar system has been designed, constructed and tested. This system is mounted on a mobile, van boom-arm during the testing phase and validation phases. There are three objectives identified for this project:

i. To design, construct and test a HRR radar system

The system design will be simple and straightforward. Simple, lightweight, compact, and low-cost design are the overall top-level requirements.

ii. To validate the HRR radar resolution

The HRR will produce a ground range resolution better than 4 meters.

iii. To use this technology as a basis towards the next evolution stage, i.e., graduating this HRR radar system to an airborne SAR platform

The experience gained in this project will be used for future evolutions towards an airborne platform SAR system.
This thesis reports on the design, development and testing results of the HRR radar system. Chapter 2 will review radar fundamentals, especially as they relate towards the design of a synthetic aperture radar system using the HRR radar as the system hardware. In particular, the concept of the stepped-chirp signal generation is highlighted.

A detailed description of the system design and its operation is presented in Chapter 3. In general, the construction of the SAR can be subdivided into five parts. These include (a) the signal generation, (b) the radio frequency (RF) section, (c) the antennas, (d) the receiver section and (e) the data acquisition section. The HRR is a quasi-monostatic radar, where a dual antenna system, in close relation to each other, is employed in this design. This design configuration will eliminate the problems associated with transmit/receive isolation, to minimized design complexity and the potential to expand the system for interferometric applications. The RF section is constructed from individual RF components. These include the use of a power splitter, band pass filter and RF amplifier. The main function of the receiver section is to filter and amplify the intermediate frequency signals received from the mixer output. The pre-processed signals are then converted into digital data by the data acquisition section, which includes an analog-to-digital (A/D) computer card, computer software and the computer storage. Analysis and processing of the data can be performed virtually in real-time while still gathering data on the aircraft or in the field.

Several radar cross section measurements have been carried out to verify the operation of the SAR system. These include the measurements of standard targets such as trihedrals of differing sizes at various downrange distances. The analysis, results and observations are presented in Chapter 4.

Finally, Chapter 5 will conclude with a summary of work performed and future recommendations for this X-band high-resolution radar.
Chapter 2

Review of Radar Theory

2.1 Radar Fundamentals

The term “radar” stands for RAdio Dectection And Ranging. This term was coined in the early 1950s, at a time when detection of ships and determination of their ranges were its primary purpose. However, after the proliferation of high speed computing, digital processing and other advances, radar has graduated beyond simple detection and ranging of targets, and is used in a wide range of applications for the military, air traffic control, police departments and remote sensing, to mention a short list.

Generally, remote sensing refers to the process of recording/observing/perceiving (sensing) objects or events at far away (remote) places. In remote sensing, the sensors are not in direct physical contact with the objects or events being observed. The technique requires an electromagnetic wave, of suitable characteristics, to travel from the objects/events to the sensors through the atmospheric medium. The choice of electromagnetic radiation wavelength (or frequency) depends upon the specific remote sensing application.

Synthetic aperture radar is basically a method of ground mapping and imaging that uses computer processing to sharpen the azimuthal resolution over what can be achieved by a conventional antenna. This technique first appeared in the early 1950s, but did not reach a high state of development for almost 30 more years. With the introduction of high-speed digital processing and other advances, SAR is rapidly emerging as a powerful remote sensing tool (Ulaby, et al., 1981).
2.2 The Radar Equation

In microwave remote sensing, the distinction between different ground targets being sensed is primarily caused by the difference in the signal strength received by the radar. Therefore, it is understood the received signal strength is the most important measurement in remote sensing. In the simplest case, the radar equation can be stated verbally as suggested by [Ulaby et.al, 1981]:

\[
\begin{align*}
\text{Power received} &= \left( \text{Power per unit area at target} \right) \times \left( \text{Effective scattering area of the target} \right) \times \\
&\quad \left( \text{Spreading loss of radiated signal} \right) \times \left( \text{Effective antenna receiving area} \right)
\end{align*}
\]

where the power at the receiver is the energy scattered back from the target (i.e. backscatter). The power per unit area at the target is the amount of energy intercepted by the ground target as a result of its physical characteristics. The effective scattering at the target represents the re-radiation of energy in all directions from the point source (ground target), to include that fraction of power received at the receiver. The spreading loss of radiated signal stems from the nature of the radiated power from the point source spreading out in all directions in the area of a sphere of radius \( R \). The further away from the point source, or the greater the radius \( R \), the less power received at the receiving antenna. The actual size of the receiving antenna will also affect the total power received.

The *phonetic* equation can be used to assist in deriving the fundamental radar equation as follows:

\[
P_r = \frac{P_t G_t A_r \sigma}{(4\pi R^2)^2}
\]

(2.1)
\[ P_r = \text{Power received} \]
\[ P_t = \text{Power transmitted} \]
\[ G_t = \text{Gain of the transmitting antenna} \]
\[ \sigma = \text{Radar cross section} \]
\[ G_t = \text{Gain of the transmitting antenna} \]
\[ G_t = \text{Gain of the transmitting antenna} \]

This equation can be broken down further for clarification:

\[ P_r = P_t \times G_t \times \frac{1}{4\pi R^2} \times \sigma \times \frac{1}{4\pi R^2} \times A_r \]

\[ (1) = (2) \times (3) \times (4) \times (5) \times (6) \times (7) \] (2.2)

The power received \( P_r \) by the radar system is a function of a linear frequency modulated chirp pulse of electromagnetic energy, \( P_t \), that has been focused down by the transmitting antenna to a beam width, so that the energy level increases by a factor of \( G_t \) over a spherically expanding wave. This focused energy illuminates an area on the ground that has a radar cross section (RCS) of \( \sigma \). The backscattered power then radiates isotropically (or spherically) in all directions from the point scatter. The receiving antenna area, \( A_r \), intercepts a fraction of this backscatter signal for processing.

Although the SAR sports a two-antenna design, both antennas are generally identical to each other in all characteristics, including gain, i.e. \( G = G_t = G_r \). As per [Ulaby, et al., 1983], the relationship between the antenna directive gain and receiving aperture area is:

\[ G = \frac{4\pi A_r}{\lambda^2} \] (2.3)

or equivalently:

\[ A_r = \frac{G\lambda^2}{4\pi} \] (2.4)
where \( \lambda \) is the wavelength of the radar system. Substituting Equation (2.4) into Equation (2.2) or (2.1) results in a modified radar equation:

\[
P_r = \frac{P_t G^2 \sigma \lambda^2}{(4\pi)^3 R^4}.
\]

(2.5)

There are several forms of this radar equation; however, one common form equates the ratio of the returned signal power, \( S \) (or \( P_r \)), to the receiver noise power, \( N \). The noise figure of the receiver, \( F_N \), and the thermal noise power, \( kTB \) (\( k = \) Boltzmann constant, \( T = \) Temperature, \( B = \) Bandwidth), as well as the system and environmental losses, \( L \), are integrated into Equation (2.6) to become the radar equation of interest. The Signal-to-Noise (power) ratio at the receiver antenna port is

\[
SNR_{\text{antenna}} = \frac{P_t G^2 \sigma \lambda^2}{(4\pi)^3 R^4 F_N kTBL}.
\]

(2.6)

2.3 SAR Principles

SAR systems can produce maps (or images) of radar reflectivity versus range and azimuth, typically in the form of a two-dimensional (2-D) image. One dimension in the image is called range and is a measure of the “line-of-sight” distance from the radar to the target. Range measurement and resolution are determined by precisely measuring the time from transmission of a pulse to receiving the echo from a target. Fine range resolution can be accomplished through pulse compression techniques, such as in this design: linear frequency modulation (LFM) chirp.

The other dimension is called azimuth (cross-range) and is perpendicular to range. It is the ability of the SAR to produce a relatively fine azimuth resolution which differentiates it from other radars. To obtain fine azimuth resolution, a physically large antenna is needed to focus the transmitted and received energy into a sharp beam. The sharpness of the beam defines the azimuth resolution. Fine azimuth
resolution is enhanced by taking advantage of the radar motion in order to synthesize a larger antenna aperture, which exploits the Doppler effect. The distance the radar platform travels in synthesizing the antenna length is known as the synthetic aperture. A narrow synthetic beamwidth results from the relatively long synthetic aperture, which yields finer resolution than is possible from a smaller physical antenna.

The quality of SAR images is heavily dependent on the size of the map resolution cell. The resolution cell is the area comprised of the range resolution by the azimuth resolution. In the following subsections, the range and azimuth resolutions are explained and calculated for this design.

Figure 2.1: *Footprint Definition.*

Figure 2.1 shows the geometry for the standard side looking SAR. As noted, the intersection of the antenna beam with the ground defines a footprint. As the platform moves, the footprint scans a swath on the ground, whose width is determined by the antenna beamwidth, range and incidence angle. The antenna 3 dB beamwidth is $\theta_{3dB}$, the range from the platform to the target is $R$, and the incidence angle is $\theta_i$. Therefore, the size of the footprint painted on the ground is

$$R\theta_{3dB} \sec \theta_i = \frac{R\theta_{3dB}}{\cos \theta_i} = \frac{RA}{L \cos \theta_i} \text{ (range direction)}$$
where $L_v$ is the vertical length and $L_h$ is the horizontal length of the physical antenna, and $\lambda$ is the radar wavelength.

2.4 Range Resolution

Ground range resolution is defined as the minimum distance on the ground at which two object points can be imaged separately, shown in Figure 2.3, as the distance $\delta R_g$.

![Figure 2.2: Definition of Range Resolution Between Two Points](image)

Two individual point targets can be distinguished if the leading edge of the pulse echo from the more distant point target arrives at the antenna later than the trailing edge of the pulse echo from the nearer object. Thus, if their distances to the antenna are different by at least half of the pulse length (product of the pulse width, $t$, and the wave propagation speed, $c$), then the two objects are resolvable [Wehner, 1995]. Range resolution is determined primarily by the transmitted pulse bandwidth, i.e. a narrower pulse yields a finer range resolution. Therefore the ground range resolution ($\delta R_g$) is:

$$\delta R_g = \frac{ct}{2}$$ (2.7)
\[
\delta R_g = \frac{\delta R_s}{\sin \theta_i} = \frac{T}{2\sin \theta_i} = \frac{c}{2B \sin \theta_i}
\]  
(2.8)

where \(\delta R_s\) is the slant resolution, \(\theta_i\) is the incident angle, \(T\) is the pulse length, \(t\) is the pulse width, \(B\) is the frequency bandwidth of the transmitted radar pulse and \(c\) is the speed of light.

The ground range resolution is infinite for a vertical look angle \((\theta_i = 0^\circ)\) and improves as the look angle is increased. Also, note that the ground range resolution is independent of the height of the radar platform. In order to achieve a fine range resolution, the pulse width must be minimized, which is accomplished via the frequency-modulated pulse (chirp) rather than resorting to using very short physical pulses. This, in effect, will reduce the average transmitted power and increase the operating bandwidth. Achieving fine range resolution while maintaining adequate average transmitted power can be accomplished by using a pulse compression (i.e. modulation) technique as described in Section 2.8.

