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Air-to-Air Radar Modes for the CP-140 Maritime Patrol Aircraft

by

Loraine L. Campbell

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

Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
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December 1996

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The undersigned hereby recommend to
the Faculty of Graduate Studies and Research
acceptance of the thesis

"AIR-TO-AIR MODES FOR THE CP-140 RADAR"

submitted by

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Master of Engineering

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December 1996

ABSTRACT

DND has a requirement to develop and demonstrate an air-to-air radar surveillance capability for the CP-140 maritime patrol aircraft. The current AN/APS-506 radar configuration is capable of pulsed operation, but can generate only low pulse repetition frequencies (PRFs). To detect and track an airborne target in a look-down scenario, the radar must make use of Doppler information to distinguish the target from ground clutter. DREO is investigating the feasibility of implementing medium PRF and high PRF pulse Doppler radar modes on the existing AN/APS-506 radar system.

It is feasible that if another transmitter can be integrated with the existing hardware in the AN/APS-506, the AN/APS-506 can be given the capability of air-to-air operation. This thesis proposes a set of air-to-air modes and signal processing techniques based on the CP-140 requirements. An analysis of the predicted performance of these modes is provided using theoretical calculations and simulated data.

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LIST OF ACRONYMS

ADC	Analog-to-Digital Converter
AEW	Airborne Early Warning
AGC	Automatic Gain Control
AI	Airborne Intercept
AWGN	Additive White Gaussian Noise
BLC	Backlobe Clutter
CA	Cell Averaging (CFAR)
CFAR	Constant False Alarm Rate
CMLD	Censored Mean-Level Detector
coho	Coherent Oscillator
CNR	Clutter-to-Noise Ratio
CPI	Coherent Processing Interval
DFT	Discrete Fourier Transform
DND	Department of National Defence
DREO	Defence Research Establishment Ottawa
FFT	Fast Fourier Transform
FM	Frequency Modulated
GMT	Ground Moving Target
GO	Greatest-Of (CFAR)
GTWT	Gridded Traveling Wave Tube
HPRF	High PRF
I and Q	In-phase and Quadrature
IF	Intermediate Frequency
IID	Independent and Identically Distributed
INS	Inertial Navigation System
JEM	Jet Engine Modulation
LEA	Long Ensemble Average
LPRF	Low PRF
MI	Mutual Interference
MLC	Mainlobe Clutter
MPRF	Medium PRF
OS	Ordered Statistic (CFAR)
pdf	Probability Density Function
PRF	Pulse Repetition Frequency
RCS	Radar Cross Section
RWS	Range While Search
SAFIRE	Simulator for Advanced Fighter Radar
SAR	Synthetic Aperture Radar
SCR	Signal-to-Clutter Ratio
SIR	Signal-to-Interference Ratio

SLC	Sidelobe Clutter
SNR	Signal-to-Noise Ratio
SO	Smallest-Of (CFAR)
STC	Sensitivity Time Control
STT	Single Target Track
TM	Trimmed Mean
TOT	Time on Target
TWS	Track While Scan
VFO	Variable Frequency Oscillator
VS	Velocity Search

LIST OF SYMBOLS

A	receiving antenna area
α	threshold
B_n	receiver noise bandwidth
BW_D	Doppler filter bandwidth
c	speed of light
d_t	transmitter duty cycle
f_E	difference in frequency between two PRFs required to achieve R_E
$\dot{f}_1, \dot{f}_2, \dot{f}_3$	frequency modulation rates for stages 1, 2, and 3 of HPRF RWS
Δf_{12}	difference in frequency measured between stages 1 and 2 (in HPRF RWS)
G	antenna directive gain
g_{sl}	sidelobe gain (relative to the mainlobe gain)
H_0	hypothesis 0 (no signal present)
H_1	hypothesis 1 (signal present)
k	Boltzmann's constant
λ	wavelength
L_a	loss due to atmospheric attenuation
L_{nr}	loss due to the non-Rayleigh characteristics of the sidelobe clutter
L_s	system losses
L_{sc}	loss due to the sidelobe clutter itself
μ	total noise power
N	number of pulses in a burst, number of cells used in the CFAR statistic
N_p	number of pause or fill pulses
Ψ	antenna grazing angle
P	cumulative (single-scan) probability of detection
P_{cd}	cumulative (multi-scan) probability of detection
P_d	probability of detection
P_{fa}	probability of false alarm

P_t	peak transmitter power
R	range to the target
R_c	range to the clutter patch
R_E	range at which a target will be eclipsed by both PRFs (in HPRF modes)
R_u	unambiguous range
ΔR	range resolution
R_0	range at which the SNR is unity
R_{50}	range at which the probability of detection is 50%
R_{85}	range at which the probability of detection is 85%
R_{90}	range at which the probability of detection is 90%
σ_t	target radar cross section
σ°	backscatter coefficient
$\left(\frac{S}{N}\right)$	signal-to-noise ratio (SNR)
$\left(\frac{S}{C}\right)$	signal-to-clutter ratio (SCR)
θ_a	azimuth look angle
θ_e	elevation look angle
T	CFAR scale factor (threshold multiplier)
T_1	lower trimming value in TM-CFAR
T_2	upper trimming value in TM-CFAR
T_c	amount of time for cumulative probability of detection
t_D	round-trip time delay
T_i	target illumination time
T_r	the scan-to-scan revisit time
T_s	system equivalent noise temperature
V_o	ownship velocity
Y	contents of the test cell
Y_0	fixed optimum threshold

Z noise power estimate statistic

CHAPTER 1: INTRODUCTION

The Department of National Defence has a requirement to develop and demonstrate an air-to-air radar surveillance capability for the CP-140 maritime patrol aircraft. The AN/APS-506 radar aboard the CP-140 was designed as a maritime surveillance radar with periscope detection capability for the anti-submarine warfare role. The current AN/APS-506 radar configuration is capable of pulsed operation, but can generate only low pulse repetition frequencies (LPRFs). LPRFs can be used to search for airborne targets in a look-up scenario where there is no mainbeam clutter to compete with the target returns. For an air-to-air look-down scenario, however, the mainlobe and sidelobe ground clutter can mask the target return.

To detect and track an airborne target in a look-down scenario, the radar must make use of Doppler (frequency) information to distinguish the target from ground clutter. The Defence Research Establishment Ottawa is investigating the feasibility of implementing pulse Doppler radar modes on the existing AN/APS-506 radar system in the CP-140. Pulse Doppler radars are employed for airborne target detection by virtue of their capability to detect small targets at long ranges over a wide field of view. In order to implement pulse Doppler operation, the current AN/APS-506 radar configuration must be altered to allow for higher PRFs.

One possible low-cost solution to this problem is to buy and integrate an off-the-shelf transmitter within the existing radar. There are several off-the-shelf transmitters that might be candidates for the CP-140 application. It is feasible that if another transmitter can be integrated with the existing hardware in the AN/APS-506, the AN/APS-506 can be given the capability of air-to-air operation.

1.1 THESIS OBJECTIVE

While there are many operational airborne early warning (AEW) and airborne intercept (AI) aircraft that use air-to-air radar modes for airborne target detection, information about these systems is limited in the open literature. More importantly, the CP-140 has a unique role and unique requirements that must be taken into consideration in the design of any air-to-air mode that will be implemented in the AN/APS-506 radar. The AN/APS-506 radar itself has unique characteristics that also affect the design of the modes. This thesis proposes a set of air-to-air modes and signal processing techniques that are specifically tailored to the CP-140 application. An analysis of the predicted performance of these air-to-air modes is provided using theoretical calculations and simulated data.

1.2 THESIS ORGANIZATION

The thesis is organized in the following manner. Chapter 2 provides the reader with a review of airborne pulse Doppler radar theory. Chapter 3 proposes a high-level design for the air-to-air modes to be used on the CP-140. Chapter 4 reviews radar signal processing techniques, and proposes techniques for use on the CP-140. Chapter 5 provides an analysis of the predicted performance of the air-to-air modes using theoretical calculations and simulated data. Probability of detection vs. range curves are given for different types of targets. Chapter 6 provides conclusions, and suggests areas for further work.

1.3 THESIS CONTRIBUTIONS

This thesis reviews state-of-the-art techniques in the design and implementation of air-to-air modes, and proposes air-to-air signal processing techniques that are tailored specifically for the CP-140. The contributions of this thesis can be summarized as follows:

- a. development of MPRF RWS, HPRF RWS, and VS mode parameters to be used in the CP-140 application,
- b. a comprehensive investigation of constant false alarm rate (CFAR) techniques, and development of a novel regional TM-CFAR technique to combat the problems in MPRF RWS mode due to the high sidelobes of the AN/APS-506 antenna,
- c. a theoretical performance analysis of the CP-140 air-to-air modes,
- d. alteration of the DREO SAFIRE simulator to allow simulation analysis of range-Doppler clutter maps based on the CP-140 antenna patterns and air-to-air modes, and
- e. a preliminary simulation analysis of detection performance of the CP-140 air-to-air modes, using the modified version of SAFIRE.

CHAPTER 2: A REVIEW OF AIRBORNE PULSE DOPPLER RADAR

In this chapter, the basic airborne pulse Doppler radar theory is reviewed. The subset of radar modes for air-to-air operation is explored, and relevant equations are given. The theory behind waveform design and detection algorithms is examined, and examples are given for air-to-air modes in use on modern aircraft.

2.1 AIRBORNE PULSE DOPPLER RADAR OVERVIEW

The term "Doppler radar" describes a radar that is capable of measuring the shift between the transmitted frequency and the frequency of backscattered echoes received from targets. Pulse Doppler radars are employed for airborne target detection by virtue of their capability to detect small targets at long ranges over a wide field of view. In order to implement pulse Doppler operation, the current AN/APS-506 radar configuration must be altered to allow for higher PRFs. Unfortunately, the transmitter currently being used is not capable of producing PRFs above 2 kHz. In order to understand the extent to which the PRFs must be increased, the types of pulsed radar modes must be briefly reviewed. A more detailed description of pulsed Doppler radar modes can be found in [1, 2, 3, 4, 5].

The three types of pulsed radar modes include low, medium, and high PRF (LPRF, MPRF, HPRF). Figure 2-1 shows an example of time-domain plots of pulse envelopes for these modes. In an airborne pulse Doppler radar system, the receiver is gated off during the transmitted pulse and vice versa, permitting the radar to operate from a single antenna.

Figure 2-1 . Low, Medium, and High PRFs

The definition of each mode is normally based on ambiguities (or lack thereof) in range and Doppler. Consider a radar transmitting a pulse train at a particular PRF. Since electromagnetic energy travels at the speed of light, c , the range to a target is calculated as:

$$R = \frac{c \Delta t}{2} \quad , \quad (2-1)$$

where Δt is the round-trip delay time. The unambiguous range, R_u , is the maximum target range that can be measured by the radar before the next pulse is transmitted. R_u is, therefore, a function of the pulse repetition interval (PRI):

$$R_u = \frac{c(PRI)}{2} \quad . \quad (2-2)$$

Beyond this range, the elapsed time between transmit and receive pulses will include multiples of the PRI, making the range ambiguous. LPRF modes employ a PRF

sufficiently low that range is unambiguously measured up to the maximum expected detection range. The transmitted pulse travels to and from the targets of interest before the next pulse is transmitted.

To understand how the choice of PRF affects detection in the Doppler spectrum from an airborne platform, first consider how target and platform motion will produce a Doppler shift. The Doppler shift of a backscattered signal is a function of the relative radial velocity of a target, $\dot{R}$, and the wavelength of the transmitter carrier frequency, λ . At aircraft velocities (i.e., neglecting relativistic effects), the Doppler shift, f_d , is defined as:

$$f_d = \frac{-2\dot{R}}{\lambda} \quad (2-3)$$

The negative sign is used to indicate that approaching targets (negative range rate $\dot{R}$) produces a positive Doppler shift.

From an airborne platform, the sidelobes and backlobes of a radar antenna will illuminate the ground below the ownship (the aircraft on which the radar is mounted). Backscatter from the ground (clutter) will be received at the antenna, and will occupy the portion of the Doppler spectrum equal to plus or minus the Doppler shift of the ownship velocity, V_o . Figure 2-2 shows a simplified example of the central band of a Doppler spectrum at baseband for a rearward-looking antenna. The mainlobe clutter (MLC) illuminates the ground behind the aircraft, and thus appears with a negative Doppler shift. The altitude line is a result of transmitter leakage and backscatter off the ground directly below the aircraft.

In the pulsed operation of a coherent radar, this central band will be repeated on either side of the central band at multiples of the PRF [4]. These bands are further modulated by an envelope of the $(\sin x)/x$ shape of a single pulse. The choice of PRF dictates how much overlap there will be between the central and sidebands, if any.

Figure 2-2 . Central Band of Doppler Profile

At LPRFs, there is extreme overlap of the central and sidebands of the Doppler profile, and no attempt is normally made to determine the Doppler shift of a target. LPRF modes are typically used for air-to-surface applications, but may perform adequately in air-to-air look-up situations. They provide accurate range, but do not provide Doppler information.

HPRF modes employ a PRF sufficiently high that all target velocities of interest are unambiguously measured. At high PRFs, the clutter bands are spread far enough

apart to open up a clutter-free region between them. The width of this clutter-free region is normally chosen to encompass all targets of interest.

Figure 2-3 . High PRF Doppler Spectrum

HPRF modes are used to detect high closing or receding velocity targets with good Doppler resolution. Since these targets normally appear in the clutter-free part of the spectrum, target returns compete only with receiver noise for detection. HPRF modes normally employ a high duty cycle, allowing maximum mean power on the target for a given peak power, and therefore maximum detection range for these types of targets. There are multiple range ambiguities, however, and HPRF modes inherently provide little range information. Coarse range resolution is possible with HPRF modes if frequency modulation (FM) techniques are applied. FM techniques will be described in Subsection 3.6.4.

MPRF modes lie between the two, and provide the best all-aspect capability by combining desirable features of the LPRF and HPRF modes. At medium PRFs, the Doppler sidebands overlap, but to a lesser extent than during LPRF operation. There is normally no region free of sidelobe clutter (SLC) in the frequency domain. The overlap of the central band and the immediate sidebands before sampling is shown in Figure 2-4. Note that in this case, the mainlobe clutter is shown with a positive

Doppler shift, reflecting an antenna pointing angle that is forward of the beam. Moving target detection is limited to detection in sidelobe clutter because the mainlobe clutter is simply too strong. Thus, for MPRF modes, the optimum antenna has very low sidelobes and a "clean" radome. MPRF modes produce ambiguities in both range and Doppler. These ambiguities are resolvable, however, when multiple PRFs are used. Good range and Doppler resolution are possible with MPRF modes, but MPRF modes normally require a higher signal processor load than for HPRF modes.

Figure 2-4 . Medium PRF Doppler Spectrum

There are no clear dividing lines between the PRFs chosen for each type of radar mode, as they depend on each application. A typical X-band (9 GHz) fire control radar on a fighter aircraft might have a high PRF of 200 kHz to ensure unambiguous velocity measurement for similar aircraft. The same radar might employ medium PRF modes between 8 kHz and 20 kHz, and low PRF modes below 2 kHz.

Additional background information about air-to-air modes and airborne radar signal processing techniques is presented in Chapters 3 and 4 in conjunction with the specific design for the CP-140.

2.2 FIELDDED SYSTEMS

The following tables outline radar characteristics for some surveillance aircraft. The values in these tables are available in the open literature, and are provided to give

a general overview of surveillance radars. Just as the role of the CP-140 is changing to include more than maritime surveillance, the role of the American P-3 aircraft has changed as well. An air-to-air mode is planned as an upgrade to the AN/APS-137(V) that is normally fitted to the P-3C aircraft [6]. In the meantime, two different alterations to the radar configuration on the P-3 have been made to allow for air-to-air target detection. These are presented in Tables 2-2 and 2-3. It should be noted that the operating conditions are not normally quoted along with the range performance in the open literature. As will be seen in later chapters, the conditions under which the range was estimated or measured greatly affect the interpretation of radar performance. The ranges quoted here may likely be the most optimistic performance for that radar.