2.5 Azimuth Resolution

The SAR system has a major advantage over the real-aperture side looking airborne radar (SLAR) in that the resolution in the azimuth direction is independent of the radar platform altitude or range distance to the target. SAR systems depend upon two conditions; (1) the motion of the radar platform relative to a stationary target, and (2) signal processing. Both of these conditions are used to synthesize an antenna (aperture) that is much longer in length than the actual antenna hardware. Condition (1) achieves this by taking advantage of the Doppler effect, and condition (2) by achieving a much narrower beamwidth in the along-track direction than that attainable with the real-aperture systems. Therefore, a larger synthetic aperture produces a finer azimuth resolution.
The SAR system transmits the microwave pulses to the imaging area and the echoes are received as "raw" SAR data, which are coherently recorded onto storage. It is these recorded echoes, which generate a coarse (raw) microwave reflectance 'image' of the illuminated ground area. The natural dimensions of this image are range and azimuth. Whereas the range resolution of the raw image is dictated by the pulse width (as described in Section 2.4), the 'raw' azimuth resolution is determined by the azimuth extent of the antenna pattern footprint on the ground, the antenna size, wavelength and the imaging geometry (SLAR properties). This 'raw' data in azimuth, would exhibit an unacceptable image for this dimension. Subsequent image processing of the 'raw' SAR data is required to dramatically improve the azimuth resolution by taking advantage of the coherence of successive echo signals as the antenna moves.

This unacceptable "raw" image is a product of the broad antenna pattern, i.e. a SLAR property, which is based only upon the magnitude of the received signal. It is the signal phase histories over azimuth that exhibits an extremely sensitive measure for the instantaneous distance between the radar receiver and the point target. After correlation of the transmitted chirp signal and the receivers echo 'raw' data in signal processing, the SAR data are focused to an azimuth resolution on the order of

\[ \delta R_a' = \frac{L_h}{2} \]  \hspace{1cm} (2.9)

where \( L_h \) is the length of the antenna. This is the finest potential synthetic azimuth resolution that can be theoretically achieved. However, Equation (2.9) is not the only parameter that has an impact on the SAR data. The coherent nature of the SAR signal will produce speckle in the image. To remove the speckle, the image is processed in digital signal processing (DSP) by averaging several looks. By increasing the looks, the interpretability of the SAR image is significantly enhanced; however this is at the
cost of the azimuth resolution. Therefore, the azimuth resolution Equation (2.9) must be adjusted to include the effects of averaging, i.e.,

\[ \delta R_a = n \times \frac{L_h}{2} \]  \hspace{1cm} (2.10)

where \( n \) is the number of looks to be averaged. Since it is desired to have a virtual square resolution cell, the nominal range resolution can be adjusted to meet this goal by selecting an appropriate value for \( n \).

### 2.6 Doppler Bandwidth

Tracking the Doppler shift history of a point target as it is illuminated by the radar provides the information necessary to resolve the azimuth location of the target. The Doppler shift history is obtained by comparing the reflected signals from a ground target with a reference signal that incorporates the same frequency of the transmitted pulse. The output is known as a phase history, and it contains a record of the Doppler frequency changes plus the amplitude of the returns from each ground target as it passed through the beamwidth of the antenna. Therefore, the Doppler bandwidth can be viewed as being constrained by the antenna beamwidth as

\[ \Delta f_D \approx \frac{2}{\lambda} v_{rel} \theta_{az} \]  \hspace{1cm} (2.11)

and since \( \theta = \frac{\lambda}{L_\theta} \), Equation (2.11) becomes

\[ \Delta f_D \approx \frac{2v_{rel}}{L_{az}} \]  \hspace{1cm} (2.12)

\( \Delta f_D \) = Doppler bandwidth

\( \lambda \) = Transmitted wavelength

\( \theta_{az} \) = Antenna beamwidth in the azimuth dimension

\( v_{rel} \) = Relative radial velocity of the radar platform to the target

\( L_{az} \) = Length of antenna along the azimuth dimension
A further detailed derivation for the Doppler bandwidth is illustrated with Figure 2.3.

Figure 2.3: Doppler spread geometry of a synthetic aperture

The Doppler frequencies received from ground-target return signals will increase as the radar platform moves from position 1, where there is a maximum Doppler shift of \(-\delta f_d\), towards position 2, where the Doppler frequency shift will become zero. As the radar platform passes through position 2, the Doppler frequency shift will continue towards \(+\delta f_d\) as the radar platform approaches position 3. It is the frequency range between positions 1 to 3, where the maximum Doppler shifts occur or when the target first enters and exits the radar beam width, respectively, which determines the Doppler bandwidth, \(\Delta f_D\):

\[
\delta f_D = 2|\Delta f_d| = 2\frac{2\nu}{c} f_s = 2\frac{2\nu_{rel}\sin\phi}{c} \frac{c}{\lambda} = 2\frac{2\nu_{rel} L_s}{R} \lambda
\]  

(2.13)

(substituting \(L_s = \frac{R\theta_{az}}{2}\) and \(\theta_{az} = \frac{\lambda}{L_{az}}\) in Equation (2.13) leads to

\[
\Delta f_D = 2\frac{2\nu_{rel}}{\lambda R} \frac{R\lambda}{2L_{az}} = 2\frac{\nu_{rel}}{L_{az}}
\]  

(2.14)
where

\[ L_s = \text{Synthetic aperture length of beamwidth} \]
\[ R = \text{Distance from radar platform to target at bore sight} \]
\[ c = \text{Speed of wave propagation} \]
\[ f_s = \text{Operating frequency} \]

### 2.7 Ambiguities and Pulse Repetition Frequency (PRF)

Ambiguities can exist with pulse radars for both range and azimuth measurements. It is the pulse repetition frequency (PRF) that controls the extent of overlap of adjacent radar echoes, accounting for the ambiguities. Generally, the radar maximum PRF must be low enough to avoid range ambiguity and the minimum PRF must be high enough to avoid Doppler ambiguity.

There is an upper limit on the PRF imposed by geometry (swath width). The PRF will need to be lower than the maximum bandwidth of the transmitted pulse (\( \Delta W \)) to prevent overlapping of the near-range and far-range return pulses, hence range ambiguity. The maximum unambiguous range (\( R_u \)) is

\[
R_u \leq \frac{cT}{2}
\]

and rearranging Equation (2.14) leads to

\[
T \geq \frac{R_u \Delta}{c}
\]

and \( R_u \) is equivalent to the slant range swath extension, \( \Delta W \), as shown in Figure 2.4. Combining Equation (2.15) with the geometry in Figure 2.4, the upper limit for the PRF can be calculated as

\[
PRF = \frac{1}{T} \leq \frac{c}{2 \Delta W} \approx \frac{c}{2 \frac{R\Delta}{L_r} \tan \theta_v}
\]
where

\[ R = \text{Line of sight range from antenna to target} \]

\[ \Delta W = \text{Transmitted pulse bandwidth} \]

\[ L_f = \text{Length of antenna along the range dimension} \]

\[ \theta_v = \text{antenna vertical beamwidth} \]

\[ \theta_i = \text{incidence angle} \]

On the other hand, Equation (2.11) pointed out that the Doppler bandwidth is constrained [Ulaby, et al., 1981] by the antenna beamwidth as it is repeated here to be

\[ \Delta f_D \approx \frac{2}{\lambda} v_{rel} \theta_{az} \]  \hspace{1cm} (2.18)

The radar PRF must be chosen to be greater than this as to limit aliasing and avoid Doppler (azimuth) ambiguity, thereby yielding an overall expression for the lower limit of the PRF given by

\[ PRF \geq \frac{2}{\lambda} v_{rel} \theta_{az} = \frac{2v_{rel}}{L_{az}} \]  \hspace{1cm} (2.19)
Most SAR systems use some form of linear frequency modulation (LFM) to achieve a high range resolution. Actually, this design is more interested in the chirp radar signal, which uses a combination of a pulse and FM signals. However, the area where the chirp and the LFM generally differentiates themselves is that the chirp requires a matched filter to de-chirp the echo signal for processing (thereby achieving maximum SNR), whereas the LFM simply uses the heterodyne process to prepare the echo for interpretation by the DSP. This thesis will use the terms; chirp and LFM, interchangeably since the design will use the process of implementing a pseudo-matched filter by shifting the echo signal.

The general principle of the transmitted chirp waveform is illustrated in Figure 2.5.

![Frequency-Time Plot](a) Frequency-Time Plot ![Amplitude-Time Plot](b) Amplitude-Time Plot

Figure 2.5: Linear Frequency Modulation (LFM) waveform

The SAR design synthesizes a stepped-chirp FM pulsed wave as it is transmitted waveform. Part (a) of Figure 2.5 shows the instantaneous frequency transmitted, that is, the FM begins with an upward sweep for a period of \( \frac{T}{2} \), followed by a downward sweep for the remainder of the pulse width. The transmitted waveform shown in Part (b) is modulated in frequency from a lower frequency to a higher frequency and return to the lower frequency in the time duration of the pulse width (T).
One of the remarkable aspects of this SAR design lies with the simple synthesis of the chirp waveform. A software programmable A/D card supplies a linear voltage signal to an RF VCO, which then produces the chirp FM waveform. A computer-timing card manages the overall timing and control of the system, to include the creation of the synthesized wave.
Chapter 3

System Design

3.1 Design Consideration

This chapter provides a simplified system design for a rudimentary, low-altitude, low-velocity synthetic aperture radar. This thesis provides for a baseline, or rather, an introductory step-evolution towards this goal. The goal of this project is to design and develop a high range resolution (HRR) radar with system parameters that can be expanded into an X-band SAR system in the future. The HRR radar is able to detect various simultaneous targets (trihe drals) and also be able to distinguish these targets in the slant-range configuration. Some important design issues that have been considered include:

(a) Transmit Waveform

In radar remote sensing, there are two widely used configurations, namely the pulse and FM-CW (frequency modulated continuous wave) schemes. The finite duration of the pulse permits range discrimination for the pulse radar at an increased average power than that of the CW signal. However, a third scheme combines the pulse and the FM: this is the chirp waveform. The duration of the chirp waveform is longer than would be required for the range discrimination, but the energy in the pulse is the same as it would be for the equivalent short pulse. This permits the same maximum range that could be achieved with a high
peak-power short pulse. Figure 3.1 shows the transmit waveform used in this system.

![Figure 3.1: Synthesized transmitted chirp waveform characteristics.](image)

The pulse duration width, $T$, is 172.4 $\mu$s; up-ramp of 86.4 $\mu$s ($216 \times 200$ ns $\times 2 = 86.4$ $\mu$s), plus the down-ramp of 215 $\mu$s ($215 \times 200$ ns $\times 2 = 86.0$ $\mu$s). The pulse repetition interval (PRI) is 2.5 ms (or equivalently, PRF = 400 Hz). The need to start the chirp 2.0 ms into the period is to ensure that the returns arrive only from the footprint illuminated on the ground at the specified aircraft altitude and incidence angle.

(b) Operating Frequency

The operating frequency of any radar system is primarily based on the function the system is to accomplish. At this preliminary stage of this SAR system, the functional goal is to simply distinguish between simultaneous slant-range targets at a determined resolution. However, ultimately this SAR will serve as an airborne-platform, remote-sensing radar for underlying terrain parameters. With this function in mind, the radar system is operated in the X-band range with a frequency band between 9.8 GHz to 10.2 GHz and a bandwidth of 400 MHz. The extent of penetration into vegetation depends upon the moisture content and density of the vegetation as well as upon the wavelength of the signal. At X-band, the wavelength is approximately 3 cm in length and roughly on the
same scale of the crop stems or vegetation it will be measuring. Since X-band signals do not penetrate into the tree canopy, it will be able to yield information about the upper layers of the vegetation and superficial layers of certain types of ground cover. Parameters that are of interest include: leaf area index (LAI), above ground biomass, and tree height.

(c) Operating Platform

To simulate the low altitude (an indication of grazing angle), low-velocity (an indication of Doppler) effects of the SAR, the radar system is mounted on a telescopic boom-mounted van. The altitude, grazing angle and velocity parameters considered by the system are restricted to the limitations imposed by this platform. Figure 3.2 shows the ground testing geometry.

![Figure 3.2: Overview of radar platform in relation to target](image)

(d) Calibration

Calibration is needed to remove measurement errors due to inherent instrumentation measurements and measurement techniques. To remove the internal system variations such as the ambient temperature fluctuations, phase change due to cable flex/length mismatches and gain drifts in the amplifiers. The DSP section will address and correct for these calibration errors during processing. Other errors, such as distortions caused by active components and antennas, will be adjusted.
for or removed when comparing the received echoes with a measured response of a known calibration target (i.e. a conducting sphere).

(e) Signal Processor

The processing of the return signal into an acceptable image is arguably one of the more complex aspects of the SAR design, following the signal generation. However, the DSP of the received echo is beyond the scope of this research and will not be considered in this thesis.

A summary of the system requirements listed in Table 3.1 are used as basic guidelines to select suitable system parameters.

Table 3.1: Summary of Design Considerations

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Design Considerations</th>
</tr>
</thead>
<tbody>
<tr>
<td>System configuration</td>
<td>Stepped-chirp FM synthesized waveform</td>
</tr>
<tr>
<td>Operating frequency</td>
<td>X-Band</td>
</tr>
<tr>
<td>Calibration</td>
<td>Internal and external measurement errors addressed in signal processing</td>
</tr>
<tr>
<td>Operating platform</td>
<td>A mobile boom-mounted van</td>
</tr>
<tr>
<td>Signal processor</td>
<td>PC-based system</td>
</tr>
</tbody>
</table>

3.1.1 Design Parameters

In Section 2.2, the radar equation derived shows how the received power is related to radar and target parameters of the system. The equation is duplicated here for convenience and is used to examine the individual parameters of the system.

\[
\text{SNR}_{\text{antenna}} = \frac{P_i G^2 \sigma \lambda^2}{(4\pi)^3 R^4 F_N k T BL} \tag{3.1}
\]
The waveform generation uses an X-band VCO that has a linear response in the required 10 GHz range and the average transmit power of 12.6 mW (+11 dBm). The operating frequency is at X-band with a wavelength ($\lambda$) of 3 cm for the reasons described in Section 3.1(b). The RF amplifier amplifies the transmitted signal ($P_t$) by +29 dBm to +40 dBm. The radar cross section (RCS) is proportionally related to the amount of reflected power diverted back towards the receiver from the illuminated area. A value of 10 dB will be used for the average RCS fluctuations ($\sigma$): this value is based on the several factors, such as terrain cover, angle of incidence, target composition, etc. [Skolnik, 2001]. The system uses two separate microstrip antenna arrays, each with virtually identical gains of 25 dBi. For a ground-based system operating at a boom-mounted van platform, the radar-to-target range ($R$) can vary between 20 meters to 100 meters.

Noise is a major factor limiting overall system performance. The SAR system design has a low noise amplifier (LNA) as the front end of the receiver. A consequence based on how the overall noise figure affects the system performance is considered. A complete measure of the noise sensitivity of the receiver takes into account the noise figures and gains of the cascaded networks; i.e. the LNA, the mixer stage, the IF amplifier stage and any losses in the RF transmission line. The receiver noise is calculated as

$$F_N = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_2G_1} + \ldots$$

with the $F_i$'s and the $G_i$'s representing the individual cascaded network losses or gains within the receiver, respectively. Finally, the remaining parameters in Equation (3.1) are considered to be constants and are system independent.

Table 3.2 summarizes the significant design parameters of the X-band SAR system.
Table 3.2: *SAR System Design Parameters*

<table>
<thead>
<tr>
<th>System Parameter</th>
<th>Selected Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>System configuration</td>
<td>Stepped-chirp FM synthesized waveform</td>
</tr>
<tr>
<td>Operating wavelength, $\lambda$</td>
<td>3 cm (X-band)</td>
</tr>
<tr>
<td>Transmit power, $P_t$</td>
<td>20 dBm – attenuated from original 10W (or 40 dBm)</td>
</tr>
<tr>
<td>Received power, $P_r$</td>
<td>-35.5 dBm (min range), -63.5 dBm (max range)</td>
</tr>
<tr>
<td>Measurement range, $R$</td>
<td>20 meters (min) to 100 meters (max)</td>
</tr>
<tr>
<td>Average RCS, $\sigma$</td>
<td>10 dB</td>
</tr>
<tr>
<td>LNA noise figure, $F_N$</td>
<td>3 dB</td>
</tr>
<tr>
<td>IF filter (center frequency)</td>
<td>30 MHz</td>
</tr>
<tr>
<td>IF bandwidth, $B$</td>
<td>7 MHz</td>
</tr>
<tr>
<td>Antenna gain, $G$</td>
<td>25 dBi</td>
</tr>
<tr>
<td>Thermal noise</td>
<td>-133.8 dBm</td>
</tr>
</tbody>
</table>

Separating Equation (3.1) into the noise and the received power considerations to realize the approximate signal to noise ratio results in:

Noise power: $P_n = KTF_NB = -103$ dBm

Received power: $P_r = \frac{PG^2\sigma\lambda^2}{(4\pi)^3R^4} = \begin{cases} -35.5 \text{ dBm} & R = 20 \text{ m} \\ -63.5 \text{ dBm} & R = 100 \text{ m} \end{cases}$

Therefore, the signal-to-noise ratio (SNR) is computed as

$$\text{SNR} = \frac{P_r}{P_n} = \begin{cases} 67.5 \text{ dB} & \text{for near reflectors} \\ 39.5 \text{ dBm} & \text{for far reflectors} \end{cases}$$
3.2 System Description

The X-band SAR design is based on a low-cost approach relying on recent advancement in digital and RF technology to develop, what is historically a large and expensive instrument, into a compact and relatively inexpensive testing instrument.

In addition to the previously stated system design parameters, the SAR design is subjected to the constraints of being built on a low budget and having a minimal impact on the delivery platform (airplane, boom-mount van, etc.). These constraints combined, dictate many of the design decisions employed into the SAR.

The schematic of the integrated X-band SAR system is shown in Figure 3.3. In general, the design may be subdivided into four major sections:

- Signal Generation
- RF section
- Receiver section
- Data Acquisition

![Figure 3.3: Top-level block schematic of SAR design.](image)
3.2.1 Signal Generation Section

The signal waveform generation is perhaps the most vital and complex section within the entire system design. The SAR operates by timing the two-way delay for a short duration RF pulse, transmitted vertically downwards and to one side of the aircraft. The required level of accuracy in range measurements (better than 4 m) calls for a pulse length of a several nanoseconds; therefore, in order to reduce the RF power requirements, a pulse compression (chirp) technique is used. The waveform can be tied directly to the slant-range resolution of the system. Additionally, the PRF, pulse width, bandwidth and the linearity of the VCO output signal are vital to the SNR of system.

An ideal chirp waveform would have a perfect ramp function of frequency versus time as depicted in Figure 3.4 (a) where at any time along the duration of the transmitted pulse \( T \) lies a unique frequency \( f_{TX} - f_{RX} = f_{IF} \), called the intermediate frequency.

![Figure 3.4: (a) Ideal LFM Waveform. (b) Actual LFM Waveform.](image)

The ideal LFM ramp could be generated by a very expensive and sizeable X-band synthesizer. However, with the constraints imposed on the design of low-cost and having a small footprint, a solution of generating the waveform using a D/A computer card and an X-band VCO was implemented. In Figure 3.4 (b), the actual
waveform generated from the D/A-VCO combination illustrates a *stair-step* ramp. It is important that each vertical step length be reduced as to approach a linear slope. The *stair-step* emerges as a result of the limitations of the generation rate (Rg) and the output bit-resolution of the D/A card, and the inherent settling time of the VCO.

The range resolution in Figure 3.4(a) is calculated as

$$\delta_r = \frac{c}{2B}$$

(3.4)

when the LFM has a linear slope, or the alternative equation

$$\delta_r = \frac{c\tau}{2}$$

(3.5)

can be used when the solving for a stair-step slope as in Figure 3.4 (b). In Figure 3.4 (a), the pulse duration of chirp is equal to the reciprocal of the bandwidth, i.e. $T = 1/B$ (or equivalently; $\tau = 1/B$). However, in Figure 3.4 (b), the pulse width $\tau$ is no longer equal to the reciprocal of the bandwidth but is now equal to the width of horizontal step in the stair-stepped LFM ($T \neq t$ and $\tau \neq 1/B$).

Table 3.3 lists the range resolution results from both equations.

### Table 3.3: Range Resolutions

<table>
<thead>
<tr>
<th>Equation 3.5 (stair-step slope)</th>
<th>Rg (Megasample/sec)</th>
<th>T ($\tau = 1/Rg$)</th>
<th>Range Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>80 Ms/s</td>
<td>12.5 ns</td>
<td>1.88 meters</td>
</tr>
<tr>
<td></td>
<td>40 Ms/s</td>
<td>25.0 ns</td>
<td>3.75 meters</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Equation 3.4 (LFM slope)</th>
<th>Bandwidth (B)</th>
<th>Pulse Width (T)</th>
<th>Range Resolution</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>400 MHz</td>
<td>2.5 ns</td>
<td>0.375 meter</td>
</tr>
</tbody>
</table>
Since the HRR system design will serve as the hardware platform for the X-band airborne SAR, certain compromises must be made at this stage between incorporating the airborne-platform SAR parameters (e.g. increased range-to-target) versus attempting to design for the greatest range resolution of a ground-based system. For an airborne-platform, the designated range-to-target is 750 meters, which has the optimal pulse duration of the chirp of 102.3875 μs (as explained in the Receiver Section of this document). The following table establishes the relationship between the output bit resolution of the D/A, the generation rate of the D/A card and the associated pulse duration.