Table 2-1 . AN/APS-138 Characteristics

E-2C Hawkeye AN/APS-138	
Frequency	UHF (430 MHz)
Power	1 MW peak, 3.8 kW average
Antenna beamwidth	Azimuth: 6.6°, Elevation: 20°
Range	Up to 300 miles (480 km)

Table 2-2 . AN/APS-125/138 Characteristics

<p>P-3 (AEW&C) Using an E-2C Rotodome and an AN/APS-125/138</p> <p>Two of these have been used by the US Customs Service (one with a '125, and one with a '138) since 1989 for drug interdiction. The '138 is an upgraded version of the '125.</p>	
Frequency	UHF (430 MHz)
Power	1 MW peak, 3.8 kW average

Antenna beamwidth	Azimuth: 6.6°, Elevation: 20°
Range	As good or better than that achieved by the E2-C

Table 2-3 . AN/APG-63 Characteristics

<p>P-3 Orion Using an AN/APG-63 (same radar as the F-15)</p> <p>Four of these have been used by the US Customs Service for drug interdiction since 1986. These scan a 120° arc in front of the aircraft.</p>	
Frequency	X, Ka band (8-20 GHz)
Power	12.975 kW peak
Antenna beamwidth	3°
Range	161 km. Typical 90 km for an F-4 sized target

2.3 PROBLEM STATEMENT AND DEFINITION

DND has a requirement to develop and demonstrate an air-to-air surveillance capability for the CP-140 aircraft. Air-to-air target detection is not currently possible in most operational scenarios with the AN/APS-506 radar due to the transmitter's maximum PRF of 2 kHz. The AN/APS-506 is an X-band radar. It is feasible that if another off-the-shelf X-band transmitter can be integrated with the existing hardware in the AN/APS-506, the AN/APS-506 can be given the capability of air-to-air operation.

While there are many operational airborne early warning (AEW) and airborne intercept (AI) aircraft that use air-to-air radar modes for airborne target detection, information about these systems is limited in the open literature. More importantly, the CP-140 has a unique role and unique requirements that must be taken into

consideration in the design of any air-to-air mode that will be implemented in the AN/APS-506 radar. The AN/APS-506 radar itself has unique characteristics that also affect the design of the modes. This thesis proposes a set of air-to-air modes and signal processing techniques that are specifically tailored to the requirements of the CP-140 application. An analysis of the predicted performance of these air-to-air modes is provided using theoretical calculations and simulated data.

CHAPTER 3: HIGH-LEVEL DESIGN

3.1 SYSTEM CONSTRAINTS

The following table outlines parameters for the potential implementation on the CP-140. Several of these parameters are imposed by the existing hardware, some are proposed modifications, and some are current operational parameters for the CP-140. The parameters that are being proposed as modifications for the CP-140 air-to-air modes are marked with an asterisk.

Table 3-1. CP-140 Implementation Parameters

<i>Parameter</i>	<i>Implementation on CP-140</i>
Antenna gain	33 dB
Antenna scan rate	24 °/s (4 rpm)*
Antenna beamwidth	2.5 ° in azimuth, 4.5 ° in elevation
Antenna sidelobes	18 dB below mainlobe, 23 dB below at $\pm 10^\circ$ (az) -20 dB spur at -18° (el)
Antenna scan limits	360 ° in azimuth, +10/-15 ° in elevation
Number of antennas	Single antenna
Air-to-air modes, L, M, or H PRF	Range while search (RWS) -M, H* Track while scan (TWS) -M, H* Velocity search (VS) -H*

Maximum aircraft speed	Mach 0.53 (350 kts) Mach 0.75 (500 kts) with tailwind
Normal cruise speed	Mach 0.3 (200 kts)
Max / Min expected closing speeds (Mach)	3 / -3
Platform altitude	1000-24,000 ft, normal cruise at 20,000 ft

3.2 PROPOSED MODES

A complete description of the radar modes proposed in Table 3-1 can be found in [4]. A brief description is provided here for each of the modes.

Range while search (RWS) is an all aspect search mode. It can provide look-up and look-down target detection capability for high closure rate head-on attacks, and low closure rate tail attacks. It alternates between both medium and high PRFs to provide coverage for all types of targets. As mentioned in Chapter 2, a medium PRF produces ambiguities in both range and Doppler. These ambiguities can be resolved when the target is detected on multiple PRFs. Because the receiver is gated off during the time the transmitter is on, part or all of the target return may arrive when the receiver is turned off, resulting in partial or total eclipsing. Target detection in MPRF modes normally follows the relatively common $M:N$ PRF selection method to allow for ambiguity resolving and to reduce eclipsing [4]. During MPRF RWS, the radar typically cycles through eight different PRFs, and detection is required on at least three of eight PRFs in order to declare a target. During HPRF RWS, high duty cycles are normally used to maximize the detection performance of the radar. The radar typically alternates between two different HPRFs to reduce eclipsing. Target ranging is

performed during the HPRF mode by modulating the pulse with a linear frequency modulation ramp.

Track while scan (TWS) provides an all-aspect look-up, look-down multi-target detection and track capability. TWS is similar to RWS, except that more processing (tracking) is performed. Within TWS, two submodes are envisioned: in the first submode, the radar would continue all-aspect coverage while maintaining tracks on selected targets. In the second submode, the radar would perform TWS only over a small volume, and for a single target. The radar would continually monitor the range, angle, and/or velocity of the one target of interest. HPRF would be used for angle and velocity tracking, MPRF would be used for angle, range, and velocity tracking, and LPRF could be used in look-up situations for angle and range tracking.

Velocity search (VS) is a long range target search mode that detects high velocity nose and tail-aspect targets (aircraft flying toward and away from the aircraft). It provides a longer detection range for these targets than RWS or TWS, but does not detect beam targets (targets flying perpendicular to the aircraft's motion), or slow moving nose/tail-aspect targets. As with other HPRF modes, target velocity is measured unambiguously, but no attempt is made to resolve range ambiguities. Several PRFs are used to reduce eclipsing.

3.3 TRANSMITTER CHARACTERISTICS

The AN/APS-137(V) on the P-3C has a peak power of 50 kW and an average power of 500 W. The 500 W average power of the AN/APS-137 and the Searchwater 2 radars and the 550 W average power of the Skymaster radar are typical values for radars of this size [6]. Typical transmitter duty cycles for airborne radars vary from

below 1% for the high peak power radars to as high as 40% [8]. A typical average power of 500 W will be used in this analysis to represent a generic transmitter for the air-to-air modes for the CP-140. Because a higher duty cycle represents a more challenging design due to greater eclipsing of the received pulse, a duty cycle of 30% will be considered in this analysis. When a transmitter is finally selected for the CP-140 application, its characteristics will have to be incorporated in the design of the air-to-air modes.

Because there is some leakage through the circulator into the receive chain, the receiver must be switched off while the transmitter is on to protect the receiver components and to avoid saturating the analog-to-digital converter (ADC). In real systems there is always a finite transmitter switching time. The transmitter switching time affects the placement and perhaps the number and width of the range gates in each of the modes. A rise and fall time of 200 ns is a reasonable estimate for current off-the-shelf X-band transmitters [38]. When more extensive measurements can be made of the transmitter that is selected for use on the CP-140, this switching time should be updated accordingly in the different modes.

3.4 SIDELOBE CONSIDERATIONS

Because the MPRF RWS mode attempts to detect targets in the Doppler region occupied by sidelobe clutter, the antenna sidelobe levels play a major role in the performance of this mode. If the sidelobe clutter level is low enough, thermal noise dominates the spectrum, and the radar can approach the detection performance that is possible in noise-only conditions. If the clutter level rises, the detection performance degrades, and prediction of the detection performance is based on clutter statistics.

Obviously, the lower the antenna sidelobes, the lower the sidelobe clutter for any of the possible scenarios. Thus, it is desirable to have sidelobe levels as low as possible for two main reasons. The first is that high sidelobe levels may cause the return signal from large radar cross section (RCS) scatterers or clutter "discretes" to be high enough to cause false alarms. The second reason is that high sidelobe levels may bring down the overall signal-to-clutter ratio (SCR) so that a true target signal may not be detected above the sidelobe clutter.

Given an existing antenna, airborne pulse Doppler radars must minimize the probability of false alarm resulting from backscatter from clutter discretes that are received through the antenna sidelobes. There are three common methods used to combat these sidelobe discretes. *Sidelobe blanking* attempts to locate target detections that may have occurred due to the sidelobes, and simply discards those detections. *Sidelobe cancellation* attempts to cancel energy entering the main antenna's radar via the use of auxiliary elements and advanced processing [4]. A third method is the use of a *guard channel* in conjunction with the main channel, such as that used on the F-15's APG-63. The guard channel uses a widebeam, low-gain antenna. The guard antenna has more gain than the main antenna sidelobes, but less than the main antenna mainlobe. The main and guard channels are processed independently. If both the main and guard channel detect potential targets (hits) in common range bins, the amplitudes of the targets are compared. If the guard channel magnitude is greater, the hit is discarded.

As noted in Table 3-1, there is no guard antenna on the CP-140. Thus, the sidelobe discretes must be dealt with in another manner. Sidelobe cancellation is not possible without modification to the antenna hardware. Sidelobe blanking can be done,

however, using postdetection sensitivity time control (STC). This approach compares received amplitudes with expected values on an unambiguous range basis, and determines whether the return is from the mainbeam or from a sidelobe of the antenna. In most cases, all but the strongest RCS sidelobe discretely are weaker at their true range than the expected main beam targets at that range. Thus, by combining postdetection STC and range correlation, sidelobe discretely are eliminated. The radar displays only those targets that have an amplitude (at the true range) that is large enough to be a main beam target of interest. This is the method used in the F-16's AN/APG-66 and AN/APG-68 radars [4]. Although this technique might not be as reliable as the guard channel processing, it avoids the cost of the duplicate channel and the additional processing.

To determine whether this method is feasible with the antenna on the CP-140, the antenna pattern must be investigated more closely. As noted in Table 3-1, the AN/APS-506 must deal with antenna sidelobes that are much higher than those enjoyed by typical AI radars. Figures 3-1 and 3-2 demonstrate typical azimuth and elevation antenna patterns for the AN/APS-506 antenna [10].

Figure 3-1. AN/APS-506 Azimuth Pattern

Figure 3-2. AN/APS-506 Elevation Pattern

The first azimuth sidelobes are 18 dB below mainlobe, dropping to 23 dB below the mainlobe at $\pm 10^\circ$ in azimuth. In addition, there is a large sidelobe 20 dB down from the mainlobe at 18° in elevation, and another at 23 dB down at minus 22° . In contrast, typical AI radars typically use antennas that have sidelobes at least 28 dB down outside the region $\pm 5^\circ$ from the mainlobe.

All other factors being equal, a target illuminated at the same range by the radar mainlobe and sidelobe have a difference in SCR equal to the two-way difference between the mainlobe and sidelobe gains. Consequently, a typical AI radar, the SCR of this target return differs by at least 56 dB. The SCR of a target varies linearly with the target RCS [11]. Suppose the targets of interest for the radar were determined to have radar cross sections between 1 m^2 and $1,000 \text{ m}^2$. The SCR of the largest target illuminated by the sidelobe would be 26 dB lower than that of the smallest target illuminated by the mainlobe. The radar could set a threshold somewhere between the two values for targets at that particular range, and eliminate detections that fell below that value as sidelobe discretes. Note that corner reflectors in urban environments may have radar cross sections on the order of $30,000 \text{ m}^2$ at X-band [2]. Virtually no radar is immune to sidelobe discretes of this magnitude if sidelobe blanking is the only rejection method used.

Now take the two-way difference of the AN/APS-506 antenna sidelobes. The elevation sidelobe that is 20 dB down results in a two-way difference of only 40 dB. If the same targets of interest were considered, the difference in SNR between the largest target illuminated by the sidelobe and the smallest target illuminated by the mainlobe would be only 10 dB. Thus, the AN/APS-506 would have a much higher potential for error for a given threshold than a typical AI radar. In practice, only partial returns

from a target may be processed due to eclipsing, thereby reducing the SNR of a mainlobe target even further. The threshold must be set low enough so that partial returns from small targets in the mainlobe are not eliminated.

Sidelobe blanking using postdetection STC is a reasonable method to use on the CP-140, as it provides at least some capability to reject sidelobe discretes. Since postdetection STC relies on the difference between the mainlobe and sidelobe gains, however, sidelobe blanking cannot provide as low a probability of false alarm for the AN/APS-506 for a given probability of detection. The probability of false alarm due to sidelobe discretes depends on the targets of interest for a particular scenario and the environment in which the radar is operating.

3.5 MEDIUM PRF RANGE-WHILE-SEARCH

3.5.1 PRF Selection

As mentioned in the previous section, the most common form of ambiguity resolution in MPRF modes in modern pulsed-Doppler radars is to use a "three of eight" scheme [4]. Eight bursts with different PRFs are transmitted, and detection is required on at least three of the eight in order to resolve ambiguities. In general, detection is only required on two different PRFs in order to resolve ambiguities for a single target. If two targets are present, however, the ambiguous detections in range and Doppler might cause a false target (or "ghost" target) to appear after resolving. Use of a third PRF resolves this problem for the two-target case [4]. The combination of PRFs and pulse widths must be chosen to optimize the performance of the proposed mode. There are several factors influencing the upper and lower limits of the choices of PRF and pulse width.

The underlying objective of any search radar is to maximize its detection range for a given mode. If a matched transmit/receive system is used, the noise bandwidth B can be set equal to $1/\tau$ where τ is the pulse width. If the receiver is range gated to match the transmitted pulse duration, the duty cycle of the transmitter, d_t , is τf_p , where f_p is the PRF. The general form of the radar range equation for pulsed Doppler radar developed in [11] becomes:

$$\left(\frac{S}{N}\right) = \frac{P_t G^2 \lambda^2 \sigma_t T_i d_t}{(4\pi)^3 R^4 (k T_s) L_s} \quad , \quad (3-1)$$

where

$$\left(\frac{S}{N}\right) = \text{signal-to-noise ratio,}$$

$$P_t = \text{peak transmitter power,}$$

$$G = \text{antenna directive gain,}$$

$$\lambda = \text{wavelength,}$$

$$\sigma_t = \text{target radar cross section,}$$

$$T_i = \text{target illumination time (coherent integration),}$$

$$R = \text{range to the target,}$$

$$k = \text{Boltzmann's constant}$$

$$T_s = \text{system equivalent noise temperature, and}$$

$$L_s = \text{system losses.}$$

In order to maximize the detection range, the radar must radiate the maximum average power, given by $P_t d_t$, toward the desired target. A pulsed Doppler radar with a fixed peak power can do so by maximizing d_t . The selection of PRFs and pulse widths are

closely linked then, and effort must be made to achieve as high a duty cycle as possible.

When using a multiple PRF scheme, there is a tradeoff between the desire to maximize the duty cycle and the system complexity. If the duty cycle were held constant as the radar cycled through PRFs in an MPRF mode, each pulse width would have to change accordingly. This would result in having to alter the bandwidth of the receive chain (to match the bandwidth of the pulse) for each PRF, and would result in different range gate widths for each PRF. The system would have to be more complex, and the task of range resolving would become more difficult. Because of this, operational radars normally use fixed pulse and range gate widths for each PRF in MPRF mode. An attempt is made to maximize the duty cycle of the highest PRF. The other PRFs then use the same pulse width. When the set of medium PRFs is chosen to ensure that the target is detectable in Doppler frequency, the ratio between the highest and the lowest PRF may be as much as 2:1. When a constant transmitter pulsewidth is used, then the average power and signal-to-noise ratio also vary by 2:1.

The eight-PRF set in MPRF RWS mode must be chosen to cover the range-Doppler regions of interest as completely as possible. All factors affecting the performance of the mode must be taken into consideration before choosing the final set. These include the expected ownship velocities, the expected target velocities, and the antenna pattern with which the radar illuminates the targets.

During MPRF operation, the mainlobe clutter is normally either removed by filtering (often referred to as the mainlobe clutter (MLC) notch) or ignored, and detections are made in the sidelobe clutter region. In addition, air-to-air MPRF modes often ignore slow-moving targets, as they are most likely ground-moving targets

(GMTs). The APG-66, for instance, blocks out targets that are moving slower than 55 knots. This corresponds to a notch of approximately ± 2 kHz at X-band [12].