<table>
<thead>
<tr>
<th></th>
<th>D/A Bit Resolution</th>
<th>Generation Rate (Rg)</th>
<th>Pulse Duration (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SAR design</td>
<td>216 bits</td>
<td>5 Ms/s</td>
<td>172.8 μs</td>
</tr>
<tr>
<td>Current design</td>
<td>216 bits</td>
<td>80 Ms/s</td>
<td>10.8 μs</td>
</tr>
<tr>
<td>With extender circuit</td>
<td>4096 bits</td>
<td>80 Ms/s</td>
<td>51.2 μs</td>
</tr>
</tbody>
</table>

The original design of the waveform generation was specifically designed to meet the parameter criterion for an airborne platform SAR. The design did not take advantage of the full output range of the D/A and was restricted to only 216 bits. This restriction of 216 bits required the pulse duration time to be significantly large to accommodate the greater range-to-target distance; therefore a low generation rate of 5 MHz is required. The 216-bit limitation results from difference in the bit representation of the output voltage from the D/A component (i.e. D/A bit 800 = 6.01 V = 10.2 GHz and D/A bit 584 = 6.99 V = 9.8 GHz; 800 - 584 = 216 bits). However, since the HRR radar is tested on the ground and at a considerable shorter range-to-target distance, a higher generation rate of 80 MHz can be considered. This higher rate will
provide a better range resolution for a shorter range. Again, this current design is still limited to the 216-bit limitation as explained with the SAR design. However, to gain a much finer range resolution with the existing hardware, an analog "extender" circuit (schematic in Appendix A.3) is included into this design and thereby taking advantage of the entire 4096-bit (216 bits = 4096) range from the output of the D/A card as illustrated in Figure 3.5.

![Diagram of D/A output resolution](image)

Figure 3.5: D/A Output Bit Resolution.

The greater number of bits representing the same voltage range (6.01 V to 6.99 V) of the D/A output would equate to a smaller voltage increment feeding the varactor inside the VCO. The 4096-bit analog output of the D/A card is combined with the precision steady state analog voltage of the extender digital circuitry, to provide the required input to the VCO. The input analog feed to the VCO will provide a very precisely correlated voltage-input to frequency-output synthesizing the desired LFM waveform. The frequency on the VCO is tuned (modulated) by varying the applied voltage from the D/A-extender circuit combination. The tuning characteristics of the VCO are provided in Appendix A.1. The extender analog circuit addresses two separate problems: (a) the settling time parameter of the VCO is excessively large (~150 nsec), and (b) the output bit resolution of the D/A card was restricted to 216 bits. By extending the output bit resolution of the D/A card, the incremental analog
voltage input to the VCO was dramatically reduced, thereby preventing the internal varactor of the VCO from fully charging and discharging. This charging/discharging corresponds directly to the settling time of the VCO. Changing the analog input voltage to the VCO to smaller increments changes the charge level on the varactor only minimally and in turn, improves the recovery time of the VCO considerably. Increasing the recovery time of the varactor reduces the settling time of the VCO and provides for a more linear ramp to the waveform. If more time and resources were available to improve on the analog design, it is believed a greater gain would result in the range resolution from such a small contribution. It will be left up to the next generation of hardware to improve on this circuit.

Figure 3.6 summarizes the synthesized LFM-chirp waveform;

![Figure 3.6: Waveform parameters.](image)

3.2.2 Radio Frequency (RF) Section

The RF section consists of the aforementioned signal generated section, the power divider and the high power RF amplifier. Setting aside the signal generation, the RF section is fairly straightforward. The power divider receives the generated waveform,
where the signal is reduced by 3 dB at each exit port. The signal at one port is used as a local oscillator to be mixed with the received echo return. The signal at the remaining port is passed onto the RF amplifier where the power level is increased by 29 dB. Although the maximum RF amplifier gain is rated at +40 dB, an 11-dB attenuator at the output port limits the amplifier gain to +29 dBm. This prevents saturation of the receiver LNA for shorter-range operation. The full amplification of the RF amplifier will be required in the airborne SAR configuration, where the range to target is increased dramatically. After the power level of the signal has been amplified by the RF amplifier, the synthesized LFM chirp signal is then coupled into free space via the transmitting antenna.

### 3.2.3 Receiver Section

The main function of the receiver section (or IF section) is to filter and amplify the return echo signal. The receiver antenna receives the re-radiated echo returns from the target, which are at a considerable lower power level than when the signal left the transmitter. The LNA amplifies the return signal by 30 dB to elevate the weak signal to required SNR levels. At this stage, the signal is at RF and is rather difficult and expensive to manipulate. The high frequency signal is down-converted into an intermediate frequency (IF) signal by a mixing process. The single side band (SSB) mixer, a down-converter, outputs the difference in frequency between the copy of the transmitted signal and the received echo return: $|f_{LO} - f_{RF}| = f_{IF}$.

Referring to the radar Equation (2.5), the received power decreases at a rate proportional to $R^4$, which means that if the range-to-target is doubled, the received power will decrease by a factor of 16. As the range increases, the maximum received power decreases but the minimum detectable power level remains the same. This is due to the fact that the thermal noise level is independent of range. Therefore,
the overall effect is the decrease in the system dynamic range. This decrease in the system dynamic range can be avoided by employing a variable gain amplifier in the IF stage. The AGC (automatic gain control) of the IF amplifier is adjusted via an external potentiometer, to accommodate the dynamic range the radar. The center frequency of the IF amplifier is at 30 MHz with a bandwidth of 10 MHz. The noise figure of this amplifier is 4 dB, which is still greater than that of the front-end LNA, and although the amplifier has the maximum gain of 70 dB, minimum gain will be used during ground testing. The last stage of the IF section is band pass filter (BPF). The BPF is inserted after the IF amplifier to reject any unwanted signal outside the passband (30 MHz ± 3.5 MHz). The IF frequency is directly proportional to the range, as can be seen the following equation

\[ f_{IF} = \frac{2BR}{cT} \]  

where \( B \) is the system bandwidth, \( T \) is the pulse duration time, \( c \) is the speed of light and \( R \) is the range-to-target. On the airborne SAR platform, the estimated range from radar to target is 1000 m, the pulse duration is 102.4 μs and the system bandwidth is 400 MHz. Substituting these values for the parameters in Equation 3.6 leads to the IF frequency of 30 MHz. To determine the bandwidth of the IF frequency, which directly determines the 3-dB cutoff requirement of the BPF, the following equation is used

\[ \Delta f_{IF} = \frac{2B\Delta W}{cT} \]  

where \( \Delta W \) is the swath width on the ground as determined in Equation (2.21). In the SAR configuration, the swath width is 100 meters, which would bring the IF bandwidth to 7 MHz. The output of the band pass filter is then fed to the data acquisition section for further processing.
3.2.4 Antenna Section

The transmitter output is coupled to the transmitting antenna where the signal traverses through the atmosphere to be scattered by a target and re-radiates back to the receiving antenna.

Table 3.5 summarizes the antenna specification.

<table>
<thead>
<tr>
<th>Item</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>9.8 GHz - 10.2 GHz</td>
</tr>
<tr>
<td>Radiating element</td>
<td>Printed Circuit Patch Array</td>
</tr>
<tr>
<td>Polarization</td>
<td>Linearly polarized</td>
</tr>
<tr>
<td>Gain</td>
<td>25 dBi</td>
</tr>
<tr>
<td>Return loss</td>
<td>-10.25 dB</td>
</tr>
<tr>
<td>Radiation pattern</td>
<td>Pencil Beam</td>
</tr>
<tr>
<td>3-dB beamwidth</td>
<td>8°</td>
</tr>
<tr>
<td>Main/Side lobe ratio</td>
<td>19.6 dB</td>
</tr>
<tr>
<td>Dimensions</td>
<td>15&quot; x 8&quot; (V x H)</td>
</tr>
<tr>
<td>Electrical connection</td>
<td>SMA female</td>
</tr>
</tbody>
</table>

The transmit and receive antennas assume the SLAR (side looking airborne radar) setup configuration, that is, they will be pointed down and to one side of the aircraft at a predetermined look angle. The transmit antenna produces a beam width that is wide vertically (8°) and narrow horizontally (5.5°). The narrow beam width is required to preserve the angular information concerning the scattering coefficient of the distributed target. However, if the beam width is too small, the illuminated footprint might not have sufficient independent scatterers to represent the target (Ulaby
et al, 1982). On average, the power level figures (in Appendix A.2) show the first side lobe level (SLL) from the main lobe is 19.6 dB and indicate that the return loss is 10.25 dB at the center frequency.

The airborne antenna system must be lightweight and have a relatively small, flat, streamlined physical profile to be considered practical. The microstrip patch array best matches the required prerequisites of having the antennas mounted on the outside of the fuselage of the aircraft. The receive and transmit antennas are identical in every aspect and can be used interchangeably.

3.2.5 Data Acquisition Section

The data acquisition section consists of an A/D computer card and the storage system on the computer itself. The function of this section is to simply convert the IF frequency analog signals into strings of digital data and store this information into a memory device. Referring to the Nyquist sampling theorem, the sampling frequency must be greater than twice the maximum IF frequency, $f_{IF_{max}}$

$$f_s > 2f_{IF_{max}}$$ (3.8)

The D/A card has a maximum sampling rate of 500 Ms/s, and since $f_{IF}$ was calculated from Equation (3.6) to be 30 MHz, this would imply the analog IF signal could be properly recorded without aliasing. In Equation (3.6), all the parameters, with the exception of the range ($R$), are fixed by the system. Since the range was selected as 1 km as the maximum range allowed for this SAR, therefore the maximum IF frequency cannot exceed 30 MHz. An ideal design would have the D/A and the A/D cards to have matched sampling rates, for every one bit generated from the D/A, the A/D would be able to sample only one bit. If the A/D sampling rate is greater than the D/A generation rate, harmonics would appear during DSP and there would
be a risk of exceeding the on-board memory of the A/D card by gathering too much information. Additionally, if too much data were gathered, there would be risk in exceeding the maximum data transfer rate between the A/D card to the computer storage disk, resulting in lost data. On the other hand, the penalty of having the sampling rate of the A/D card being less than that of the D/A card would be a loss of data. Since oversampling is inevitable with this system, the DSP can sample-average the data gathered with the A/D. This is accomplished by averaging 20 to 60 received pulsewidth returned signals. This averaging of the return signals reduces system noise, but more over, it reduces the "environmental" noise. This is the noise which is random and received from the surrounding terrain environment (i.e. target shifts, wind gusts, clutter movement, etc.).

Referring to Figure 3.1, it is necessary to have a lead-time of delay from when the period starts to when the chirp pulse begins. The lead-time accounts for a minimum time the VCO needs to stabilize before constructing the LFM pulse, also for an inherent delay between the D/A and the A/D, and for the two-way travel time to the target. It is also during this time period which the system transfers the data from the on-board memory of the A/D card to the storage unit in the computer.

There is an on-board processor on each of the D/A and the A/D cards, as such; there is a mismatch in starting times (~ 100 nsec) from when the each card receives the timing signal to when they start functioning. This mismatch is accounted for during the trailing end where there is a dead period (after the chirp pulse) of the waveform in Figure 3.1.
Chapter 4

Analysis and Results

4.1 Laboratory Test Results

Laboratory testing involves verifying, under controlled conditions that the RF and digital hardware of the HRR is indeed functional within the parameters and specifications outlined. Test procedures will verify the entire HRR radar design as a single unit-under-test (UUT), as opposed to verifying each individual system component as a stand-alone system. The UUT will consist of the four subsections as outlined in Section 3.2: the waveform generation, RF, Receiver and Data Acquisition. The antenna subsection will be bypassed during the laboratory-testing phase, but will be tested and verified during field-testing. In the sections to follow, many of the same procedures designed to test the compatibility of the individual component are used to test and debug the inherent problems of system design. Therefore, rather than testing and reporting duplicate results, the test procedures and plots to follow will focus on identifying the shortcomings of the HRR radar system.

Testing identified two significant problems which appeared in the systems: (a) identification of the low-frequency induced interference and implementation of a solution to overcome this problem, and (b) identification of the signal IF gain-saturation in the Receiver and Data Acquisition sections and provision of general guidelines to obtain and process valid data.
In most of the test procedures, except where specifically noted, the output bandwidth of the VCO will be limited to 40 MHz, rather than the full bandwidth of 400 MHz. The 10% bandwidth restriction is a response to the A/D sampling rate limitation of 100 MHz. To avoid aliasing, the 40 MHz bandwidth is restricted to less than the aliasing frequency of 50 MHz \((f_s \geq 2f_{max})\).

4.1.1 IF Gain-Saturation Problem

Once the RF components and digital circuitry are selected as the framework of the system design, the system boundaries must be tested from which the complete system will operate properly. One of the more significant problems of the HRR radar design deals with maintaining a proper power level in the IF section of the design. The power level of the received echo, at the receiver antenna, must be higher than the noise level of the receiver in order to distinguish the target signal from the system noise. The LNA will amplify the receive echo and the mixer downconverter will output the IF. The power level of the IF range is adjusted by the AGC in the IF amplifier. The power level of the signal available after the IF amplifier is the focus of this subsection. If the power level is too high, then the input to the A/D component will become saturated and the acquired data will be incomplete and invalid. On the other hand, if the power level is too low, the A/D will sample noise and the data will be corrupted and invalid.

4.1.1.1 LFM Transmitted Signal and Power Spectrum (Saturation UUT)

Figure 4.1 represents the unit-under-test for this sequence of testing. This UUT setup tests the functional compatibility of the transmit and receive channels of the system design, as well as the data acquisition section. With the limited bandwidth of 40 MHz, the combination D/A - VCO generates the transmit chirp waveform of approximately 9.8 GHz to 9.84 GHz.
Figure 4.1: UUT- Testing LFM, Transmit, Receive and Data Acquisition.

With the coarse resolution of the two combined components allowing for a minor variance in accuracy of starting/stopping at a specific frequency, the output range might look something like 9.785 GHz to 9.832 GHz (as an example). The inputs to the mixer are the LFM from the VCO and a constant generated signal of 9.8 GHz from the frequency synthesizer. The internal frequency bandpass range of the mixer is 10 MHz to 500 MHz, so the output is a linear frequency signal of ~ 10 MHz to 40 MHz. An attenuation of 40 dB is provided at the output of the power divider so as to limit the input power to the LNA. The attenuation simulates the signal strength from a down-range target. The input voltage level of the A/D component is set at 200 mV, which is related to noise sensitivity of the acquired data.

In Figure 4.2, the top plot presents a time-domain signal as acquired by the A/D, at a sampling rate of 100 MHz, and the bottom plot represents the Fast Fourier Transform (FFT) of the signal representing the power spectrum density (PSD) of the same signal.

It is clear from the time domain plot, the acquired data reveals a LFM waveform. Point 1 represents the lowest frequency of the up-ramp, point 2 is the highest frequency of the triangular waveform, and point 3 is the lowest frequency of the down-ramp.
Figure 4.2: VCO-50dB-LNA-9.8GHz-Mixer-200mV.

It would be ideal for the lowest frequency, points (1) and (2) to reside at the beginning of sampling and at the end of sampling the data, respectively. However, due to inability of the chirp generation section to generate an exact 9.8 GHz to 9.84 GHz, the mixer will output the difference between the static 9.8 GHz and the LFM from the VCO.

The PSD plot shows the bandpass frequencies starting at baseband and cutting off around 32 MHz. This plot should ideally display a bandpass frequency range of ~10 MHz to 40 MHz. There are two reasons for this disparity, the first is explained by chirp generation section limitations of providing exact start/stop frequencies, and the second is due to the internal passband of the mixer. The passband specification of the mixer indicates a range of 10 MHz to 500 MHz, however, the 3-dB cutoff of this BPF could allow for slightly more, or less, signal to pass through the passband.
The following plots in Figure 4.3 are included to illustrate the importance of the attenuation level (i.e. simulative of target strength) of the transmitted signal and the input voltage level of the A/D. Using the same UUT setup, increase the attenuator to 60 dB and lower the input voltage level to the A/D component.

![Acquired Data](image1.png)

![Power Spectrum Density](image2.png)

Figure 4.3: VCO-70dB-LNA-9.8GHz-Mixer-50mV.

In Figure 4.3, the time domain plot exhibits a lower overall voltage level in the acquired data, as well as the waveform appearing to be coarser than previous. This fact is explained by the LNA receiving weaker signal strength (through added attenuation). The PSD plot exhibits a lower SNR level compared to the previous plot. With the input voltage level decrease on the A/D, the threshold level of the input signal is limited while the noise level is allowed to rise. This effect reduces the dynamic range of the input signal in reference to the noise floor, thereby lowering the SNR. Although not as clear as in the previous PSD plot, a passband is still visible with a similar cutoff frequency.
The conclusion interpreted from both sets of plots provides verification for the functionality and interface of each component in the UUT. In addition, it points out the effects of the attenuation level to the LNA and the input voltage level to the A/D component.

4.1.1.2 Saturation UUT - Include The IF BPF

Figure 4.4 represents the unit-under-test for this sequence of testing. This UUT setup includes the intermediate frequency bandpass filter. The UUT will test the IF passband boundaries of the system design.

![Diagram](image)

Figure 4.4: UUT-Testing IF passband.

The UUT setup is identical to Figure 4.1, with the addition of the IF BPF, having a center frequency of 26 MHz and a bandwidth of 7 MHz. The input to the BPF is 10 MHz to 50 MHz from the downconverter mixer, and the output frequency range of the BPF is 22.5 MHz to 29.5 MHz. Figure 4.5 shows plots of the A/D acquired data in the time domain, and the FFT of the sampled data representing its power spectrum.

The frequency range of the generated chirp waveform is ~ 9.8 GHz to 9.84 GHz and the static synthesized frequency is at 9.79 GHz. Shifting the synthesized frequency to 9.79 GHz allows the passband of the downconverter to start at 10 MHz,
versus at baseband. All other system parameters and components in the UUT are the same as in the system tested in Section 4.1.1.1.

Figure 4.5: VCO-50dB-LNA-9.79GHz-Mixer-BPF-100mV.

The acquired data in the time domain plot shows a compact signal of the up-ramp and the down-ramp of the transmitted triangular waveform. This time domain plot shows how the BPF restricts only 7 MHz bandwidth of the 40 MHz bandpass signal through the IF section of the system. The encircled area on the down-ramp section of the time domain plot displays some deviation from that of the up-ramp section, ideally it would be expected these two samplings would be similar (but inverted) to each other. This deviation is explained by the inclusion of the inherent system noise and due to non-ideal hardware corruption. Several samples were acquired with very similar results of the up-ramp and down-ramp differing slightly from each other.

In the PSD plot, it is fairly evident the passband is approximately 22.5 MHz to 29.5 MHz, indicating a bandwidth of 7 MHz.
The conclusions interpreted from this set of plots provides a high confidence level for the integration of the IF BPF into the system design.

4.1.1.3 Saturation UUT - Include The IF Amplifier

Figure 4.6 represents the unit-under-test for this sequence of testing. This UUT setup includes the intermediate frequency amplifier with automatic gain control (AGC). The UUT will test the effects of gain-saturation levels from the IF amplifier to the system design.

![Diagram of UUT with IF Amplifier and AGC Saturation Level](image)

**Figure 4.6: UUT-Testing IF Amplifier and AGC Saturation Level.**

The UUT setup is identical to Figure 4.1, with the addition of the IF amplifier having a center frequency of 30 MHz and a bandwidth of 10 MHz. Controlling the saturation level of the IF amplifier output is essential to maintain a proper voltage level to the input of the A/D component to adequately sample the data. The AGC of the IF amplifier has a dynamic range of 70 dB. The amplifier will exhibit no gain when there is a -4 V at the input AGC port, in turn, the amplifier will boost the received signal the full 70 dB when the AGC port is either grounded or there is no input. The amplifier will not differentiate between a valid signal and noise, rather it will amplify whatever the signal is present at its input port. The input frequency to the amplifier is 20 MHz to 60 MHz from the mixer downconverter, and the output...
is 25 MHz to 35 MHz due to the internal passband of the IF amplifier. All other system parameters and components in the UUT are the same as in the system tested in Section 4.1.1.1.

Since a ground to the AGC port boosted the input from the IF amplifier 70 dB, the input voltage level of the A/D component was raised to the maximum 10 volts. This test demonstrates the need to keep the AGC gain adjusted to proper levels relative to the signal strength of the received signal from the target. If the receiver antenna receives a strong signal from a high reflectivity target or a close-in target, the incoming signal will be strong and the need to reduce the AGC gain to avoid saturating the input to the A/D component is essential. The time domain plot in Figure 4.7 shows a signal that is clipped at the top and bottom, resulting from the applied 70-dB AGC gain and thereby saturating the A/D input. As a result of this
saturation, the power spectrum in the PSD plot is unable to clearly show the proper passband SNR.