It is desirable to keep the MPRFs low to minimize the foldover in range, offering a potential decrease in sidelobe clutter (and thereby improved target detection). The MPRFs must be high enough, however, so that targets can be detected in Doppler as well as in range. Recall from Figure 2-4 that the central and sidebands normally overlap. Target detection is attempted in the Doppler spectrum that remains after the MLC notch. A common design criterion for the lowest PRF in the set is to choose the PRF such that there is still 50% of the Doppler spectrum available for detection processing [2]. Reducing the PRF further brings the unusable regions of the spectrum closer and closer together, resulting in diminishing returns. On the basis of a 50% design criterion and a ± 2 kHz notch for slow-moving targets, the radar would require a minimum MPRF of 8 kHz. Indeed, most airborne intercept (AI) radars have medium PRFs in the range 8-20 kHz [4].

The question arises as to whether a full ± 2 kHz notch required for the CP-140. The aircraft maximum speed is slower than that for an AI radar, so the mainbeam clutter spread is smaller. Also, slow-moving targets are perhaps of interest during maritime surveillance. These factors point toward the possibility of lowering the lowest PRF below 8 kHz.

Defeating the potential for a lower MPRF, however, are the sidelobe levels of the AN/APS-506 antenna. As noted in Table 3-1, the AI radar antenna sidelobes are typically at least 28 dB below the mainlobe. Low sidelobe levels are absolutely critical to achieving a signal-to-clutter ratio (SCR) that allows for target detection. The immediate AN/APS-506 azimuth sidelobes, however, are only 18 dB below the

mainlobe. Thus, after the MLC notch, there is a portion of the range-Doppler map for which the (two-way) sidelobe clutter return is 20 dB higher. If a target lies in this region, there will be a 20 dB difference in the SCR. Consider, for example, a target that is just detectable to the AI radar in this region. All other factors remaining equal, a 20 dB difference in the SCR means that the target RCS would have to be 100 times as large for the AN/APS-506 air-to-air mode to detect it. At $\pm 10^\circ$ in azimuth, however, the sidelobe levels in the AN/APS-506 azimuth pattern drop to 23 dB below the mainlobe. This corresponds to a two-way difference of only 10 dB in the SCR compared to a typical AI radar. Thus, it is more reasonable to detect targets outside the clutter illuminated by the $\pm 10^\circ$ pattern.

In order to determine the minimum MPRF that accommodates these sidelobe levels, the effect of notching out the $\pm 10^\circ$ beamwidth must be investigated. The widest mainlobe clutter (MLC) spread in azimuth occurs when the antenna is pointed perpendicular to the direction of motion with an elevation angle of 0° . The frequency of the mainlobe clutter for a given azimuth and elevation is equal to:

$$f_{mlc} = \frac{2V_o \cos\theta_a \cos\theta_e}{\lambda} \quad , \quad (3-2)$$

where V_o is the ownship velocity, θ_a is the azimuth angle from the direction of motion, θ_e is the elevation angle, and λ is the wavelength of the radar. With the antenna pointed perpendicular to the direction of motion, the $\pm 10^\circ$ beamwidth extends from 80° to 100° in azimuth. At an ownship velocity of 257 m/s (350 kts), the clutter from this beamwidth extends from minus 2 kHz to +2 kHz in the Doppler spectrum. Thus, if 50% of the spectrum is to be usable, the minimum PRF must be 8 kHz. This is demonstrated in Figure 3-3. The region marked "50% usable" will be referred to as

the weak sidelobe region, and the area between the MLC and this region will be referred to as the strong sidelobe region.

Figure 3-3. Minimum MPRF

As will be seen in Chapter 4, it actually appears worthwhile to provide some detection sensitivity in the strong sidelobe clutter region. To avoid a sharp increase in the false alarm rate, however, the same probability of detection cannot be achieved here. Thus, the width of only the weak sidelobe region is considered in choosing the minimum PRF. Consequently, the minimum MPRF for the CP-140 implementation is chosen to remain at the 8 kHz value typically used by AI radars, despite the slower ownship velocity of the CP-140.

As mentioned earlier, MPRFs typically lie between 8 and 20 kHz. Choosing eight PRFs to best cover the range and Doppler spectrum is a complex task, and is outside the scope of this thesis. An example is presented here to demonstrate the coverage of a set of eight PRFs. Take, for example, a minimum PRF of 9,305 Hz and a maximum of 18,695 Hz. Table 3-2 displays the resulting PRFs if eight PRFs are evenly spaced between these two frequencies. Both the APG-63 and APG-66 use a 13 μ s pulse with a 13-bit Barker code in their MPRF modes [4, 12]. This is the configuration that will be considered here. An explanation of pulse compression using

Barker phase coding can be found in [1, 2, 4]. This choice of pulse width and Barker code results in a compressed pulse width of 1 μs or a range gate width of 150 m. For the maximum PRF, this pulse width results in a duty cycle of 24.3%, which is within the 30% duty cycle of the "typical" transmitter used in this investigation. Also listed in Table 3-2 are the number of range gates, N_{rg} , available from each interpulse period based on the compressed pulse width. These assume a shortening of the interpulse period of 400 ns to allow for transmitter settling, as discussed in Section 3.4.

Table 3-2 Evenly-Spaced MPRFs

	PRF (kHz)	PRI (μs)	N_{rg}
1	9.305	107.47	94
2	10.647	93.92	80
3	11.988	83.42	70
4	13.329	75.02	61
5	14.671	68.16	54
6	16.012	62.45	49
7	17.354	57.62	44
8	18.695	53.49	40

3.5.2 Coherent Dwell

Coherent dwell or coherent processing interval (CPI) refers to the amount of time during which a pulse Doppler radar carries out coherent integration of pulse returns. Recall from Equation 3-1 that the longer the coherent integration, the higher the target SNR. Because the platform and expected targets move much slower for the CP-140 than for typical AI radars, the longer revisit time that must accompany a longer

dwell can be tolerated. Computational efficiency must be taken into account when setting the number of integrated pulses per burst. Many digital signal processors are optimized to perform the fast Fourier transform (FFT), so it is common to set the number of pulses to a power of two. Based on the antenna scan rate that is investigated in Section 3.8, it is reasonable to set the length of the coherent integration to 128 pulses for MPRF RWS on the CP-140.

The length of time that must be spent at a particular PRF must be longer than simply the number of pulses multiplied by the PRI. The radar must use "fill" or "pause" pulses at the beginning of a particular PRF so that each processed PRI receives radar returns from the maximum possible detection range. Consider, for instance, the radar returns seen after the very first transmitted pulse. By definition, the PRI is short enough in MPRF that the radar returns received during the first interpulse period do not encompass the entire range of interest. There will be no information about the more distant targets in the first PRI. Indeed, many PRIs may occur before the backscatter from the first pulse off a distant target of interest is received.

In order to provide the maximum processing gain for more distant targets, the radar must add enough fill pulses at the beginning of a burst of pulses so that the first PRI that is actually processed could contain radar returns from a target at the maximum detection range. As will be seen in Chapter 4, the required maximum unambiguous detection range for the MPRF RWS mode on the CP-140 should be approximately 350 km.

Using Equation 3-1, there must be a fill time of 2.347 ms in order to account for this maximum unambiguous range. The number of fill pulses required at each PRF is different, since each PRI is different. Table 3-3 outlines the number of fill pulses,

N_p , required to account for this maximum unambiguous detection range (rounded to the nearest integer). Two additional PRIs are added to the 128 plus N_p to allow for settling time for automatic gain control (AGC). If AGC is not used, this will simply allow a slightly longer maximum unambiguous detection range.

Table 3-3. PRIs for Evenly-Spaced MPRF Set

<i>Sequence Number</i>	<i>PRI length (μs)</i>	N_p	<i>Total dwell (μs)</i>
1	107.47	22	16.12
2	93.92	25	14.37
3	83.42	28	13.01
4	75.02	32	12.00
5	68.16	35	11.11
6	62.45	38	10.37
7	57.62	41	9.74
8	53.49	44	9.20

Adding up the dwell at each of the actual PRFs to be used on the CP-140, the total dwell is 104.8 ms. Thus, the antenna beamwidth must encompass a target of interest for at least 104.8 ms. This corresponds to a required scan rate of approximately 4 rpm for the 3-dB antenna pattern of 2.5°. Antenna scan patterns to accomplish this dwell are discussed in Section 3.8.

3.5.3 Range and Doppler Coverage

As mentioned earlier, the target must be detected on at least three of the eight PRFs in order to allow range and Doppler ambiguity resolving. The question arises as

to whether to attempt range or Doppler resolving first. Both approaches should produce the same result. Computational efficiency, however, favours range resolving prior to velocity resolving, as explained below.

The same pulse width (13 μs) and range gate width (1 μs) were chosen for all eight PRFs. Therefore, it is not necessary to calculate the actual range that corresponds to a range gate for purposes of comparison. While the total number of unambiguous range gates for each PRF is different, they all begin approximately 13 μs after the start of a PRI (immediately after the pulse transmission plus 200 ns for settling time). Each 1 μs range gate then follows sequentially. This situation is different for the Doppler filters. The width of the Doppler filters for the PRFs shown in Table 3-2 varies from 73 Hz to 146 Hz, depending on the PRF. Based on a transmit frequency of 9.6 GHz, this represents a radial velocity width from 1.1 m/s to 2.3 m/s. In order to compare hits on different PRFs, the radar must compute the velocity that corresponds to each hit rather than just comparing the target Doppler bin numbers. This necessitates a look-up table of the filter bandwidths for each PRF, and multiplication of the target Doppler bin number by the filter bandwidth. Thus, the velocity resolving routine is more computationally intensive. Because hits are required to correlate on at least three of the eight PRFs, the range resolving is likely to eliminate a large percentage of any false alarms that may occur, reducing the computations required for velocity resolving. Consequently, it is advantageous to carry out the range resolving before the velocity resolving. Once targets have been resolved in range, the resolved hits are passed to the velocity resolving routine.

The use of receiver blanking and MLC rejection causes blind zones in the range and Doppler coverage provided by each PRF. The overall range and Doppler coverage

achieved with the eight PRFs must, therefore, be considered in order to predict the performance of the radar. It is desirable that the range and Doppler regions of interest are covered by as many of the PRFs as possible. Recall that the target must be seen on at least three of the eight PRFs in order to be declared. Coverage by more than three of the PRFs guards against possible signal dropout due to target radar RCS scintillation or jamming.

While a pulse is being transmitted, the receiver is blanked and no data is processed. This results in a blind zone that eclipses any target returns during that period of time. During the interpulse period, the receiver is turned back "on" to allow data collection. Based on increments of 1.5 m, Figure 3-4 shows the effect of receiver blanking for the eight selected PRFs on the range coverage out to 350 km. Notice the initial band of 1980 m in range that is blind on all eight PRFs due to the dead time while transmitting the first pulse in each burst.

Figure 3-4. Medium PRF Receiver-Blanked Regions

Figure 3-5 shows the range coverage as a cumulative distribution. For example, 96.2% of the range is uneclipsed on at least five PRFs. Figure 3-5 also shows that 98.3% of the range is uneclipsed on at least four PRFs.

Figure 3-5. Percent of Range Covered by At Least X PRFs

While these eclipsing plots are useful in choosing PRF combinations and looking at the overall performance of the PRF set, they do not provide a complete representation of the range coverage of the mode. Consider, for example, the range coverage that is achieved with a 20 kHz PRF, or a 50 μ s PRI. The unambiguous range for this PRI is 7500 m. Barker encoding involves placing a binary phase code (0° or 180°) on different parts of each transmitted pulse. With a 13-bit Barker code and a 13 μ s pulse, each range gate is processed as if it were 1 μ s wide. The pulse-compressed output for the current range gate is then obtained through a summation of the current range gate output with the 12 subsequent range gate outputs. Each of the 12 is weighted with a 1 or a minus 1 prior to the summation.

Due to the 200 ns of dead time after pulse transmission to allow for switching, no results can be processed until at least 13.2 μs after the beginning of pulse transmission. Therefore, the range at which a target must be encountered in order to lie completely within the first range gate is 1980 m. The first range gate has a partial response, however, to targets lying closer to the radar, as the first range gate catches the tail end of the backscattered pulse from these closer targets. The 13 bits in the Barker code are such that pulse compression produces 13 times the amplitude normally received for a target lying exactly in the range gate, but produces range sidelobes that are only equal to zero or one times the amplitude. This effect is demonstrated in Figure 3-6 for ranges below 1980 m for a 13 μs pulse and 50 μs PRI. The graph is plotted to show the maximum response for a target as a function of target range out to R_u . In this case the interpulse period is 36.6 μs (including a total of 400 ns of dead time), so there are 36 possible pulse-compressed range gates. The response shown in this figure repeats itself over and over again out past R_u . Results are aliased back into the range-Doppler spectrum.

Notice that despite the fact that $R_u = 7500$ m, the maximum amplitude begins to decline after 5520 m. This is due to the fact that not all 13 samples are available for pulse compression when the trailing edge of the pulse begins to enter the eclipsed region of the second pulse. Indeed, the last range gate only has its own response with which to attempt to detect the target. Despite this, the 36th range gate can still provide some sensitivity to a target lying at 7380 m, performing just as well as if there were a 1 μs transmitted pulse (with the same amplitude as that of the 13 μs pulse) backscattered off this target.

Figure 3-6. Range Gate Response for MPRF RWS

The radar designer must decide the extent to which to process the data in each interpulse period is processed. When computational capability is limited, it may be advantageous to process only the range gates that achieve maximum response. For the CP-140 application, there is no reason to exclude any of the range gates, as computational power should be available, and inclusion of all range gates provides the maximum range coverage possible.

Repeating the coverage analysis for the Doppler spectrum, Figure 3-7 shows the number of PRFs unclipped at each point in the Doppler spectrum out to 65 kHz, based on increments of 1 Hz. In the case of the Doppler coverage, the blind zones occur in the areas of the spectrum occupied by mainlobe clutter.

Figure 3-7. Medium PRF Doppler Coverage

In each case the mainlobe clutter was shifted to DC, and a notch of ± 2 kHz was taken out of the spectrum. Thus, none of the PRFs cover the region from 0 to 2 kHz. Again, these Doppler-coverage plots are pessimistic. As mentioned earlier, Chapter 3 will demonstrate that it is worthwhile to provide some detection sensitivity in the strong sidelobe region. Thus, the blind zones in the Doppler spectrum are not truly ± 2 kHz wide.

Figure 3-8 shows the Doppler coverage as a cumulative distribution. It indicates, for example, that 89.7% of the Doppler spectrum is mutually covered by at least five PRFs, and 99.7% of the Doppler spectrum is covered by at least four PRFs.

Figure 3-8. Percent of Doppler Covered by at least X PRFs

The eight PRFs must be optimized to provide as complete a coverage of the range-Doppler spectrum as possible.. Judicious choice of the eight PRFs to be used on the CP-140 has provided good range and Doppler coverage.

Figure 3-9 plots the total coverage of the range-Doppler space based on range increments of 200 m and frequency increments of 100 Hz. There is a grayscale legend up the right hand side; white represents areas covered by eight PRFs, and black represents areas covered by none of the PRFs.

Figure 3-9. Range-Doppler Coverage for MPRF RWS

If it is stated that the MPRF RWS mode does not detect targets within 1980 m and will not detect slow-moving targets (within ± 2 kHz or within about ± 31 m/s radial velocity relative to the CP-140), then the coverage in the rest of the range-Doppler space can be investigated separately. From Figure 3-9, 97% of the range-Doppler space is covered by at least three PRFs, and 83% of the range-Doppler space is covered by at least four of the PRFs. Thus, targets in only 3% of the range-Doppler space should escape detection based on the range-Doppler eclipsing alone.

3.6 HIGH PRF RANGE-WHILE-SEARCH

3.6.1 PRF Selection

In HPRF operation, the PRF is chosen so that the target Doppler is unambiguous. The duty cycle of the transmitter is normally kept at its maximum for the greatest possible detection range. In HPRF mode, the radar normally searches only the noise-limited region for targets. In order to be in the noise-limited region, a target must be either closing or receding at a high enough rate to appear outside the sidelobe clutter. Discussion of HPRF design in the literature normally focuses on implementation in a fire-control radar aboard an AI aircraft where the radar antenna is forward-looking [2, 4, 13]. In this type of radar, there are very few cases where a target recedes at a rate such that the target appears with a negative Doppler frequency outside the sidelobe clutter for the central line of the spectrum. If the ownship were traveling at Mach 0.8, for instance, the target would have to be traveling with a radial velocity of at least Mach 1.6 away from the ownship (i.e. at Mach 2.4) in order to appear below the central clutter spectrum. Thus, for these radars, the HPRF modes only concentrate on the high closing velocity targets that appear with a positive Doppler shift.

The closing speed measured by the radar is a function of the ownship velocity, V_o , and the target velocity, V_T . In order to ensure that this target appears in the noise-limited region for an AI radar, the PRF must be chosen as shown in Figure 3-10.

Figure 3-10. Minimum PRF for an AI HPRF Mode

Implementation of a HPRF mode on the CP-140, however, is different from implementation on an AI aircraft. In a 360° azimuth search from a surveillance aircraft, the radar is likely to detect targets with both high closing and receding velocities. The HPRF modes must be designed with a different Doppler spectrum in mind, as shown in Figure 3-11. Note that on the CP-140, there is actually a small arc in the scan when the antenna is pointing aft where the transmitter is blanked to protect the aircrew from radiation.

Figure 3-11. HPRF for the CP-140

The PRF must be high enough that the radar can unambiguously detect the maximum closing and receding velocities. From Table 2-1, the maximum expected closing/receding velocities are $\pm$ Mach 3, or $\pm 1,026$ m/s. Assuming the maximum transmit frequency of 9.87 GHz, the HPRF must be at least 135 kHz. Once again, the duty cycle should be maximized to achieve the highest average transmitter power for the maximum detection range. The PRI of a radar operating with this PRF is 7.407 μ s.

Since the transmitter is capable of a maximum duty cycle of 30%, the pulse width should be set to 2.222 μs .

Because the HPRF RWS mode uses a duty cycle of 30%, almost one third of the range is eclipsed because of receiver blanking during pulse transmission. In order to mitigate this problem, HPRF modes usually use two different PRFs to minimize range blind zones due to eclipsing. The second PRF is staggered so that the range is eclipsed as little as possible for targets of interest.

The second PRF is chosen as follows. A range, R_E , is chosen at which it is acceptable for the target to be eclipsed on both PRFs. Then, the effective eclipsing repetition frequency, f_E , is calculated as follows [4]:

$$f_E = \frac{c}{2R_E} \quad . \quad (3-3)$$

The second PRF, PRF_2 , is then

$$\text{PRF}_2 = \text{PRF}_1 + f_E \quad . \quad (3-4)$$

A choice must be made, then, for R_E . As will be shown in Chapter 4, the probability of detection for HPRF RWS mode is expected to drop below 50% for a 5 m^2 target outside a range of 130 km. Thus, it is reasonable to choose $R_E = 130$ km, resulting in $f_E = 1,154$ Hz. The second PRF in this set, then, is 136.154 kHz.

3.6.2 Range Gate Placement

With a pulse width of $2.222\ \mu\text{s}$ and a PRI of $7.407\ \mu\text{s}$, there are several possibilities for the placement of range gates. Pulse Doppler radars often divide the interpulse period into range gates that are the size of the pulse width. When a target return arrives back at the radar such that the reflected pulse lies entirely within the chosen range gate, maximum SNR is achieved. If matched-width range gates are used with this transmitter, there are two range gates that are each set to 30% of the width of the PRI, and 10% of the PRI is not processed.

DREO has developed a simulator for advanced fighter radar (SAFIRE) to investigate ECCM techniques on AI radars. SAFIRE assumes that each range gate has the same temporal width as the transmitted pulse, and that the two range gates are contiguous, starting when the receiver is switched on. Another possible placement is shown in Figure 3-12 for range gates that have the same temporal width as the pulse. As the HPRF modes are normally used to detect long range targets, the second range gate will process returns from the farthest possible targets.

Figure 3-12. Possible Range Gate Placement in HPRF Mode

Very rarely, however, will the received pulse lie entirely within a range gate. Normally, there is some eclipsing and/or blanking of the received pulse. Fortunately, the average loss factor can be reduced by the use of overlapping range gates, at the expense of increased computational requirements [7]. Because greater computational capacity is available on the CP-140, the HPRF modes on the CP-140 can benefit from the use of three overlapping range gates. Their placement is shown in Figure 3-13. Note that the first and second range gates overlap by almost $1\ \mu\text{s}$, as do the second and third. In this manner, the full interpulse period is covered by at least one range gate.

Figure 3-13. Final HPRF RWS Range Gate Placement

3.6.3 Coherent Dwell

As with the MPRF RWS mode, it is desirable to have the longest coherent dwell possible in HPRF RWS. Given the slow antenna scan rate for MPRF RWS suggested in the previous section (4 rpm), it is feasible to have a long HPRF RWS dwell as well. A coherent dwell of 4096 pulses still allows for reasonable antenna scan rates, as discussed in Section 2.8.

Three different stages are used by the HPRF RWS mode for frequency modulation (FM) ranging, as described in the next subsection. Given the probability of

detection vs. range curves presented in Chapter 5, the HPRF RWS for the CP-140 should be designed around a maximum unambiguous detection range of approximately 500 km. Assuming that both HPRF RWS modes have the same average transmit power, the coherent dwell time in Equation 3-1 dictates the final range performance. The HPRF RWS on the CP-140 transmits 4096 pulses at 135 kHz, resulting in a coherent dwell of approximately 30 ms.

Based on an unambiguous range of 500 km, 3.333 ms of fill pulses must be transmitted at the beginning of a burst. At a PRF of 135 kHz, this results in $N_p = 450$. Thus, the dwell time for each stage in HPRF RWS is 33.7 ms. With three stages as outlined in the next subsection, the total dwell is equal to 101.0 ms.

3.6.4 FM Ranging

In order to obtain unambiguous range, the radar must determine the time delay between the transmitted and received signal. Since there are only three range gates in HPRF RWS mode, time delay cannot be measured accurately with range gating as it was in MPRF RWS. In order to calculate time delay in HPRF RWS modes, AI radars normally use frequency modulation (FM) ranging.

A three-stage modulation waveform is often used [4]. Only two stages are actually required to resolve a single target, but just as a third PRF is required in MPRF RWS to combat ghosting due to multiple targets, a third stage is used in the same manner for FM ranging. Stage one represents no modulation of the FM carrier, stage two represents a linear modulation of value $\dot{f}_2$, and stage three represents a linear modulation of value $\dot{f}_3$. One possible modulation sequence is shown in Figure 3-14. During stage one, the frequency shift of a backscattered target signal is simply the

Doppler shift, f_d , due to the relative motion of the target. During stage two, the backscattered signal is demodulated with a reference waveform having a linear FM of $\dot{f}_2$. This results in the target signal having a frequency shift equal to $f_d - \dot{f}_2 t_D$, where t_D is the round-trip time delay. During stage three, the frequency shift after appropriate demodulation is equal to $f_d - \dot{f}_3 t_D$.

Figure 3-14. FM Ranging Waveform

Since the Doppler shift is measured in stage one and the frequency modulation rates are known, it is possible to calculate the time delay, t_D , after the second stage. Given the Doppler filter in which the target appeared for each stage and knowing the filter bandwidths, the difference in frequency, Δf_{12} , measured in the two stages can be calculated. Then, the time delay is equal to

$$t_D = \frac{\Delta f_{12}}{\dot{f}_2} \quad (3-5)$$

For a target to be declared, it must be detected during all three stages and the detections must correlate to a common range. A more detailed explanation of range resolving using FM ranging can be found in [3, 4].

The range resolution of this FM ranging technique is dependent on the filter bandwidth:

$$\Delta R = \frac{c(BW_D)}{2\dot{f}_2} \quad , \quad (3-6)$$

where c is the speed of light, and BW_D is the Doppler filter bandwidth. Note that simply doubling the number of Doppler filters used for each PRF would reduce the Doppler filter bandwidth by two, thereby improving the range resolution by a factor of two.

There are several factors that must be considered in choosing the best $\dot{f}_2$ and $\dot{f}_3$ for a HPRF RWS mode. If there is a required range resolution for the HPRF RWS mode, $\dot{f}_2$ and $\dot{f}_3$ must meet this resolution. If the finest range resolution possible is desired, Equation 3-6 would have the designer choose as high a modulation rate as possible. The modulation rate is limited, however, by the total frequency range of the transmitter. If the modulation scheme shown in Figure 3-14 is used, the starting frequency for the third phase is the same as the ending frequency for the second phase. In this case, the total frequency change over the second and third stages cannot exceed the total frequency range.

Another factor that must be taken into consideration is the transmitter power. The transmitter peak power may vary according to the selected frequency. It is desirable to have a fairly flat response across the entire range in use so as to readily compare outputs from the three phases. This may require limiting the FM rate to a smaller value. There may also be an upper limit as to what FM rate can be handled by the transmitter.

Finally, the frequency modulation rate should be chosen to account for multiple CP-140s in the same vicinity. It might be desirable to split the usable frequency ranges into smaller portions, allowing each aircraft a specific range or a specific bank of frequencies. Each frequency range would have to be wide enough to account for the FM sweep and the bandwidth of the transmitted pulses. This would minimize mutual interference (MI) between the aircraft.

Consider, for example, FM rates of $\dot{f}_2 = 4$ MHz/s and $\dot{f}_3 = 2$ MHz/s. With 4096 pulses sent in each phase at a PRF of 135 kHz, this corresponds to a total change in frequency of approximately 182 kHz. The 3-dB bandwidth of the 2.222 μ s pulse is approximately 450 kHz. Typical off-the-shelf X-band transmitters can operate over a much wider range of frequencies, thereby allowing any CP-140s in close proximity to choose different frequency banks. Using a PRF of 135 kHz, a burst length of 4096 pulses, and $\dot{f}_2 = 4$ MHz/s with Equation 3-6 results in a range resolution of 1236 m. Final values for $\dot{f}_2$ and $\dot{f}_3$ must be chosen based on the characteristics of the selected transmitter.

3.6.5 Range and Doppler Coverage

Based on the PRF set of 135 kHz and 136.154 kHz chosen in Subsection 3.6.1, the range coverage of the HPRF RWS mode can be calculated. Recall that $R_E = 130$ km. Assuming zero response for targets that lie at a range corresponding to the transmission of a pulse, the range coverage is depicted in Figure 3-15 for this PRF set. Note that the coverage shown here would repeat past 130 km.

Figure 3-15. Range Coverage for PRF Set in HPRF RWS

Note that this staggering of the PRFs has allowed a region between 40 and 90 km to be covered by at least one if not both PRFs. The region between zero and 40 km and the region between 90 and 130 km experience some eclipsing in range.

Despite the somewhat pessimistic view of the range coverage that the eclipsed regions show, the true range coverage is somewhat better. Consider, for example, a target that lies at 3.6 km. The round-trip transit time for each pulse backscattering off this target is $24 \mu\text{s}$. With a PRI of $7.407 \mu\text{s}$, the beginning of this pulse arrives while the receiver is being blanked for the fourth transmitted pulse, and the contribution of this PRF to Figure 3-15 would be zero. The response to a target at this range, however, is not zero. Energy from the trailing edge of this backscattered pulse is received and processed by the first range gate in the fourth interpulse period. Unlike the MPRF RWS mode, no phase coding is applied to the HPRF RWS pulse. Thus, the first range gate provides some response for the trailing edge of the pulse.

The first range gate in the second interpulse period starts processing data 24.643 μs after the beginning of a burst. Thus, 1.579 μs worth of backscattered energy is actually processed by the first range gate even though the range of the target was such that the leading edge of the backscattered pulse arrived during a pulse transmission. This phenomenon is demonstrated in Figure 3-16.

Figure 3-16. Range Gate Response for HPRF RWS

Because HPRF modes notch out the mainlobe and sidelobe clutter, the Doppler coverage for the HPRF RWS mode depends on the groundspeed of the aircraft. At the maximum ownship groundspeed of 500 kts, the sidelobe clutter extends out to approximately ± 16.5 kHz. Thus, no targets are detected if their relative radial velocities lie between minus 500 and 500 kts. If the ownship is traveling slower, the clutter notch narrows accordingly.

3.7 VELOCITY SEARCH MODE

3.7.1 PRF Selection

As the VS mode is another HPRF mode, most of the characteristics of the HPRF RWS selection apply here. The PRF is driven by the same requirement to be able to detect a maximum closing/receding velocity of $\pm$ Mach 3. Based on this requirement alone, the PRF set for VS would be the same set as was chosen for HPRF RWS. In selecting the PRF for VS mode, however, the interrelationship of the antenna scan rate, the time-on-target (TOT), and the burst length must be considered.

In order to achieve the desired burst length described in Subsection 3.7.3 for the TOT described in Section 3.8, a slightly higher PRF of 170 kHz must be used. Increasing the PRF over that required for the Mach 3 targets results in an increase in the clear region between the repeated Doppler spectra. At 170 kHz, the VS mode can actually detect targets closing or receding at approximately Mach 3.9. Once again, the duty cycle should be maximized to achieve the highest average power for the maximum detection range. The PRI of a radar operating with this PRF is 5.882 μ s. The pulse width is set to 1.765 μ s to achieve the maximum 30% duty cycle that the transmitter is capable of sustaining.

Like the HPRF RWS mode, a second PRF is normally staggered with the first to provide some improvement in range eclipsing. The second PRF is staggered so that the range is eclipsed as little as possible. A choice must be made, then, for R_E . As will be shown in Chapter 4, the probability of detection for VS mode is expected to drop below 50% for a 5 m² target outside a range of 150 km. Thus, it is reasonable to

choose $R_E = 150$ km, resulting in $f_E = 1,000$ Hz. The second PRF in this set, then, is 171 kHz.

3.7.2 Range Gate Placement

With a pulse width of $1.765 \mu\text{s}$ and a PRI of $5.882 \mu\text{s}$, there are several possibilities for the placement of range gates. As with the HPRF mode, three overlapping range gates are used to cover the entire available interpulse period. Recall that a dead time of 200 ns is left before and after the transmitted pulse to allow for transmitter switching time. The final range gate placement is shown in Figure 3-17.

Figure 3-17. Final VS Range Gate Placement

3.7.3 Coherent Dwell

As with the RWS modes, it is desirable to have as long a coherent dwell as possible in VS mode. Given the slow antenna scan rate for the RWS modes, it appears feasible to lengthen the VS mode dwell as well. Using a coherent dwell of 8192 pulses still allows for reasonable antenna scan rates, as discussed in Section 3.8. Given the probability of detection vs. range curves presented in Chapter 5, the VS mode for the CP-140 should be designed around an unambiguous range of approximately 550 km.

Based on an unambiguous range of 550 km, 3.67 ms of fill pulses must be transmitted at the beginning of a burst. At a PRF of 170 kHz, this results in $N_p = 624$. Thus, the dwell time in VS is 51.9 ms. Raising the PRF in VS to 170 kHz has allowed the total dwell to be equal to approximately half the length of that required for MPRF and HPRF RWS. This allows the transmission of both PRFs for VS within one time on target, which is an extremely desirable feature. This issue will be further discussed in Section 3.8.

It is important to reiterate that the only influence that the maximum unambiguous detection range has on this stage of the design is in determining the number of fill pulses used. Even if no fill pulses were used in VS mode, the target detection capability at the maximum unambiguous range would only be missing backscatter from 624 of the total 8,192 pulses. In reality, most targets of interest that are detected by this mode are at much shorter ranges, and do not require nearly as many fill pulses.

3.7.4 Range and Doppler Coverage

In VS mode, the return signal is divided into three range gates, and no further attempt is made to determine the target's range. Only velocity and azimuth information are displayed to the pilot or radar operator. It is important to consider, however, the kind of coverage that is provided for the possible ranges at which targets may be located.

Based on the PRF set of 170 kHz and 171 kHz, the range coverage of the VS mode can be calculated. Assuming zero response for targets that lie at a range

corresponding to the transmission of a pulse, the range coverage is depicted in Figure 3-18 for this PRF set.

Figure 3-18. Range Coverage for PRF Set in VS

Note that this staggering of the PRFs has allowed a region between 45 and 105 km to be covered by at least one if not both PRFs. The regions between zero and 45 km and the regions between 105 and 150 km experience some eclipsing in range.

Again, despite the somewhat pessimistic view of the range coverage that the eclipsed regions show, the true range coverage is somewhat better. The three range gates can process backscattered signal from targets even if the pulse does not lie exactly in the range gate. The true range gate response is demonstrated in Figure 3-19 for one unambiguous range interval.