After lowering the AGC gain below the maximum level, the time domain plot in Figure 4.8 displays the sampled signal that is not saturated and the PSD plot of a power spectrum with a clear passband.

![Acquired Data](image1)

![Power Spectrum Density](image2)

**Figure 4.8: VCO-50dB-LNA-9.78GHz-Mixer-IF-AGC2v-10v(Gain < 70dB).**

The PSD plot exhibits a passband of approximately 20 MHz to 45 MHz, which is slightly higher than the internal passband of the IF amplifier (25 MHz 35 MHz). As previously stated, this irregularity can be explained by the 3-dB cutoff of the internal BPF of the IF amplifier.

### 4.1.1.4 Saturation UUT - Include The LNA And IF Amplifier

Figure 4.9 represents the unit-under-test for this sequence of testing. This UUT setup represents the complete system design. The UUT will test the compatibility of all
the RF/digital components operating as a complete system.

The UUT setup is tested with the 10% limited bandwidth of 40 MHz and all RF components and digital circuitry included completing the system design (minus the antennas). The AGC port of the IF amplifier was biased to a proper gain level (< 70 dB) and the input voltage level of the A/D card was selected to the maximum level of 10 volts.

Referring to Figure 4.10, the time domain plot exhibits a non-saturated signal sampled from the A/D component. The distortion observed in the down-ramp of the time domain plot can be caused by system noise and non-ideal hardware corruption. The PSD plot exhibits a satisfactory SNR with the proper passband frequency range as expected from the IF BPF. Overall, these plots illustrate how the complete system operates within the prescribed specifications of the HRR.
In HRR radar systems, operating in the microwave region, it is common to find residuals of undesired low frequencies. These residuals become problematic since they induce harmonics within the system, and they also extend the minimum range of the close-in target where the system can begin to acquire valid data. These low frequencies usually reside in the kilohertz range, which can be on the same scale of magnitude of the IF range, and thereby inducing itself (or its harmonics) along with valid data.

In Section 3.2.3, the relationship between the IF frequency and range was explained with Equation (3.6). It is repeated here for convenience.
The system variables $B$, $c$ and $T$ are grouped as a single constant $\alpha$, leaving the following equation.

$$f_{IF} = \frac{2BR}{cT}$$

(4.1)

From Equation (4.2), it is easy to see that $f_{IF}$ is directly proportional to range. When the IF frequency is corrupted with low-frequency effects, there is a proportional range (distance) which the system will be unable to use as valid data, thereby placing a higher limit on a minimum range the system can begin to detect a target.

4.1.2.1 Identifying Low Frequency (Low Frequency UUT)

Figure 4.11 represents the unit-under-test for this sequence of testing. This UUT setup test will identify the low frequency and locate the origin of where it is being induced into the system. The UUT consists of the minimum required RF components to generate a 40-MHz bandwidth chirp-waveform and simultaneously acquire the IF through sampling with the A/D component.

![Diagram of UUT setup test](image)

Figure 4.11: UUT- Testing the Low Frequency System Component.

The power divider provides the same generated chirp signal to both inputs of the mixer downconverter; therefore the mixer IF is expected to output at baseband.
Figure 4.12: VCO-BW40-PD-Mixer-2V.

Figure 4.13: VCO-BW40-PD-Mixer-20mV.
The two interconnecting cables between the power divider to the mixer are of the
same type, have the same connectors and are virtually the same length. Having
identical cables reduces the possibility of inducing phase mismatch (via reflections)
at the component ports and indirectly producing system noise. Two set of plots are
listed in Figures 4.12 and 4.13, the differentiating factor between each set is the input
temperature level for A/D sampling.

The time domain plots in Figures 4.12 and 4.13 indicate the presence of a sta­
ble low frequency signal. The frequency of this signal is $1/((216+215)*400 \text{ ns}) \approx$
6 kHz. There are two significant factors contributing to the low frequency; the dis­
crete step intervals from the D/A to the VCO generating the transmit waveform,
and the impedance mismatches (i.e. reflections) at the interface between RF system
components (i.e. the RF interface).

More specifically, the low frequency interference originates at the VCO due to the
discrete step output frequency versus a linear ramp frequency output and including
the cross coupling between the two ports of a non-ideal power divider. It is not
the RF alone that generates the low frequency, rather the RF interface, which adds
constructively or destructively to the system. A basis for this conclusion resides in
the fact that noise is random and virtually impossible to replicate, whereas the RF
interface will produce similar results each time. This reasoning supports why similar
(but not exact) power spectra near baseband are observed in each of the PSD plots.
The PSD plots from Figures 4.12 and 4.13, indicate no higher frequency components,
except close to DC. The power density observed near baseband, is a product of the
time delay invoked by mismatched cables between the power divider and the mixer.
Several measurements were taken with differing mismatched cable lengths and each
displayed a shifted power spectra near baseband.
The plots in Figure 4.14 involve widening the transmission bandwidth to its full range. By expanding the testing procedure to include the entire bandwidth of 400 MHz, the test will demonstrate similar outcomes as with the 10% bandwidth. More importantly, this broader testing profile will verify that the system's hardware and software do function correctly under normal operating system conditions.

![Graph showing acquired data and power spectrum density](image)

Figure 4.14: VCO-BW40-PD-Mixer-20mV.

The time domain plot continues to exhibit a low frequency component under the full system bandwidth. The frequency is different than the previous test which can be attributed to the coarse resolution of the D/A discrete step resolution and aliasing at the A/D component.

4.1.2.2 Low Frequency UUT Changing the A/D Sampling Rate

Figure 4.15 represents the unit-under-test for this sequence of testing. This is the same UUT configuration that was used in the previous Subsection 4.1.2.1 and will
test what effect the A/D component will have on the low frequency interference within the system. The difference in this configuration from the previous configuration is the increase of the sampling rate of the A/D component from 100 MHz to 400 MHz.

Figure 4.15 displays the time domain plot exhibiting no discernable difference from the time domain plot in Figure 4.13, which is sampling at the system default rate of 100 MHz.

Therefore, the conclusion to exclude the A/D component as a contributor to the induced low frequency is valid.

4.1.2.3 Low Frequency UUT - Including the IF BPF

Figure 4.16 represents the unit-under-test for this sequence of testing. This UUT setup test will include the IF bandpass filter to confirm that the low-frequency effect will be filtered out before the A/D begins sampling data. The UUT will be tested
with the chirp bandwidth of 40 MHz (9.8 GHz - 9.84 GHz) and then will be repeated with the full bandwidth of 400 MHz (9.8 GHz - 10.2 GHz).

In the previous Subsection 4.1.2.1, a low frequency component was identified to be present in the system as early as in the signal generation section of the design.

This UUT has included the IF BPF to filter out this low frequency, as well as those frequency components due from the power divider/mixer cable mismatch. The plots of Figure 4.17 and Figure 4.18 represent the sampled data with a system bandwidth of 40 MHz and 400 MHz, respectively. In both of the time domain plots in Figure 4.17 and 4.18, there is no valid signal. The signals observed in both cases are very small and are attributed to system noise. The PSD plots both exhibit various spikes in the power spectrum, but they are random and possess no significant value. Therefore, the conclusion taken from this test confirms the IF bandpass filter can satisfactorily eliminate the low frequency components.
Figure 4.17: VCO-BW40-PD-Mixer-BPF-50mV.

Figure 4.18: VCO-BW400-PD-Mixer-BPF-50mV.
4.1.2.4 Low Frequency UUT – Including the RF Amplifier

Figure 4.19 represents the unit-under-test for this sequence of testing. This UUT setup test will include the RF amplifier to confirm the effects of this system component will not be additive to the low frequency problem. The RF amplifier will boost the transmitted signal to +40 dBm, and connected to the output will be a 40 dB attenuator to lower the amplified signal, thereby avoiding damage to the mixer.

![Diagram of UUT testing the effect of the RF Amplifier](image)

Figure 4.19: UUT Testing the Effect of the RF Amplifier.

This configuration tests what effect the addition of the RF amplifier will have on the system. The UUT is tested with the system bandwidth limited to 40 MHz.

Even with the addition of the RF amplifier and attenuator, the time domain plot in Figure 4.20 displays how the low-frequency effect is still present and is quite similar to the UUT without the amplifier. The PSD plot exhibits no valid power spectra, except close to baseband, which has previously been explained as the cable mismatch between the power divider and the mixer. Therefore, it is safe to conclude the RF amplifier contributes nothing to the low-frequency component.
4.1.2.5 Low Frequency UUT – Normal System Configuration

Figure 4.21 represents the unit-under-test for this sequence of testing. This UUT setup represents the complete system design, minus the antennas. The UUT tests whether the low frequency will have any effect on the overall system design and expresses the relationship between the input voltage level of the A/D component and noise sensitivity.

This configuration pools together the test conclusions resulting from the previous subsections regarding the low frequency problem. The UUT is tested with the full system bandwidth of 400 MHz with the AGC port biased reducing the maximum gain in the IF amplifier. The following sets of plots result from the same testing.
procedure; however, the one variable factor is the input voltage level to the A/D component. The time domain plots in the Figures 4.22 through 4.25 show the A/D sampling virtually zero input signal. This concludes the system as a whole does filter out any residuals of the low frequency. On the other hand, the PSD plots illustrate the effects of selecting a proper input voltage level to the sampling A/D component.

Figure 4.21: Testing Normal System Configuration.

Figure 4.22: VCO-BW400-PD-50dB-LNA-MIXER-IF(AGC3v)-BPF-2v.
Figure 4.23: VCO-BW400-PD-50dB-LNA-MIXER-IF(AGC3v)-BPF-1v.

Figure 4.24: VCO-BW400-PD-50dB-LNA-MIXER-IF(AGC3v)-BPF-500mv.
It is evident in Figures 4.22 - 4.25, the PSD plots display different power spectra as the input voltage levels decrease from 2 volts to 200 millivolts. At higher input levels, the A/D samples a wider dynamic range of power spectra above the noise floor. This is evident in Figure 4.22, with the input level at 2 volts, the PSD shows random spikes in the power spectrum that originate from the discrete output steps at the VCO. In Figure 4.24, as the selected input level lowers to 500 millivolts, a much narrower dynamic range is sampled by the A/D and evidence of noise begins to appear in the PSD plot. When the input voltage level is lowered to 200 millivolts, the power level of the signal is on the same order of magnitude as the noise floor and the PSD plot in Figure 4.25 displays band-limited amplified noise. At these lower voltage levels, the A/D component becomes quite sensitive to detecting the low level signals and care must be taken not to mistake the power spectrum of band-limited noise as
a valid low-level signal. Of the four sets of plots, Figure 4.23 shows the best selection of input voltage level to be at 1 V for the conditions given in this UUT. Here, the time domain plot exhibits a virtual zero input signal and the PSD shows no power spectrum.