Figure 3-19. Range Gate Response for VS

The Doppler coverage of the VS mode is limited by the same factors as for the HPRF RWS mode. Again, because high PRF modes notch out the mainlobe and sidelobe clutter, the Doppler coverage for the VS mode depends on the groundspeed of the aircraft. At the maximum ownship groundspeed of 500 kts, the sidelobe clutter extends out to approximately ± 16.5 kHz. Thus, no targets are detected with relative radial velocities between minus 500 and 500 kts. If the ownship is traveling slower, the clutter notch narrows accordingly.

3.8 ANTENNA SCAN PATTERNS

While AI radars normally scan only the forward sector, a surveillance radar often scans a full 360° in azimuth. A typical air-to-air search mode employs interleaved MPRF and HPRF antenna scan bars to search for all types of targets. In a downlook scenario, the current antenna elevation scan limits are from 0° to minus 15° .

With a 4.5° elevation beamwidth, an elevation bar spacing of 2° would provide an overlap of both the HPRF and MPRF scans. An example of a six-bar scan pattern is shown in Figure 3-20. With this scan pattern, both the MPRF and HPRF waveforms cover 86% of the 14.5° elevation extent.

Figure 3-20. Interleaved PRF, Six Bar Scan

The advantage of this relatively small beamwidth is that the antenna gain is concentrated toward a small illuminated region. The disadvantage for an air-to-air surveillance aircraft is that a smaller beamwidth takes longer to search a potentially large volume. Some AEW aircraft use a fan-beam antenna pattern that is narrow in azimuth and much wider in elevation. This allows the radar to search larger volumes over a given period of time without requiring separate bars to scan in elevation. If a target is detected, the radar can switch to a pencil-beam type pattern to determine the elevation of the target. Since the antenna pattern is fixed for the APS-506 radar, the modes are designed based on this pattern.

AI radars typically use rapid antenna scan rates on the order of 60° per second. This allows the AI radar to process and detect targets in the most demanding scenarios. AEW radars, on the other hand, provide a surveillance role from a much slower moving (and less dynamic) platform. Typical AEW antenna scan rates are around 5 rpm ($30^\circ/\text{s}$), providing complete azimuthal coverage every 12 seconds. It is desirable to have a common antenna scan rate for all modes when in normal search operation.

The coherent dwell in MPRF mode of 128 pulses requires a total length of time required on target to 104.8 ms. Given a 3 dB azimuthal beamwidth of 2.5° , an antenna scan rate of 4 rpm results in a TOT of 104.2 ms, almost accommodating the full eight PRFs from the MPRF set (3.98 rpm would be required to catch all eight within the 3 dB beamwidth).

The three stages of the HPRF RWS set require a total TOT of 101.0 ms, which is provided nicely by the 4 rpm scan rate. Because there is no time to alternate between the HPRFs during one TOT in HPRF RWS, different PRFs should be used on alternating bars. This method is suboptimal in terms of eclipsing performance. This represents a tradeoff, however, between the length of the coherent integration and the ability to reduce target eclipsing.

Recall that in VS mode, the PRF is increased slightly over the HPRF RWS case, so that one burst of pulses requires a total of 51.9 ms. This allows time for the radar to transmit a second burst at the second PRF to minimize eclipsing. The total dwell with the two bursts is 103.8 ms, falling within the 104.2 ms of the 4 rpm scan rate.

The scan rate of 4 rpm can be used as a basis for the search modes. As noted earlier, it is close to the 5 rpm quoted for some other AEW radars. In situations where a target must be quickly acquired, it may be desirable to increase the antenna scan rate or decrease the antenna search volume. A doubling of the antenna scan rate would not be recommended for the current structure of the RWS modes, as each requires the full TOT to get reasonable performance. If it were determined that a doubling of the antenna scan rate would be desired often, the radar could be programmed to halve the number of pulses integrated in each of the RWS modes for an 8 rpm scan rate. The detection range would not be as great, but both modes would operate correctly. A doubling of the antenna scan rate for the VS mode would have less drastic ramifications. The VS mode would still have a full PRF burst transmitted during the TOT, although the single burst would result in additional range eclipsing as compared to the two-burst 4 rpm case. If the approximate location of the target were known, another option would be to reduce the antenna scan volume. Halving the scan volume would result in the same target revisit time as doubling the scan rate and keeping the full scan volume.

3.9 SUMMARY OF HIGH-LEVEL DESIGN

The modes outlined in this chapter provide a baseline design for the CP-140 air-to-air detection capability. Final decisions on the parameters in each mode will incorporate operator input. Figure 3-21 shows the final burst structure during the TOT for the air-to-air modes. N_p represents total number of fill pulses, and N represents the total number of coherently integrated pulses for each mode. Note that N_p is not listed for the MPRF RWS mode, as the selection of the medium PRFs is outside the scope of this thesis.

Figure 3-21. Burst Structure for Air-to-Air Modes

The MPRF RWS mode cycles through eight different PRFs during the TOT. A target must be detected on at least three PRFs for range and Doppler resolving.

The HPRF RWS mode cycles through three different frequency-modulated stages at a single PRF during one TOT. A target must be detected on all three stages in order to carry out range resolving. HPRF RWS mode alternates between two different PRFs for alternating scan bars.

The VS mode cycles through two different PRFs during the TOT. A target need only be detected on one of the two PRFs to be declared.

CHAPTER 4: RADAR SIGNAL PROCESSING

The relationships between radar signal processing algorithms and the corresponding radar subsystem requirements are particularly critical in an airborne pulsed Doppler radar due to the presence of a complex time-varying and spatially-varying interference process composed of thermal noise, clutter, target returns, and potential jamming. The next section reviews signal processing techniques used in a pulse Doppler radar. Section 4.2 investigates *constant false alarm rate* (CFAR) techniques. Section 4.3 outlines a CFAR technique designed specifically for the air-to-air modes on the CP-140.

4.1 PULSE DOPPLER RADAR SIGNAL PROCESSING

When no clutter or jamming is present, the detection performance of a radar system is ultimately limited by the thermal noise generated in the antenna and receiver. In the noise-limited region of the range-Doppler spectrum, maximum detection range is determined by the ratio of the energy received from a target to the energy of the background noise. In clutter-limited detection, the total clutter return power in the range-Doppler cell is normally greater than any other interference source at the receiver output. Detection range in clutter-limited cells varies with the relative geometry of the target, the radar platform, and the clutter sources. Detection range can degrade significantly due to clutter or jamming. Detection implementation and performance characteristics vary significantly among LPRF, MPRF, and HPRF radar modes.

4.1.1 Coherent Integration

Detection can be considered to be a three-stage process in a modern pulse Doppler radar system [4]. The first stage includes all analog components in the receiver up to and including the *analog-to-digital converter* (ADC). The second stage involves the radar digital signal processing. The third stage involves target tracking and the interface to the radar operator.

In the first stage, the analog signal received at the antenna is coherently downconverted by the receive chain. The digitized baseband versions of the in-phase and quadrature (I and Q) components of this signal are obtained through one of two methods: by mixing the received signal to baseband and sampling with two ADCs, or by mixing the received signal to an *intermediate frequency* (IF) and sampling with a single ADC. In both cases, all stages of the downconversion are done with phase-locked *coherent oscillator* (coho) outputs. In the former, the I and Q components of the signal are formed by mixing the received signal with a coho signal and a signal that is 90° out of phase with the coho (analog demodulation). In the latter case, the I and Q components are formed as the sampled signal is digitally demodulated to baseband (digital demodulation).

With the advent of high-speed, high-resolution ADCs, digital demodulation is becoming more and more common, as it better preserves the information contained in the received signal. Digital demodulation alleviates the need for a second ADC, and avoids the images generated when there is an imbalance in the I and Q analog channels [14]. In older radar systems an analog matched filter is applied to the signal, while in newer systems, digital matched filtering can be performed. Thus, digital demodulation

offers more flexible filtering as well. Digital demodulation is planned for the CP-140 implementation.

Once the received signal is digitized, it is divided temporally into range gates. One I and one Q sample are normally stored for each range gate. This process is then repeated for each interpulse period. The I and Q components of each range gate can be considered as forming a complex phasor. Since frequency is defined as $d\phi/dt$, where ϕ is the phase shift, the phase relationship between phasors in the same range gate of different interpulse periods determines the Doppler shift. The discrete Fourier transform (DFT) is a digital filter bank that can measure this phase relationship. The combination of pulse-matched filtering and coherent pulse integration constitutes a matched filter for the coherent pulse train. Coherent integration of a nonfluctuating target with no signal processing losses provides a gain of N in the SNR. The DFT can be expressed in the following form [4]:

$$S(n) = \sum_{k=0}^{N-1} X(k) W^{nk} \quad , \quad (4-1)$$

where

$X(k) = e^{j2\pi f k T} = e^{j2\pi f k / PRF}$, a series of complex samples of the input time waveform,

f = the Doppler frequency of the input waveform,

T = the pulse repetition interval,

$W = e^{-j2\pi/N}$,

N = the number of time samples and the number of Doppler filters,

n = the filter number, and

k = the time sample number.

If the number of I and Q samples N in a particular range gate is some power of two, this can be implemented as a computationally efficient *fast Fourier transform* (FFT). The FFT effectively forms a digital filter bank of N filters, each with a magnitude response in the form of a $(\sin x)/x$ function. If the PRI is equal to T , each of these filters has a passband equal to approximately $1/NT$. The null-to-null bandwidth of the filter is $2/NT$.

If no windowing is used, each filter will have its first sidelobes 13 dB down from the mainlobe. This is sometimes referred to as a rectangular window, because it can be considered that each sample is weighted with a unity multiplier. These sidelobe levels are normally unacceptable, since they may cause a large return to mask a smaller return at the same range several filters away. To alleviate this problem, a window is normally applied to the data before taking the FFT. The windowing results in a reduction in the sidelobe level of each filter while broadening the 3 dB bandwidth of the filter mainlobe. Weighting functions are chosen to attenuate the filter sidelobes to acceptable levels while maintaining as narrow a mainlobe as possible.

The FFT of the data is taken for each range gate. After detection of the output of the FFT, the range-Doppler map consists of data in N Doppler bins for each range gate. Linear or envelope detection is commonly used, which involves taking the square root of the summed squares of the real and imaginary components of the signal. Alternatively, square-law detection is simply the summed squares of the real and imaginary components of the signal. The intersection of a particular range gate with a particular Doppler bin is often termed a *range-Doppler cell*. The FFT algorithm

performs optimally with a target return signal that has a constant phase shift from sample to sample. This is true only for a constant radial velocity target. If, however, the target is accelerating, the amount of the phase shift will differ from pulse to pulse, and the Doppler filters will not be matched to the return signal. The higher the acceleration, the more mismatch will occur, and the more the target will spread in Doppler. In extreme cases such as an engagement between AI aircraft or between an aircraft and a missile, the relative ownship/target radial acceleration might be enough that the target return moves across several Doppler bins during the *coherent processing interval* (CPI). When this happens, it is possible that the SNR will drop below that required to detect the target. Several options are available if it is expected that there will be a significant radial acceleration. If the acceleration is known, the phasors can be derotated prior to FFT processing. Alternatively, time-frequency techniques that are matched to targets with certain accelerations can be used.

For a large platform such as the CP-140, there will rarely be engagements where the radar will measure a high target radial acceleration. In a search mode the ownship normally travels at a constant velocity or with mild acceleration, so any radial acceleration would normally come from the target's movements. Even if the target were accelerating, however, the coherent processing intervals are relatively short, meaning that the total change in target velocity would be small for one CPI. Approaching the worst case scenario, assume, for example, a relative radial target acceleration of two times the acceleration of gravity (two "g"s = 19.62 m/s^2), and the maximum CPI (during VS mode) of $8192/170000 = 48 \text{ ms}$. Although AI aircraft are capable of acceleration above two "g"s, they only experience this much acceleration infrequently. When they do, it is normally for a short duration, and it is likely that

only a component of this acceleration would be along the radial line to the surveillance aircraft. At a relative radial acceleration of two "g"s, the target's velocity will shift by approximately 1 m/s over the interval, resulting in a Doppler frequency shift of $2(\Delta V_T)/\lambda = 60.5$ Hz. Since each Doppler bin is $170000/8192 = 21$ Hz wide, the target will move approximately three bins during the interval. Thus, the target signature is only integrated within each Doppler bin for about one third of the CPI. Table 4-1 repeats this calculation for HPRF RWS and MPRF RWS. For HPRF RWS, the signal will be spread at worst into an adjacent Doppler bin. In MPRF RWS, the signal will normally stay well within one Doppler bin.

Table 4-1 Doppler Spread for a 2g Target

<i>Mode</i>	<i>Coherent Dwell (ms)</i>	ΔV_T (m/s)	Δf_T (Hz)	<i>Doppler Bin Width (Hz)</i>
VS	48	0.95	61	21
HPRF RWS	30	0.59	38	33
MPRF RWS	9	0.18	11	110

For high-acceleration targets, the long coherent integration in VS mode provides no added SNR gain over a CPI that is three times smaller. Despite this, the Doppler spread is not excessive. In most cases, the radial velocity is expected to be nearly constant, allowing for coherent integration of the signal in one Doppler bin for the full CPI. Thus, it is reasonable to keep this long CPI for the VS mode. A full coherent integration of 8,192 pulses provides the CP-140 air-to-air modes with a detection range that would not otherwise be achievable with this configuration. However, the Doppler spreading that can occur for high acceleration targets highlights

the need to avoid the cells directly surrounding the range-Doppler cell under test in any constant false alarm rate technique, as will be discussed in the next section.

4.1.2 Noncoherent Integration

The VS mode could potentially benefit from employing noncoherent integration (or post-detection integration) techniques. If it were known that the VS mode would be required to detect targets with two "g"s of radial acceleration quite often, it might be advantageous to avoid the Doppler spreading by using a shorter coherent integration period. The dwell time could be divided, for instance, into four CPIs of 1,024 pulses. After coherent integration with the FFT, the data from the four CPIs could then be noncoherently integrated, thereby avoiding the Doppler spreading. Each Doppler bin would be four times as wide, and the target would more readily stay within one Doppler bin. Noncoherent integration provides a smaller SNR gain than coherent integration, however, because the phase information is lost. Noncoherent integration provides an SNR gain between $\sqrt{N}$ and N , in contrast to the gain of N with coherent integration [1]. For a constant velocity target, coherent integration would provide about 39 dB of SNR gain, while the coherent/noncoherent combination suggested above might only provide about 36 dB. Thus, for a constant velocity target, this method might represent a SNR loss of up to 3 dB over the full coherent integration.

4.1.3 Detection in Thermal Noise

Detection is based on establishing a threshold level after the FFT. The threshold forms the dividing line between one of two hypotheses: that the receiver output $r(t)$ is due to noise $n(t)$ alone, or that the output is due to the target signal $s(t)$ plus noise. Symbolically:

H_0 : no signal present [$r(t) = n(t)$]

H_1 : signal present [$r(t) = s(t) + n(t)$]

There are two types of errors that may occur. A type I error is to decide that there is a signal present (D_1) when only noise is present. In the field of radar this is called a false alarm. The probability of false alarm, P_{fa} , is the probability that a type I error is committed times the probability that there is no signal present. Symbolically, this is $P(D_1|H_0)P(H_0)$. A type II error is to decide that there is only noise present (D_0) when a target return is actually present ($P(D_0|H_1)P(H_1)$). This is often termed a missed detection. The average probability of error P_e is:

$$P_e = P(D_1|H_0)P(H_0) + P(D_0|H_1)P(H_1) \quad . \quad (4-2)$$

The threshold represents a compromise between the two types of errors. The probability of detection (that is, we say that the target is present when it actually is present), P_d , is equal to $P(D_1|H_1)P(H_1)$.

The detection process is as follows. The I and Q components of the receiver output $r(t)$ at the i th range gate and j th Doppler filter can be expressed as the following when the interference is a zero-mean band-limited Gaussian process [4]:

$$\begin{aligned} I_{ij} &= A \cos(\theta) + x_{ij} \\ Q_{ij} &= A \sin(\theta) + y_{ij} \end{aligned} \quad , \quad (4-3)$$

for a return magnitude A and a relative phase θ at the output of the receiver. The noise components, x_{ij} and y_{ij} , are independent zero-mean random variables with variances σ_x^2 and σ_y^2 . Each is equal to one-half the noise power, P_n , such that

$$2\sigma_x^2 = 2\sigma_y^2 = P_n = B_n k T_s = \sigma^2, \quad (4-4)$$

where

B_n = the receiver noise bandwidth,

k = Boltzmann's constant, and

T_s = the equivalent system temperature (K).

Based on the received signal, some criterion must be selected for determining the threshold. In some communication systems the *a priori* probabilities of the two hypotheses are known (for instance, the probability of receiving a 0 or a 1), and each type of error can be assigned a certain weight or cost. The detection system is often designed to minimize the total cost, and is known as *Bayes' criterion* [15]. A subset of the Bayes' criterion involves assigning an equal cost to both types of errors. In this case the average error probability is minimized, and the threshold is set based on the *minimum error probability criterion*.

In a radar system, however, the *a priori* probabilities are difficult or impossible to determine. For instance, it is not possible to determine the *a priori* probability that there will be an airborne target in the beam at any one point in time. For such cases there is another criterion which involves neither *a priori* probabilities nor cost estimates. It is the Neyman-Pearson criterion, and is the most commonly used criterion for radar systems [4, 11]. It works with the conditional probabilities P_{fa} and P_d rather

than cost functions or *a priori* probabilities. Its objective is to maximize the probability of detection for a given probability of false alarm. To determine these probabilities, the statistics of the noise and target returns must be considered.

The probability density function of the magnitude, m , in both the in-phase and quadrature channels due to Gaussian noise alone is given by

$$f_{gn}(m) = \frac{1}{\sqrt{2\pi\sigma_x^2}} e^{-m^2/2\sigma_x^2} \quad . \quad (4-5)$$

After coherent integration and envelope detection, the magnitude of a range-Doppler cell that contains only noise has a Rayleigh *probability density function* (pdf) [4]:

$$f_n(m) = \frac{m}{\sigma_x^2} e^{-m^2/2\sigma_x^2} U(m) \quad , \quad (4-6)$$

where $U(m)$ is the unit step function; $U(m)=0$ for $m<0$, $U(m)=1$ for $m\geq 0$. Note that a quadratic detector obtains the square of the envelope. If a quadratic detector were used instead of an envelope detector, the output would be proportional to the IF power $x = m^2$, having an exponential pdf [1]:

$$f_n(x) = \frac{1}{2\sigma_x^2} e^{-x/2\sigma_x^2} U(x) = \frac{1}{\sigma^2} e^{-x/\sigma^2} U(x) \quad . \quad (4-7)$$

If a sinusoidal target return $A \cos(2\pi f_t t)$ is introduced at the receiver, the pdf of the envelope-detected magnitude of a cell containing the target return plus noise is Rician [4]:

$$f_{sm}(m) = \frac{m}{\sigma_x^2} I_0\left(\frac{Am}{\sigma_x^2}\right) e^{-(m^2 + A^2)/2\sigma_x^2} U(m) \quad , \quad (4-8)$$

where $I_0(\cdot)$ is the zero-order modified Bessel function. In the absence of a target return, $A=0$, and the Rician density is equal to the Rayleigh density. For a high SNR, the Rician density approaches a Gaussian density [4]. The probability density functions of noise alone and signal-plus-noise are depicted, along with one possible threshold, in Figure 4-1 for envelope detection with $A=3$ and $\sigma_x = 1$.

Figure 4-1. Threshold Detection

The probability of detection is calculated as the area under the signal-plus-noise curve to the right of the threshold α :

$$P_d = \int_{\alpha}^{\infty} f_{sm}(m) dm \quad . \quad (4-9)$$

The probability of false alarm is represented by the shaded area under the noise only curve, to the right of the threshold:

$$P_{fa} = \int_{\alpha}^{\infty} f_n(m) dm \quad . \quad (4-10)$$

Thus, decreasing the P_{fa} by moving the threshold to the right will result in a decrease in P_d . It is also true that increasing the target's SNR will shift the signal-plus-noise graph to the right, thereby increasing the P_d for any given threshold. The decision threshold is set to perform at a *constant false alarm rate* (CFAR) so as to generate detection reports at a manageable rate. CFAR techniques will be discussed in Section 4.2.

4.1.4 Detection in Clutter

One principal factor limiting target detection in a MPRF mode is ground clutter. Because MPRF pulse Doppler waveforms are included in the design of air-to-air modes for the CP-140, detection performance in sidelobe clutter must be considered. The radar characteristics of the MPRF mode constitute an integral component of the clutter modelling process.

Recall from Equation 3-2 that for a given ownship velocity V_o , the Doppler frequency of the ground clutter return due to the motion of the ownship is a function of the radar look angle. The antenna azimuth and elevation beamwidths determine the size of the patch of ground that is illuminated by the mainlobe of the antenna at a particular look angle. The size of this clutter patch changes depending on the look angle.

The loci of points having a constant apparent Doppler due to the platform motion are cones of revolution about the velocity vector. If the aircraft is in level flight, the intersection of the cone and the earth plane is a hyperbola. All points located along each hyperbola produce returns at the same Doppler frequency, and are termed *isodops*. Equal ranges from the radar, on the other hand, may be visualized as a series of concentric spheres centered on the radar. In level flight, the clutter return at a particular range comes from the circular region where the concentric sphere intersects the ground. The concentric regions are termed *isoranges*. Due to the range and Doppler ambiguities in MPRF mode, many different regions of ground clutter may contribute to the same final range-Doppler cell output.

The clutter RCS measured at the output of a single Doppler filter of the receiver is proportional to the product of the transmitter power and two-way antenna gain multiplied by the product of the clutter patch area, A_c , and the surface backscattering coefficient, σ° . The total clutter power, C , can be defined as the summation of an ensemble of clutter patches such that [4]

$$C = \frac{P_t G A}{(4\pi)^2 L_{cs}} \sum_i \frac{A_{ci} (\sigma^\circ)_{ci} (g_{sli})^2}{(R_{ci})^4} , \quad (4-11)$$

where

P_t is the transmitter peak power,

G is the antenna gain,

A is the receiving antenna area,

σ° is the backscatter coefficient,

R_c is the range to the clutter patch,

g_{sl} is the sidelobe gain (relative to the mainlobe gain),

L_{cs} is the clutter system-loss, and

A_c is the clutter patch area.

Note that for most grazing angles and pulse widths, $A_c \propto R$, so $C \propto R^{-3}$. The backscatter coefficient, σ° , is defined as the radar reflectivity per unit area (m^2 / m^2). This parameter varies significantly with terrain, polarization, frequency, and grazing angle. The reflection coefficient can be approximated as:

$$\sigma^\circ = \gamma \sin(\psi) \quad (4-12)$$

over a wide angular extent where ψ is the grazing angle to each clutter patch. This approximation is inaccurate as the grazing angle approaches either 0° or 90° . Typical values for γ are listed in Table 4-2 [4]. For a given frequency, polarization, and grazing angle, sea clutter can be reasonably well related to sea state [4]. Sea state is a measure of the local sea roughness. Urban areas have a high backscatter coefficient and a relatively large number of specular reflectors, such as human-made metallic surfaces.

Table 4-2 Typical Values for γ at X-band

<i>Surface</i>	<i>γ, dB</i>
Smooth water	-45
Desert	-12
Wooded area	-10
Urban	-3

Pulse Doppler radar performance modelling requires calculation of the composite clutter return given in Equation 4-11 for the ambiguous range-Doppler output space defined by the radar parameters. The simulator SAFIRE described in the next section calculates σ^0 for each clutter patch based on Equation 4-12 with a value for γ that is appropriate for vegetation clutter. The Box-Muller algorithm is then applied to give the clutter return a Gaussian distribution, with variance proportional to σ^0 .

Comment should be made about the validity of using the method described above to model clutter for prediction of the performance of the air-to-air modes on the CP-140. Considerable effort has been spent over the last twenty years to characterize clutter as seen by a radar system. Distributions that are commonly used to describe the amplitude variations of echo from clutter include Rayleigh, Ricean, log-normal, Weibull, and K -distributed [16, 17]. Recall that the Rayleigh distribution accurately describes the thermal noise contribution. In MPRF modes, there will normally be many different clutter patches that contribute to a single range-Doppler cell due to range and Doppler ambiguities. If clutter is received from a large number of scatterers of nearly equal magnitude, the clutter spectrum in each cell can be assumed to have an exponential power distribution (equivalent to a Rayleigh amplitude distribution) as well [18]. Thus, using Equation 4-12 and an isotropic antenna, SAFIRE would produce Rayleigh-like sidelobe clutter returns for an MPRF mode.

Sea clutter appears non-Rayleigh for low grazing angles and pulse lengths of 0.5 μ s or less, and is often modelled as K -distributed [17]. With a short pulse length the radar is able to illuminate specific features of the sea for a given sea state, and the assumption of a large number of equal-amplitude scatterers is violated. The shorter the

pulse length, the less the clutter appears Rayleigh-distributed. Thus, for a high-resolution radar, the use of Equation 4-12 does not accurately characterize the clutter return, and will not result in a valid performance prediction. For the CP-140 application, however, the compressed pulse lengths for the air-to-air modes are at least 1 μ s. The *K*-distribution approaches the Rayleigh distribution for these pulse lengths, and Equation 4-12 provides an accurate calculation of σ^0 [33].

If the clutter were backscattered from one large scatterer and many small scatterers, the clutter return would be better approximated by a Rician distribution. Recall that the Rician distribution is also assumed when there is a target return embedded in a thermal noise background. This type of clutter spectrum would arise if there were a single large-RCS scatterer on the ground illuminated by the radar. It would also arise if the illuminating antenna had large sidelobes such that equal-RCS ground scatterers did not result in equal-magnitude contributions to the clutter return. Because SAFIRE uses the actual antenna pattern for the specific radar antenna, this phenomenon is modelled even with the current method. Thus, it should not be expected that the simulated sidelobe clutter returns will neatly conform to Rayleigh statistics. Indeed, as will be seen in the next section, they do not.

Therefore, the method of clutter calculation used in SAFIRE is a reasonable one to predict the performance of the air-to-air modes on the CP-140. The results will characterize the performance of the modes over unpopulated areas of land and sea. However, the contribution of independent large-RCS ground scatterers is not modelled. It is reasonable to assume that over urban areas, the false alarm rate of the radar will be higher for the same threshold setting used over unpopulated areas in MPRF RWS mode.

4.1.5 Target Fluctuation Effects

Virtually all targets of interest for the CP-140 have dimensions that are much greater than the radar wavelength. In addition, the range resolution cells for each waveform are, for the most part, larger than the target dimensions. Radars using very short pulse widths and range gates are sometimes able to distinguish specific features of a complex target. This will not be the case for the air-to-air modes for the CP-140. Again, the smallest range resolution cell for the modes recommended in this thesis is 150 m, and occurs with the MPRF waveform. Thus, for all targets of interest, the backscattered signal is normally the sum of returns from many individual scattering centers. Variation among the scatterers in a single range gate results in coherent constructive and destructive addition. As a result, the RCS of a complex target varies with aspect angle.

Because of this, the RCS variation associated with a complex scatterer target such as an aircraft undergoing arbitrary motion is not readily modelled in a deterministic fashion. It is customary to employ statistical RCS fluctuation models that emulate the RCS variation anticipated with the type of target to assess radar system performance. To accomplish this, the target RCS is typically modelled as a random variable with a specified pdf.

In his classic paper [20], Swerling developed four target RCS models. These models have stood the test of time, and are still commonly used today for radar performance prediction. The four RCS models are based on combinations of two pdfs and two temporal autocorrelation conditions. For cases 1 and 3 ("slow" fluctuation), the temporal conditions are such that the target RCS is completely correlated from pulse to pulse, but fluctuates from scan to scan. For cases 2 and 4 ("fast" fluctuation),

the target RCS is uncorrelated from pulse to pulse. The nonfluctuating or constant RCS model is often termed case 0 or case 5.

Both pdfs are based on the χ^2 family. Swerling cases 1 and 2 use a Rayleigh pdf of signal amplitude to represent the target return. This model approximates the fluctuation associated with a target composed of a large number of randomly distributed scatterers, with no single scatterer being dominant. The amplitude pdf has the same form as that of the random noise envelope. It is generally considered to be reasonably representative of aircraft return magnitude fluctuations [4].

For Swerling cases 3 and 4, the pdf is the χ^2 pdf with four degrees of freedom:

$$f(x) = \left(\frac{4x}{\bar{x}^2}\right) e^{(-2x/\bar{x})} U(x) \quad , \quad (4-13)$$

where the random variable x corresponds to the target RCS, and $\bar{x}$ is the mean RCS. This pdf is representative of a complex target comprising one dominant constant reflector and multiple smaller reflectors. The four cases are summarized as follows:

- a. *Case 1:* The echo pulses received from a target on any one scan are of constant amplitude throughout the entire scan but are uncorrelated from scan to scan (slow fluctuation). A Rayleigh pdf describes the target signal amplitude.
- b. *Case 2:* The target signal amplitude is also Rayleigh-distributed, but the fluctuations are independent from pulse to pulse instead of scan to scan (fast fluctuation).

c. *Case 3*: The target fluctuation is assumed to be independent from scan to scan (slow fluctuation), and the pdf is given by Equation 4-13.

d. *Case 4*: The pdf for the target signal amplitude is also given by Equation 4-13, but the fluctuations are independent from pulse to pulse instead of scan to scan (fast fluctuation).

The target fluctuation corresponds to broadening the pdf of the signal-plus-noise receiver output. Thus, the SNR must be increased over the nonfluctuating case to achieve the specified P_d . This increment in SNR is termed the *fluctuation loss*. For $P_d > 0.3$, there is the greatest fluctuation loss for type 1, and the smallest loss for type 4 [3]. All four cases result in poorer performance than for case 0 for $P_d > 0.3$. A series of pulses will often be partially correlated, however, so that the performance lies somewhere between cases 1 and 4. In that sense, the cases 1 and 4 provide limiting performance bounds.

Despite this, in the preface to [21] (almost 20 years after his initial publication), Swerling notes that the four models do not bracket the behaviour of all physically interesting targets. Indeed, if the targets of interest are precisely known and their RCS measured, more accurate models can be derived. For the CP-140 application, however, there may be a wide variation in the types of the targets of interest. Use of the case 1 model provides a conservative performance estimate. In general, radar systems operating at a fixed frequency are assessed using the Swerling 1 fluctuation model [4].

4.1.6 Pulse Compression

As mentioned in the previous chapter, the pulse compression scheme that is used in the MPRF RWS mode is a 13-bit Barker code applied to a 13 μs transmitted pulse. The Barker 13-bit phase code is most often found in fielded systems, including the APG-63 and APG-70 on the F-15, as well as the APG-66 and APG-68 on the F-16 [4]. The 13 μs pulse is considered to consist of 1 μs *chips*, and each chip is modulated to be either in phase with the carrier, or 180° out of phase with the carrier. After pulse compression, the range resolution of a 1 μs pulse is achieved while maintaining the higher average power of the 13 μs pulse.

In computing detection range for a pulse-compressed target return competing with thermal noise, there are two methods that give a close approximation to the correct answer for the matched-filter receiver [22]. With the first method, the noise bandwidth term B in the denominator of the radar equation can be made equivalent to the matched-filter bandwidth of the uncoded pulse. With the second method, B is set to the matched-filter bandwidth of the chip width, and then the number of chips, N_c , is placed in the numerator to account for the coherent integration of the chips. For the MPRF RWS example, then, B could be set to the matched-filter bandwidth of the 1 μs chip, and a factor of 13 could be added to the numerator. Note that this is equivalent to transmitting a 1 μs pulse at 13 times the peak power.

The noise bandwidth term has not been used explicitly in the equations given in this thesis. Recall from Chapter 3 that if a matched transmit/receive system is used, the noise bandwidth B can be set equal to $1/\tau$ where τ is the pulse width. To include the performance of the phase coded waveform in MPRF RWS mode, then, d_r in Equation 3-1 is calculated using the 13 μs pulse width rather than the 1 μs chip width.

There is little discussion in the literature about the performance of a pulse-compressed target return competing with clutter. The basic premise underlying the discussion of the performance in noise was that there is a gain in the SNR of the target return during pulse compression due to the correlation of the target return across (compressed-width) range gates. Because the noise samples in adjacent range gates are assumed to be uncorrelated, the overall gain in SNR (over a chip-width pulse) is N_c . Clutter returns, on the other hand, may show some correlation from one range gate to the next. Indeed, echoes from discrete clutter scatterers illuminated by the sidelobes benefit from the same type of pulse compression as a target. Because the characteristics of the clutter are dependent on so many factors in MPRF modes (i.e. the flight scenario, the terrain, the PRF, the antenna sidelobes, etc.), the gain in SCR due to pulse compression is not readily calculated. As suggested earlier, the best way to predict the performance of the radar in MPRF modes is to simulate the clutter returns. This is addressed in Chapter 5.