4.1.3 Laboratory Testing Conclusions

In the previous subsections, the HRR radar system design was tested to identify and understand the inherent problems associated when RF components, digital hardware and controlling system software interface one another at X-band frequencies. Testing to understand the relationship between target reflectivity, return echo strength and signal saturation is essential to aid interpreting acquired data and avoid acquiring erroneous data. The AGC control governs the level of gain the IF amplifier would boost the incoming signal. Too much gain from the IF amplifier would saturate a valid input signal to the A/D component, whereas too little gain would not contribute to a proper level of signal-to-noise ratio. The level of gain is dependent on the power level of the signal return and should be adjusted according to the environment or conditions the measurements are taken. Similarly, identifying the low frequency component in this system design is essential to understanding the boundaries or limitations the HRR radar can operate within. There were several factors identified contributing to the most significant amount of noise to the system; the D/A-VCO discrete step interval waveform generation, the RF interface between RF components resulting in reflections, and cable mismatches adding time delays in certain signal paths. The low frequency places a limitation on close-range measurements ($f_{IF} \propto R$), i.e. the minimum distance the radar can illuminate a target and acquire valid data. The inherent low frequency was filtered out via the IF bandpass filter. Lastly, the A/D component was tested and a relationship was identified between input voltage levels and the sensitivity to low-level signals. Higher input voltage levels are
necessary when there exist higher power levels of the signal being sampled. As the
input voltage levels become higher, also comes a wider dynamic range in the power
spectrum to be sampled. When the input voltage level decreases, the narrower the
dynamic range becomes and the A/D becomes more sensitive to noise. Similar to the
conditions for AGC in the IF amplifier, the input voltage level to the A/D component
should be properly adjusted to the system configuration and environment in which
the measurements are made.

4.2 HRR Radar Data Processing

The data acquired by the A/D converter is related to the reflectivity of the target as
received by the radar receiver. This data are stored as a raw digital format that cannot
provide intelligible information in its current form. Therefore, some manipulation to
this data format is required to interpret what the radar is seeing downrange. The
data reformatting takes place in software data processing. There are two software
applications that are used to acquire and process the HRR radar data; these are the
LabView software interface program and Matlab, a mathematical software program.
LabView is the graphical user interface used to adjust system parameters, on-the-fly,
while acquiring and storing the data onto a storage media. Matlab will use the stored
data to provide digital filtering, pulse averaging, Fast Fourier Transforms and produce
both time domain, and PSD plots.

4.2.1 Data Structure

The generation of transmitted waveform is described from a hardware viewpoint in
Section 3.2.1. However, the data structure of the transmit and receive waveforms is a
more detailed description of generating data and acquiring data from a point/sample
viewpoint.
4.2.1.1 Generating Pulse

The data structure of the transmitted and received waveforms can be explained by referring to Figure 3.1 from the previous chapter. There is an elapsed time from the beginning of the transmission to the beginning of the LFM chirp of 2 ms. This elapsed time is necessary to account for the stabilization time of the VCO, the intrinsic time delay between the A/D and the D/A components and the two-way round trip time delay of the transmitted signal. The number of points for this elapsed time is $2 \text{ ms} \times 5 \text{ MHz} = 10,000$ points from the D/A. Figure 4.26 shows the parameters for the generation of the LFM chirp.

![Diagram of LFM Chirp waveform](image)

Figure 4.26: Parameters of the LFM Chirp waveform.

The received waveform is a virtual replica of the transmitted waveform, with a time delay of $\Delta t = td = 2R/c$, equating to the two-way round trip time the transmitted wave takes to propagate to the target and return to the radar receiver. The D/A converter generates samples at a rate of 5 MHz (200 ns). The RF VCO does not use the full resolution of its output range, rather only those points between 584 and 800,
equating to 10.2 GHz and 9.8 GHz, respectively. This difference is $800 - 584 = 216$ steps from the D/A, and since there is an up-ramp and a down-ramp to the transmitted wave, the initial number of steps to create the up-/down-ramps are $216 + 215 = 431$ steps. However, since each step consists of two generated sample points, the total number of generated sample points for the waveform is $431 \times 2 = 862$ points. As a result of the internal time delay of the D/A, this component cannot be relied upon to terminate the waveform transmission at a consistent end point every time. This is the reasoning for the arbitrary dead time sometime after the end of the down-ramp and the start of the following period. Summarizing the data structure of a complete period of the transmitting waveform is given in Figure 4.27.

![Figure 4.27: Point Data Structure of the Transmitted Waveform.](image)

4.2.1.2 Acquiring Data

The A/D converter is responsible for acquiring the return data from the receiver in the form of sampling this data. The sampling rate of the A/D is set for 100 MHz. Once the D/A starts generating the waveform for transmission, the A/D begins to acquire the data from the IF bandpass filter. Figure 4.28 shows the number of sampling points in each section of the waveform. There are a total number of 243,840 points being sampled by the A/D, however only a fraction of this is actually being saved.
The LFM chirp return appears after a time delay equal to the two-way propagation of the transmit signal, and it is at this time when the A/D begins to save the data into storage. Referring to Figure 4.26, the length of time the A/D actually saves the sampled data is equal to the time pulse duration of the chirp wave minus the time delay of two-way propagation from the transmitted signal. In other words, 
\[ \Delta t = \frac{2R}{c} = \frac{(2 \times 20m)}{3 \times 10^8} = 133.3 \text{ ns} \] or equivalently, 
\[ 133.3 \text{ ns} \times 100 \text{ MHz} = 13 \text{ sample points}. \] Therefore, the total number of saved sample points per transmitted period is 
\[ 17,240 - 13 = 17,227 \text{ points}. \]

![Figure 4.28: Sampled Data Points from received waveform.](image)

In summary, the total number of sampled points from the A/D components is 243,840 points, but only 17,227 points of this total are actually being stored for data processing. It is during this period of saved data which the frequency difference between the transmitting and receive signals can range profile be extracted.

After the data are acquired and stored onto a storage disk, Matlab will process this data through a software program that will develop the various plots representing the data. The data processing involves averaging a certain number of return waveform periods to filter out some of the signal randomness (caused by random noise, changing environmental conditions, etc.) With the amount of over-sampling the A/D component samples, the number of times the return periods are averaged varies from data to data to develop the most representative example of the data. In addition
to the filtering out some of the random noise components, the processing software program also throws away a portion of the beginning acquired data. The reasoning behind a portion of the return signal being discarded is cut out that data that might be corrupted by the VCO not having enough time to settle down. Again, like the number of return period averaging, the amount of sample points discarded is selected on a case-by-case basis to best represent the target data.

4.3 Field Test Results

Field-testing of the HRR radar system involves taking the theoretical design out of the laboratory-controlled environment and subjecting the radar to practical environmental stimuli. It is during field-testing when the researcher learns of the shortfalls and inaccuracies of the radar design, which comes about from missed parameter assertions to over-optimistic assumptions. It is also at this stage when the limitations of the radar system are revealed and additional adjustments in hardware, or in expectations, are warranted. In the following subsections, the HRR radar will resolve a single trihedral, to serve as a reference target, and then include a secondary trihedral that will vary in range in relation to the reference target. Another test will replace the secondary trihedral with a different type of a target, namely a pickup truck. These testing configurations will provide information on range resolution that the HRR radar will be able to resolve.

4.3.1 HRR Radar Profiling Results

Figure 4.29 illustrates the target arrangement profile used to measure the range resolution of the HRR radar system.

The HRR radar hardware is mounted on a boom and is elevated 20 meters above the ground. In every field test configuration, there is a reference trihedral target
located 20 meters from the base of the boom carrying the radar. The combination of the elevation distance, along with the distance to the reference target, will provide a grazing angle of 45°, once the tilt of the radar is adjusted for maximum signal power return. Due to the height limitation of the HRR platform to such a shallow elevation, there is a corresponding limitation to the distance the targets can be set down range. If the targets are set too far down range, then the grazing angle will become too small and difficulty in resolving these targets becomes an issue. For example, the interference of two-way bounce from the terrain surface increases, increased clutter from the terrain becomes additive with the return signal or even the ability for the radar to illuminate targets behind each other becomes more difficult to resolve. Another issue concerning the height limitation of the radar platform and target range restriction is the decrease in the size of footprint illuminated on the ground. The decrease in swath size limits the measurement area where targets must reside in order to be within the 3 dB pencil beamwidth of the transmitting antenna.

The secondary trihedral target is offset to one side of the reference trihedral by a length of 3.5 meters. This offset will allow the radar to illuminate both targets
without the closer target blocking the secondary target located at some distance behind. The offset will add additional range to the secondary target during the two-way round trip propagation path than if it were located at the same range and directly behind the reference target. However, the additional range due to the target offset is negligible and adds only centimeters to the overall range-to-target.

![Secondary target offset diagram](image)

**Figure 4.30: Secondary target offset.**

In Figure 3.3 shows the schematic of the HRR radar system design. The functionality and necessity of each component within this schematic has been explained in Chapter 3. However, the HRR radar system is optimally designed for a radar platform operating at a height above terrain level of 750 - 1000 meters and a grazing angle of 45°. These parameters would give a range-to-target distance of 530 to 700 meters; a considerable increase from the 20 meters the profile configurations that the field-testing is setup for. In view of the minimum range-to-target of 20 meters, the low-noise amplifier and the IF amplifier will be excluded during field-testing measurements. Both of these amplifiers serve to boost the received signal at the different stages in the receiver section due to the low power returns from distant, or smaller
targets. By using 1-meter trihedral targets at a distance of 20 meters from the radar, the received signal returns possess power levels sufficiently high enough to eliminate the need for these amplifiers.

Included in the field measurement figures are a time domain plot of the acquired signal return, the second plot is the FFT (PSD) of the same acquired data as it is averaged over a number of return periods, and the third plot is a close-up view of the second PSD plot representing the frequency range of interest. Equations (4.1) and (4.2) describe the relationship of Range versus Frequency and are used to identify the target returns in the PSD plots of the figures that follow.

Figure 4.30 shows the profile configuration of a second trihedral target, identical to the reference target, that is positioned behind and to the side of the reference target. This is the setup profile configuration that is used in the remaining field test measurements where there are two targets being resolved.

The first radar measurement involves detecting a single trihedral target at a range of 20 meters. Once this measurement is obtained, the positioning of the radar and the trihedral target will be static, and this configuration is used as the referenced target to which the secondary target will refer to when considering range resolution.
4.3.1.1 Reference Target Range Results

Figure 4.31: A single Trihedral target at 20 meters down range (Reference Target.)