There is one additional factor that can degrade the performance of the pulse compression technique from the theoretical performance shown above; the performance of a phase-coding technique is dependent on the Doppler shift across the received pulse. Recall that the FFT algorithm is optimal when there is a constant phase shift from pulse to pulse (i.e. from a target with a constant radial velocity). The Barker code compression scheme, on the other hand, is optimal when there is no phase shift over the length of the pulse (i.e. from a target with zero radial velocity). The Barker code compression scheme depends on the relative phase shift of 0° or 180° on each chip for compression. When echoes are backscattered from a target with a nonzero radial velocity, the target return contains a Doppler phase shift from chip to chip. This

phase shift alters this binary code such that the peak response is less than 13 times that of the amplitude of the coded pulse. Suppose, for instance, a target had a closing velocity of 180 m/s, resulting in a Doppler shift of 11.65 kHz. A Doppler shift of 11.65 kHz produces a phase shift of $(11,650 \times 360)$ degrees per second, or 4.2° per microsecond. For the 13 μ s pulse, then, there would be a total phase shift of 55° over the length of the pulse. If the received pulse were processed with no derotation, there would be a relative peak amplitude of 11.2 rather than 13. If each chip could be derotated by 4.2° , however, then optimal pulse compression could be carried out.

Because of the code's sensitivity to a Doppler shift, the compression is often preceded by some attempt to compensate for the Doppler shift [4]. An AI radar might be tuned, for instance, to compensate at a derotation rate that minimizes the attenuation of high-priority targets. A compensating velocity, V_{comp} , is scaled to the radar platform velocity. Byrum gives an example where $V_{comp} = 1.25(V_o)$ [4]. The derotation rate, $d\phi / dr$, in degrees/range cell is then

$$\frac{d\phi}{dr} = 720 \frac{V_{comp} \tau}{\lambda} , \quad (4-14)$$

where τ is equal to the compressed pulse width.

This type of compensation is worthwhile for a forward-scanning AI radar whose main concern is incoming threats. For a surveillance platform, however, both closing and receding targets are often of interest. The surveillance platform has an advantage over an AI aircraft, though, based on the fact that the slower platform motion and the types of targets of interest result in smaller expected relative closing velocities. For these reasons, no compensation is recommended for use with the CP-140. It should be

made possible, however, to allow derotation tuned to a particular target of interest in a track-while-scan implementation.

4.2 CFAR

In order to determine if a target is present, some estimate must be made of the clutter-plus-noise (interference) power in a range-Doppler cell. If the characteristics of the interference power were known exactly, a fixed theoretical threshold could be set for each range-Doppler cell, and optimum detection (i.e. maximum detection probability) could be achieved for a desired probability of false alarm. If the average interference power (variance), μ , in each of the I and Q channels were known *a priori*, the false alarm probability would be:

$$P_{fa} = P[Y > Y_0 | H_0] = e^{(-Y_0/2\mu)} \quad (4-15)$$

where Y_0 denotes the fixed optimum threshold, and Y is the content of the cell under investigation [23]. Since the characteristics of the interference power are not known at any given range and Doppler bin, a fixed theoretical threshold detection scheme cannot be used. The interference power must be estimated based on samples collected within a reference window of the range-Doppler spectrum.

In this case, constant false alarm rate (CFAR) processors are normally used. CFAR processors set the threshold adaptively based on local information. The threshold in a CFAR processor is set on a range-Doppler cell-by-cell basis by processing a group of reference cells surrounding the cell under investigation. A detection is declared if the power in the cell under test exceeds a certain threshold. This threshold is proportional to the estimated interference power. This threshold

value is set just high enough above the mean background level to reduce the probability of false alarms to an acceptably low value.

A block diagram of a typical one-dimensional CFAR processor is shown in Figure 4-2. A one-dimensional CFAR processor would normally be used with a radar that measures only range. The processor often extends to two dimensions for a radar that measures range and Doppler.

Figure 4-2. Typical CFAR Processor

The statistic Z is proportional to the estimate of total noise power, and is formed by processing the contents of N reference cells surrounding the test cell. The "guard" cells immediately adjacent to the test cell are not used in the computation of Z in order to prevent target self-masking. If the content of the test cell is Y , then a target is declared to be present if Y exceeds the threshold TZ . T is a constant scale factor used to achieve a desired constant false alarm probability for a given window of size N when the total background is homogeneous. Typical values for N might be anywhere from 4 to 200. Typical values for T are discussed later in this section.

There is an inherent loss of detection probability in a CFAR processor compared with the optimum processor. This is because the CFAR processor sets the threshold by estimating the total interference power within a *finite* reference window. The optimum processor, on the other hand, sets a fixed threshold under the assumption that the total noise power is known. The situation is further exacerbated for a CFAR processor if there are multiple targets in neighbouring cells. This loss is sometimes termed *CFAR loss*.

4.2.1 CFAR Techniques

Most modern radars use a variant of the well-known *cell averaging* (CA) CFAR [5]. The CA-CFAR processor adaptively sets the threshold by estimating the mean level in a window of N range and/or Doppler cells. The CA-CFAR maximizes the probability of detection in a homogeneous background when the reference cells contain *independent and identically distributed* (IID) samples governed by an exponential distribution (i.e., for quadratic detection) [21]. Based on this uniform statistic, cell-averaging techniques derive the interference power estimate using the unbiased and most efficient arithmetic mean estimator. As the size of the reference window increases, the detection probability approaches that of the optimum detector which is based on a fixed threshold.

The CA-CFAR processor performs well for noise-limited detection. Thus, simple CA-CFAR techniques are used successfully for HPRF RWS and VS modes. Unfortunately, in MPRF modes, the data seldom conform to the criteria given above. In general, there are two common situations that occur which cause a degradation in

the performance of the CA-CFAR technique: 1) clutter edges, and 2) multiple interfering targets.

Consider, for example, a ground based radar with a rotating antenna. A *clutter edge* might occur when the rotating antenna illuminates a mountain range or an urban area. Figure 4-3 shows an example of clutter edges in range.

Figure 4-3. Clutter Edges

In this example, the extent of the large-RCS backscatterer is shorter in range than the total unambiguous range measured by the radar. If the reference window contains both noise-limited and clutter-limited values but the test cell is in the noise-limited region (i.e., range gate 60), a target return may be masked due to an increase in the threshold. If the cell under test is a clutter cell (i.e. range gate 40), however, the threshold might not be high enough to achieve the design false alarm rate because the interference

estimate includes values from the noise-limited region. *Interfering targets* occur when there is a primary target in the test cell and one or more targets that lie within the CFAR reference window. Modifications to the CA-CFAR scheme have been studied to overcome the problems associated with nonhomogeneous interference backgrounds and interfering targets.

The *greatest-of* (GO) CFAR detector splits the reference window into leading and lagging parts symmetrically about the cell under test. The leading and lagging windows are then summed, and the greater of the two is chosen. The GO-CFAR processor can help to regulate the false alarm rate in a region of clutter transition. Hansen showed a false alarm probability almost three orders of magnitude lower for GO-CFAR (as compared to CA-CFAR) for a discontinuity of 20 dB in the clutter level [24]. The additional losses incurred when using the GO-CFAR technique in homogeneous backgrounds are typically small. One study showed a typical loss of SNR of between 0.1 and 0.3 dB for the GO-CFAR as compared to the CA-CFAR for a homogeneous background [25]. When there is a single interfering target with strength equal to that of the primary target, however, the detection probability decreases intolerably with the GO-CFAR [26].

In order to prevent the suppression of closely spaced targets, a *smallest-of* (SO) CFAR scheme was proposed [26]. This algorithm takes the smallest of the leading and lagging windows to estimate the interference power. The SO-CFAR is able to prevent the suppression of closely spaced targets, but its performance begins to degrade if there are interfering targets in both the leading and lagging windows. Also, the SO-CFAR algorithm fails to maintain a constant false alarm rate at clutter edges. Gandhi and

Kassam have shown that a 15 dB clutter edge amounts to an increase of more than five orders of magnitude in the false alarm rate at $N=24$ and a design $P_{fa} = 10^{-6}$ [23].

More recently, an *ordered statistic* (OS) CFAR scheme has been introduced, which alleviates these problems to a certain extent [27]. OS schemes have been successfully used in applications such as image restoration. The OS-CFAR processor estimates the interference power by simply selecting the k th smallest cell in the reference window of N cells. For $k=N/2$, this is a median threshold. This procedure has the effect of allowing up to $N-k$ interfering targets to be present without serious degradation in detectability. Alternatively, this procedure allows the reference window to encompass $N-k$ cells on the high side of a clutter edge without reducing the sensitivity for a test cell that lies in a noise-limited region. Typical values for an OS-CFAR scheme are $k=27$ for $N=32$.

A more generalized form of the OS-CFAR scheme is known as the *trimmed mean* (TM) CFAR scheme, which is similar to the *censored mean-level detector* (CMLD) [28, 29, 30]. The TM-CFAR processor first orders the range cells according to their magnitude and then trims T_1 cells from the lower end and T_2 cells from the upper end. The threshold is formed by summing the remaining range cells and multiplying by a threshold multiplier. The TM-CFAR scheme reduces to the CA-CFAR and OS-CFAR schemes for specific trimming values.

In general, the performance of CA-CFAR can degrade rapidly in non-ideal conditions caused by extraneous targets and nonuniform clutter. CFAR techniques based on ordered statistics such as OS and TM trade a small loss in detection performance in homogeneous environments for much less performance degradation in

nonhomogeneous backgrounds [27]. The trade-off in terms of implementation is an increase in complexity.

The performance of OS-CFAR, however, depends heavily on the choice of k and the scenario in which the radar finds itself [33]. TM-CFAR, on the other hand, can tolerate interfering targets and has a more graceful degradation in the event that the number of interfering targets exceeds the number of censored samples because it averages the rest of the samples instead of using only a single ordered sample [31]. TM-CFAR also allows a more graceful degradation around clutter edges when the OS value of k is not well suited to the particular window.

An airborne pulse Doppler radar faces additional clutter edges as a result of its motion and altitude. These include clutter edges from the altitude line, from the mainlobe/sidelobe transition, and from where thermal noise-limited and clutter-limited regions meet. As will be demonstrated, the CFAR technique for the CP-140 is designed to deal with these clutter edges using *a priori* information. There will still exist, however, clutter edges that will not be known *a priori*, such as illumination of a coastline where both land clutter and sea clutter are backscattered. Indeed, a maritime surveillance platform might often face such a scenario. A CFAR technique will not have any information about the temporal and spatial location of this type of clutter edge. As such, the use of an OS technique (where the threshold is based on a single sample) might risk unpredictable performance. Choosing k near the maximum in an OS-CFAR processor in this situation would avoid an excessive false alarm rate, but would suffer a greater loss in detection performance [23]. The use of some type of cell-averaging technique provides more robust performance overall.

For these reasons, TM-CFAR is selected here to be investigated for possible use on the CP-140. TM-CFAR provides the means to handle interfering targets, while still providing robust performance in the presence of a clutter edge. CA-CFAR is included in the analysis to design the initial technique and as a baseline against which to compare different trimming values.

4.2.2 CA and TM-CFAR

The following description of the CA and TM-CFAR processors assumes that the clutter-plus-noise returns in different cells are independent and governed by an exponential distribution (i.e. from a quadratic detector). The false alarm rate of the two processors is independent of the power of the exponentially distributed interference. Note that it is impossible to detect $N+1$ contiguous and independent radar targets of identical strength. It is therefore assumed in the derivation of the performance of the processor that the number of targets is small relative to the number of cells in the CFAR window.

In the CA-CFAR processor, the total noise power, μ , is estimated by the sum of N range cells of the reference window. The statistic Z is formed as [23]:

$$Z = \sum_{i=1}^N X_i \quad . \quad (4-16)$$

Recall from Figure 4-2 that the content of the test cell is Y . A target is declared to be present if Y exceeds the threshold TZ . T is a constant scale factor used to achieve a desired constant false alarm probability for a given window of size N when the total background is homogeneous. Note that some CFAR processors are designed to use an

average interference power (i.e. divide the sum Z by N before calculating a threshold multiplier). Following Ghandi and Kassam, it is the threshold multiplier T (and not Z) that is a function of N [23].

The probability of detection for the CA-CFAR processor given an exponential distribution of returns in the test cells can be shown to be:

$$P_d = \left[1 + \frac{T}{(1+S)} \right]^{-N}, \quad (4-17)$$

where S is the average SNR of the target [23]. The probability of false alarm can be obtained from Equation 4-17 by setting $S=0$. Consequently, the scale factor T is computed from Equation 4-17 for a desired P_{fa} as:

$$T = (P_{fa})^{-1/N} - 1. \quad (4-18)$$

Figures 4-4 and 4-5 show two possible reference windows around a test cell. In Figure 4-4, a 5x5 cell region is used. The range-Doppler cells directly adjacent to the cell under test are excluded to prevent target self-masking from occurring, since a radar target return may be spread over several cells. This results in a total of 16 cells used in the CFAR estimate. For the 7x7 window, a total of 40 cells is used.

where $X_{(1)}, \dots, X_{(N)}$ are the ordered samples. Given the same assumptions of a homogeneous exponentially distributed background, the false alarm probability is found to be [23]:

$$P_{fa} = \prod_{i=1}^{N-T_1-T_2} M_{V_i}(T) \quad (4-20)$$

where

$$M_{V_1}(T) = \frac{N!}{T_1!(N-T_1-1)!(N-T_1-T_2)} \sum_{j=0}^{T_1} \frac{\binom{T_1}{j} (-1)^{T_1-j}}{\frac{N-j}{(N-T_1-T_2)} + T} \quad , \quad (4-21)$$

$$M_{V_i}(T) = \frac{a_i}{a_i + T}, \quad i = 2, \dots, N-T_1-T_2 \quad , \quad (4-22)$$

and

$$a_i = \frac{(N-T_1-i+1)}{(N-T_1-T_2-i+1)} \quad . \quad (4-23)$$

The probability of detection is obtained for TM-CFAR by replacing T with $T/(1+S)$ in Equation 4-20, resulting in

$$P_d = \prod_{i=1}^{N-T_1-T_2} M_{V_i}\left(\frac{T}{1+S}\right) \quad , \quad (4-24)$$

where T changes to $T/(1+S)$ in Equations 4-21 and 4-22 accordingly. When there is no trimming, the value for T corresponds to that of the CA-CFAR processor. If the

same 5x5 reference window in Figure 4-4 were used with $T_1 = T_2 = 2$, the scale factor would be $T=2.38$.

4.3 CFAR FOR THE CP-140

4.3.1 *A Priori* Knowledge

CFAR techniques have been closely investigated over the last thirty years. As mentioned earlier, because the performance of the traditional CA-CFAR degrades quickly in the presence of a clutter edge or multiple targets, most of the modified CFAR techniques were proposed to deal with one or both of these difficulties. Little effort has been made, however, to make use of *a priori* knowledge inherent in an airborne pulse Doppler radar situation. Most of the techniques found in the open literature were developed for the conditions encountered with ground-based or ship-based radars.

There are several characteristics of an airborne pulse Doppler spectrum that are known *a priori*. Some of these characteristics are already exploited by pulse Doppler CFAR processors. For instance, airborne pulse Doppler systems are normally designed to filter out the full clutter return in HPRF mode with spectral notch filtering. In a system where an analog filter is used, the receiver downconversion can be adjusted to heterodyne the clutter component into the notch filter. In modern systems, the frequency spectrum can be shifted to center the clutter at zero frequency, and digital high-pass filtering can be used. Thus, in HPRF mode, no attempt is made to look for targets in the Doppler bins occupied by sidelobe clutter, which represents a frequency range of $\pm 2V_o/\lambda$.

Pulse Doppler radars, however, could make use of several other characteristics of the pulse Doppler spectrum. For example, in a properly designed LPRF mode, there is no foldover in range. Thus, there is no clutter return from ranges nearer than the height of the aircraft. Second, there are likely to be strong clutter returns in the range gates corresponding to the altitude of the ownship in low and medium PRF. This *altitude line* is due to the large clutter backscatter associated with the short ranges and near 90° grazing angle. Third, the antenna pointing angle can provide additional information about the *mainlobe clutter* (MLC) return for an MPRF mode. Fourth, for the CP-140, it is known that there are strong sidelobe clutter returns due to the large antenna sidelobes.