Figure 4.31 clearly shows a distinct target return at 20 meters (0.78 MHz), with a power level greater than 90 dB, which is 25 dB above the noise level. A 50 dB noise level shows up in the power spectrum plot with sporadic peaks and troughs, which can be attributed to the near distance of the target in relation to the grazing, also results from the rough surface of the terrain. The time domain plot in Figure 4.31 shows the superposition of sinusoidal waves (various frequencies) from the reference target along with the surface terrain and other noise contributors.
4.3.1.2 4 meter range results

Figure 4.32: Reference Target at 20 meters and Secondary Target at 24 meters.

Figure 4.32 shows the radar measurement from a setup profile of having the secondary target positioned at 24 meters down range, 4 meters behind the reference target. The PSD plots in Figure 4.32 illustrate two targets at points B and C; there is also a significant peak depicting a false target return at point A. The peak at point B has a power level of 76 dB showing the reference target at 20 meters range, while point C shows a peak of 69 dB power level from the secondary target located at 0.94 MHz, this frequency corresponding approximately to 24 meters range. The reference target at point B remains solid with a power level greater than 15 dB above the nearest peak. However at point C, the target return is within 5 dB above the
nearest sidelobe. There is the advantage of knowing in advance the distance the target is from the radar, in turn, the frequency at which the target should appear is known. Knowing that at the frequency 0.94 MHz a return peak should be present corresponding to a range of 24 meters, it is determined that this peak does actually represent a valid target return. Point A in the PSD plot shows the effects of the interaction between the multiple targets. These effects are possibly attributed to the bouncing of multiple reflections between the two targets, or perhaps emanating from the 2nd-time around echo effect.

4.3.1.3 3 meter range results

![Acquired Data](image)

![Power Spectrum Density](image)

Figure 4.33: Reference Target at 20 meters and Secondary Target at 23 meters.

Figure 4.33 shows the radar measurement from a setup profile of having the secondary target positioned at 23 meters down range, 3 meters behind the reference
target. The peak power level of the reference target is at 84 dB and still located at 20 meters down range. The secondary target shows a 69 dB power level located at 0.91 MHz, this frequency corresponding approximately to 23 meters range. The reference target remains solid with a power level greater than 20 dB above the noise level. However, the power level for the secondary target is within 5 dB above the nearest sidelobe. For the same reasoning provided in the 4 meter range results, by having advance knowledge of the target range (hence, frequency), it is determined that the return peak at 0.91 MHz does actually represent the valid target return.

4.3.1.4 2 meter range results

Figure 4.34: Reference Target at 20 meters and Secondary Target at 22 meters.

Figure 4.34 shows the radar measurement from a setup profile of having the secondary target positioned at 22 meters down range, 2 meters behind the reference
target. The PSD plots in Figure 4.34 illustrate two distinct targets. The peak power level of the reference target is at 79 dB and still located at 20 meters down range. The secondary target shows a 73 dB power level located at 0.86 MHz, this frequency corresponding approximately to 22 meters range. The power level for the secondary target is within 5 dB above the nearest sidelobe. For the same reasoning provided in the 5 meter range results, by having advance knowledge of the target range (hence, frequency), it is determined that the return peak at 0.86 MHz does actually represent the valid target return.

The next field test measurement involves replacing the secondary trihedral target with another specular target, namely the side view of a pickup truck.

4.3.1.5 Truck range results

Figure 4.35 shows the radar measurement from a setup profile of having the truck target positioned 25 meters down range, 5 meters behind the reference target. The PSD plots in Figure 4.35 illustrate two distinct targets. The power level of the reference target is at 73 dB located at 0.78 MHz (20 meters). The PSD plot does show a peak at 0.98 MHz, a frequency consistent with 25 meters, with a power level of 62 dB. The reflective qualities of the truck is not known, however, with most of the outward shell comprised of metal, it is assumed that the truck will have a strong specular component. At X-band frequency, the wavelength is approximately 3 cm. Therefore, with a 1-2 cm change in target aspect, major scintillations will occur in reflected returns. The increase in scintillations will have a greater effect contributing either positively or negatively to the radar signal return to the transmitter. Additionally, the 20-foot long truck positioned to the side of the reference target and possibly towards the boundary of the 3-dB beamwidth of the main lobe, it is thus possible that the entire side view of the truck (from bumper to bumper) is not being fully illuminated.
by the main lobe and only a fraction of the transmitted power is being reflected due to the limited exposure of the target. These factors considered by themselves, or in conjunction, could result in the power level of the truck target being relatively low.

4.4 Concluding Analysis

The HRR radar system hardware was rigorously tested in the laboratory to identify certain problematic areas of interest and then solutions were implemented to work around, or eliminate these issues. In laboratory testing, identifying the source of the low frequency interference is essential so that a suitable solution can be applied, while in the system design, a BPF with a center frequency relative to the range-to-target, is implemented in the IF section. The adjustment in the dynamic range of
gain (0 dB - 70 dB) in the IF amplifier, is shown to be a major factor in obtaining an acceptable power level of the downconverted signal from the mixer. The AGC is be adjusted according to the range the target is from the radar receiver and the RCS of the target being illuminated. Similarly, the input voltage level to the A/D component is adjusted according to the voltage level of the return signal from the BPF. Both the AGC and the input voltage level to the A/D are not to be considered a one-setting-fits-all adjustment for every measurement, rather samples of data from the given target/terrain should be taken prior to acquiring actual data and then adjusting these levels accordingly.

In operational radar systems, the need for calibration is to obtain accurate measurements of the radar cross section, \( \sigma^0 \). There are two types of calibration; internal and external. Internal calibration determines the relative scattering coefficient, whereas the external calibration determines the absolute scattering coefficient using a target of known RCS. It is determined that calibration on the HRR radar system would not be value-added at this early stage of evolution towards the airborne SAR. At this stage, the HRR system is primarily concerned with resolving two separated targets and identifying the best range resolution of the radar.

The van-mounted boom has a height limitation to 20 meters above ground level, restricting the average range-to-target distance to around 25 meters (± 10 meters) in order to maintain an adequate grazing angle of approximately 45°. Preliminary field tests provided feedback as to the power levels of the return signals being received at the input to the receiver. The height/range limitation, in addition to the sizeable 1.5-m sided trihedral targets, concluded that the need to boost the received signal is not required. Therefore, the LNA and the IF amplifier components were bypassed during the field testing measurements. Also, the BPF was replaced with a 10.5 MHz LPF adequate for the range of the target profile configurations.
During field testing, it was determined that the HRR radar was shown to be capable of accurately resolving a single trihedral target down range at various ranges between 20 meters to 30 meters. The power level of the return signal is a sharp peak, greater than 20 dB above the noise level, with a narrow bandwidth and low sidelobes as depicted in Figures 4.31 through 4.35. To determine the range resolution characteristics of this radar, a secondary target was placed along side the referenced target and return measurements were taken as the distance between the two targets decreased. With a priori knowledge of range in hand, the frequency is calculated where to locate the targets. Although the PSD plots in Figures 4.32 through 4.35 indicate peaks of the individual targets at the appropriate ranges, the power levels of the secondary targets are less than 5 dB above nearest sidelobes (or noise level). The noise level is persistent and hovers about 50 dB. A definite requirement to further implement software algorithms to suppress spurious peaks and for noise reduction is needed. Such data processing software additions are outside the scope of this thesis of hardware design and were not incorporated. However, implementing such algorithms would serve only to improve the SNR of the return signals and thus, adding to the validation of the existing data. Figure 4.34 shows two distinct targets resolved at a separated range distance of 2 meters. Measurements obtained with separated distances shorter than 2 meters in range, displays a severe degradation in range resolution due to the increase in noise from the nonlinearity of the VCO.

A 0.5% of nonlinearity in the VCO frequency versus voltage would increase the noise sidelobes from a value of -40 dB for 0% linearity to a level of -19 dB, as illustrated in the Figure 4.36.
Referring to Section 3.2.1 in Chapter 3, the limitations of the output voltage resolution of the D/A component being severely limited to 216 steps, the VCO will display nonlinearity greater than 0.5%. It is determined via theory and validated through testing measures, the most significant area of improvement should focus on the waveform generation of the LFM chirp. No other singular modification to the system will improve the noise suppression, and the range resolution, greater than improving the linearity of the VCO (or its equivalent).
Chapter 5

Conclusions

5.1 Summary of Work

Chapter 1 outlined the scope of this thesis as the design, construction and testing of a high resolution radar system with a range resolution greater than 4 meters. Chapters 2 through 4 systematically laid out the foundation to accomplish this goal. A review of the fundamental concepts towards understanding this radar system, along with fundamentals towards SAR systems, where explained in the early chapters. Then transitioning to explain the hardware design in detail down to component level and finally, giving validation to the working radar system.

Chapter 3 addressed the need to further improve on the linearity of the VCO through increasing the output voltage resolution of the D/A component. An analog extender circuit was design and tested with the radar system under laboratory conditions with inadequate results. The failing results are directly tied to the analog components, in the extender circuit, not being fast enough to keep up with the changing speed of the D/A component.

The HRR radar design uses the concept of designing the system with primarily off-the-shelf parts, but most notably is the use of an RF VCO to serve as the primary component to synthesis the transmitted waveform. Both these concepts add to keeping the cost of the design to a minimum.
The HRR hardware system in its current state is a functional radar system which can provide accurate target detection, and is a validated radar system to have a range resolution of approximately 2 meters. The HRR can not be considered an operational system with the given limitations to the VCO linearity, and the added need for noise reduction algorithms during data processing. The use of this HRR system to serve as the hardware platform building up to the airborne SAR is a distinct possibility, depending on the specifications required of the SAR system.

5.2 Recommendations for Future Work

The first item to address is improving the linearity of the VCO-D/A combination. If the design specifications remain "low-cost", then revisiting an improved version of the "extender" circuit is suggested. If lower noise and greater resolution is required, then the current design of the waveform generation can be simply replaced by a frequency synthesizer with a resolution of 1 Hz.

The implementation of improved algorithms to suppress spurious peaks and add noise reductions is additional area to consider. Although a frequency synthesizer will reduce the noise level considerably, there will still exist noise emanating from the system itself, changing atmospheric conditions, or clutter from the ground, all which require suppression or elimination.

If the HRR hardware system will graduate up to an airborne SAR system, then additional basic hardware/software are needed to assure valid data retrieval. This added hardware should address relative positioning (GPS), motion changes (motion compensation techniques) and digital matched filtering (software).
Bibliography


Appendix

A.1 VCO Data Sheets

Figure A.1: VCO tuning characteristics.
A.2 X-band Antenna Data Sheets

Figure A.2: Antenna: E-Plane radiation pattern.

Figure A.3: Antenna: H-Plane radiation pattern.
Figure A.4: Antenna: Return-Loss.
A.3 Analog Extender Circuit

Figure A.5: Analog EXTENDER circuit.