Weber et al. began to make use of this *a priori* knowledge with the suggestion of nadir-blanked CFAR [18]. Because the range gates at or near the altitude line are likely to contain strong clutter returns near zero Doppler, Weber et al. suggest blanking these range gates entirely. For the MPRF case studied, nadir-blanked CFAR produced a 13 dB reduction in CFAR loss over the CA-CFAR case, and an 11 dB reduction in CFAR loss over the GO-CFAR case.

Vrckovnik and Faubert suggest notching out the MLC in MPRF modes to avoid including the MLC edge in any CFAR technique [32]. Their Doppler bin blanking technique attempts to determine the MLC edge adaptively by looking for a sharp rise in the magnitude of the clutter. Given knowledge of the ownship velocity and the antenna pointing angle in MPRF mode, however, a CFAR algorithm can use Equation 3-2 to calculate the expected location of the MLC edge.

4.3.2 Clutter Simulation

As discussed in Subsection 4.1.4, the total clutter return in MPRF mode for any range-Doppler cell is normally the combination of returns from many different ground patches. Each of these is dependent on all of the factors listed in Equation 4-11. Thus, it is extremely difficult to develop CFAR algorithms deterministically. Weber et al. state that to design a CFAR processor, one should simulate the clutter return with the waveform parameters of the radar system [18].

Clutter simulations for the CP-140 air-to-air modes were done with the simulator SAFIRE [33]. Complete descriptions of the simulator are not repeated here. Descriptions of the modifications required to generate simulated radar data for the CP-140 platform, however, are included here.

An antenna pattern representative of the AN/APS-506 antenna was needed in order to simulate clutter returns that would accurately represent the clutter seen from the CP-140. SAFIRE requires a 2-D matrix representing the amplitudes of the antenna pattern in elevation and azimuth over a specified angular region. Unfortunately, a complete antenna pattern for the AN/APS-506 is not available. A representative pattern was generated using the azimuth and elevation patterns shown in Figures 2-4 and 2-5. A linear interpolation was carried out between the amplitudes of the two, rotated about the mainbeam. The resulting pattern is shown in Figure 4-6.

Figure 4-6. AN/APS-506 Pattern

Most of the parameters required to run a simulation with SAFIRE are held in two files: *initparm.inc* and *sim_init.dat*. The file *initparm.inc* contains radar system parameters such as burst lengths, filter poles, PRIs, pulse widths, number of range gates, and number of fill pulses. These files were altered to reflect the design for the MPRF RWS, HPRF RWS, and VS modes for use on the CP-140. The file *sim_init.dat* contains parameters to specify the engagement scenario such as ownship altitude and speed. Also in this file is a switch to turn the *variable frequency oscillator* (VFO) on or off. When turned on, the VFO translates the MLC return to zero frequency. The parameters in this file were modified depending on the particular scenario under investigation.

Before simulating a sample MPRF RWS range-Doppler map with SAFIRE, it is instructive to consider what one might expect for a particular scenario. Take, for instance, a baseline search scenario where the ownship is traveling at an altitude of 3,000 m at a velocity of 180 m/s. The antenna is pointed at minus 15° in elevation and 0° in azimuth, and the radar is operating at 9.7 GHz. Data are captured for the second burst in the set of eight evenly-spaced MPRFs, corresponding to a PRF of 10,647 Hz.

In this case, the *sidelobe clutter/backlobe clutter* (SLC/BLC) spreads $\pm 11,650$ Hz due to the ownship motion. Since one I and one Q sample are retained for each range gate in a PRI and there are 128 pulses in a burst, the represented Doppler spectrum is 10,647 Hz wide and consists of 128 samples (Doppler bins). Thus, we would expect the SLC/BLC to extend right across the Doppler spectrum. Using Equation 3-2, the centre of the MLC has a Doppler shift of 11,254 Hz. When the VFO is turned on, the MLC is translated to zero Doppler. The peak in the altitude line is translated by minus 11,254 Hz as well (along with the rest of the Doppler spectrum). If the range-Doppler map is displayed from zero Hz to 10,647 Hz, a clutter return at minus 11,254 Hz will be aliased back into the spectrum, and will appear at 10,040 Hz.

Recall that MPRF RWS uses a pulse width of 13 μ s, and that a 200 ns delay was used at each end of each pulse for transmitter switching. With a 10,647 Hz PRF, each PRI is 93.9 μ s. If the PRI starts at the rising edge of the transmitted pulse, the interpulse period (during which data can be collected) begins at 13.2 μ s and extends out to 93.7 μ s. Since the compressed range gate width is 1 μ s, there is room for a total of 80 range gates in the interpulse period. This is shown in Figure 4-7. As in Doppler, backscatter from ground clutter at longer ranges is aliased back into the range-Doppler map. Using Equation 2-1, the altitude line should appear with a time delay of 20 μ s.

This corresponds to approximately range gate seven. The MLC should appear with a time delay of about $77\text{ }\mu\text{s}$. This corresponds to approximately range gate 64.

Figure 4-7. MPRF Range Gates

Figure 4-8 shows the range-Doppler map output from SAFIRE for the scenario described above. This range-Doppler map is output after the FFT and pulse compression processing. In this map, white represents the strongest return, and black represents the weakest. For all scenarios the VFO was turned on. The Doppler axis extends from zero Doppler to 10,647 Hz (represented by 128 Doppler bins). The range axis extends from 1,980 m to R_{max} for this PRF, 14,088 m (represented by 80 range gates). Note that it would have been equally valid to display the Doppler spectrum as extending from minus $\text{PRF}/2$ to $\text{PRF}/2$, placing the MLC in the center of the Doppler spectrum.

Figure 4-8. Clutter Map with AN/APS-506 Pattern

Several features are evident in the clutter map. The first is the altitude line, which appears at range gate seven. The altitude line is strongest at zero Doppler, which appears, as expected, near Doppler bin 121 (10,040 Hz). The strength of the altitude line clutter decreases moving away from zero Doppler. The MLC is strongest at range gate 64 (11,591 m), and extends across at least twenty range gates. The elevation sidelobes between the mainlobe and nadir create a "clutter hook" that decreases in range and Doppler from the mainlobe. The clutter hook becomes asymptotic with the altitude line as the elevation decreases to nadir.

The ground clutter illuminated by antenna backlobes has a negative Doppler shift, and decreases in Doppler frequency from the peak of the altitude return. Just as

the sidelobes between the mainlobe and nadir created a clutter hook, the antenna backlobes extending out from nadir directly behind the aircraft create clutter that increases in range as it becomes more negative in Doppler frequency. The magnitude of the backlobe clutter hook is less than that associated with the sidelobes due to the antenna pattern, but appears above the noise floor in the range Doppler map nonetheless. The region between the MLC peaks and away from the immediate altitude line is composed of sidelobe clutter, but with a lower amplitude than that near the altitude line.

The same scenario was run with a simple pencilbeam pattern to demonstrate the relative SLC levels. The pencilbeam pattern shown in Figure 4-9 is normally used with SAFIRE to represent a typical AI aircraft antenna pattern. Outside the $\pm 20^\circ$ region depicted here, the pencilbeam pattern has sidelobes 47 dB below the mainlobe, and backlobes 60 dB below the mainlobe. A cross section of this pattern is shown in Figure 4-10.

Figure 4-9. Pencilbeam Pattern

Figure 4-10. Cross Section of Pencilbeam Pattern

The range-Doppler (clutter-plus-noise) maps for the pencilbeam antenna pattern and the AN/APS-506 pattern are shown below in Figure 4-11. The same scale factors were used for each. Clearly, the SLC is lower for the pencilbeam pattern than for the AN/APS-506 pattern. The clutter hook still exists, but at a much lower level due to the lower sidelobes and backlobes.

Figure 4-11. Clutter Map Comparison with a) Pencilbeam and b) AN/APS-506 Patterns

The clutter and noise were extracted from SAFIRE for these two cases just before the point at which they are summed (and before any scaling due to the automatic gain control). In Figure 4-12, the clutter and noise are plotted for the two cases for range gate 47. Note that in range gate 47, most of the sidelobe clutter has fallen below the noise floor for the pencilbeam pattern. This is true for many of the outer range gates. With the AN/APS-506 pattern, however, the sidelobe clutter still lies above the noise floor. This is true for most scenarios with the AN/APS-506 pattern. Thus, the CP-140 MPRF RWS mode will normally carry out clutter-limited detection.

Figure 4-12. Clutter and Noise in Range Gate 47 for the Two Antennas

Variations to the range-Doppler clutter map occur depending on the scenario. The location of the altitude line varies depending on the aircraft altitude. The location and width of the mainlobe clutter depend on the aircraft altitude and elevation angle. The strength of the clutter varies with altitude as well, decreasing at higher altitudes, and increasing at lower altitudes. The extent of the sidelobe clutter depends on the ownship velocity. There will most certainly be scenarios in which the ownship is traveling slowly enough that the sidelobe clutter does not cover the full Doppler spectrum. The Doppler spectrum reverses in frequency for an azimuth angle of 180° .

4.3.3 Regional CA-CFAR

Clearly, there are regions of the range-Doppler clutter map that do not exhibit a simple (noise-like) exponential distribution. Given that the MLC region is blanked, the regions most affected are the altitude line region and the clutter hook regions. While the regional characteristics are present with the low-sidelobe pencilbeam antenna, they are accentuated by the higher sidelobes of the AN/APS-506 antenna. These regions exhibit characteristics for which a traditional CFAR window is not optimum. Because of this, a CFAR technique that splits the range-Doppler space into several regions has been developed for the CP-140. These regions are shown in Figure 4-13. The exact locations of these regions in the range-Doppler map depend on the particular scenario.

Figure 4-13. Regional CFAR

These regions can be characterized as follows:

- a. *Region 1: Weak SLC/Noise.* In this region, the spectrum is composed of thermal noise and/or weak sidelobe clutter. The lower the aircraft flies to the ground, the stronger the SLC contribution will be. In general, the weak SLC is approximately exponentially distributed. Variations from an exponential distribution are due in part to non-uniformity of the antenna sidelobes.
- b. *Region 2: Strong SLC.* In this region, the strong sidelobes about the mainlobe of the antenna pattern result in SLC that has non-exponential characteristics. The range-Doppler cells are more alike in amplitude across range gates than they are across Doppler bins.
- c. *Region 3: Altitude Line.* In this region, there is a significant altitude line when the ownship is flying at lower altitudes. There is also a large contribution from antenna sidelobes pointing toward near-nadir. The range-Doppler cells are more alike in amplitude across Doppler bins than across range gates, as the clutter comes from a small width of ranges.

Each of these regions is characterized by different types of clutter, so it is logical to tailor CFAR techniques for each of them.

In the strong SLC region, the SLC tends to be strongest closest to the MLC, and falls off in strength toward the weak SLC/noise-limited region. In this sense, the extent of the strong SLC region in range is relayed to the extent of the MLC in range. The extent of the MLC across the range gates is a function of the antenna elevation angle and the ownship altitude. The larger the downlook angle and the lower the

altitude, the smaller the range gate spread will be. In straight and level flight, the maximum downlook angle is 15 degrees (an elevation angle of minus 15°). With a 3-dB beamwidth of 4.5° in elevation, the mainlobe illuminates the downlook angles between 12.75° and 17.25°. Table 4-3 shows the total spread in range assuming 1 μ s range gates for a 10° and 15° downlook angle.

Table 4-3. Range Gate Spread

Ownship Altitude	# Gates Spread for a 10° downlook angle	# Gates Spread for a 15° downlook angle
300 m	5	3
1,000 m	18	8
3,000 m	54	23
5,000 m	90	39
10,000 m	180	77

With an average MPRF of 14 kHz, there is an average of 58 range gates. Thus, for all but the largest downlook angles at the lowest altitudes, the MLC will spread across most if not all of the range gates. Because of this, it is reasonable to assume that the strong SLC region will extend across the entire range spectrum. The fact that the range-Doppler cells are more alike across range gates suggests that a CFAR window that is longer in range than in Doppler should be used for Region 2. The fact that the range-Doppler cells are more alike across Doppler bins in the altitude line region (Region 3) suggests a CFAR window that is longer in Doppler than in range.

The questions that must be answered are as follows: 1) how large should each region be, 2) what type of reference window should be used in each, and 3) how much better does regional CFAR perform than a standard CFAR technique? Note that the location and width of each region must be determined either *a priori* or based on *inertial navigation system* (INS) and radar altimeter data that would be available to the radar signal processor during flight. To begin to answer these questions, the 3,000 m baseline scenario was tested with many combinations and variations to the regions shown above.

First, the width of the MLC notch had to be determined. In practice, more than a single Doppler bin separation is required between a possible target and the MLC. This is because instabilities and jitter in the radar transmit-receive chain broaden the extent of the MLC and other strong clutter returns. Also, Doppler filters are normally designed with spectral sidelobe weighting, resulting in a resolution degradation. It is desirable, however, to have the MLC notch as narrow in frequency as possible to allow maximum room in the CFAR map for possible target detections. As a baseline, the MLC notch filter was designed to be the width of the MLC (as determined by the antenna pattern and look angle) rounded up to the nearest integer, plus one Doppler bin on each side. For the scenario described earlier, this results in a notch of three bins on each side of the map. As a result, 480 of the total 10,240 cells in this map are excluded as test cells. They are, however, kept for use in the reference window of other test cells. This leaves a total of 9,760 cells for which CFAR is carried out in this baseline scenario.

When the test cell lies close to the edge of the range-Doppler map, there may not be enough cells to form the desired reference window. In these cases, the cells that

are available for the reference window are averaged, and the threshold is calculated according to Equation 4-18 (where the value of N used corresponds to the number of available cells). Note that this results in a slightly lower P_d along the edge of the range-Doppler map in order to maintain the same P_{fa} . The other option in this case would have been to allow the reference window to shift relative to the test cell in order to maintain N cells in the average. This is done, for instance, in the CFAR techniques used in [33]. In cases where a known homogeneous background is being averaged, this technique is valid. For the cases described here, however, the cells included in the interference power estimate are chosen explicitly due to their proximity and relative orientation to the test cell. Thus, the window is not permitted to shift in relation to the test cell.

The 5x5 and 7x7 reference windows shown in Figures 4-4 and 4-5 were applied to all test cells using CA-CFAR to determine the performance associated with traditional windows. Threshold multipliers were calculated using Equation 4-18 (i.e., for an exponentially distributed background). Because these multipliers are not actually being used with exponentially distributed data, it should not be expected that the design false alarm rates will be realized. Approaching the task in this manner, however, allows examination of how far the simulated range-Doppler map strays from the ideal. Any time "design P_{fa} " is mentioned in the text, it is for a threshold that has been calculated for a homogeneous exponentially distributed background. Where results are determined empirically using the simulated range-Doppler map, the term "actual P_{fa} " is used. A relatively high design P_{fa} of 10^{-3} was used to help determine where the reference windows were weakest. The results are shown in Figures 4-14 and 4-15. Each mark on the map represents a false alarm.

Figure 4-14. 5x5 Window for All Test Cells

Figure 4-15. 7x7 Window for All Cells

If the range-Doppler clutter map consisted of exponentially distributed clutter, approximately 10 false alarms would be expected. In this case, 38 false alarms were generated for the 5x5 window, and 50 false alarms were generated for the 7x7 window. A closer investigation of Figures 4-14 and 4-15 reveals areas in which the performance of the traditional square windows suffer. Clearly, the altitude region is not well served by the square windows. As well, there is a high false alarm rate in the region occupied by the strong sidelobe clutter.

In order to better deal with the altitude line region (Region 3), a reference window that is long in Doppler and short in range was devised. Recall that using fewer cells in a CFAR window results in a lower P_d for the same design P_{fa} . Therefore, it is desirable to use the same number of cells as used previously to achieve the same theoretical P_d and to avoid any seemingly enhanced false alarm performance due to a higher threshold with fewer cells. Thus, the 11x3 reference window shown in Figure 4-16 was used, for a total of 16 cells in the CFAR average. This is the same

number of cells used in the 5x5 window, so the performance of the two can be readily compared. Figure 4-17 shows the result of applying the 11x3 window to all test cells.

Figure 4-16. 11x3 Window Used in Region 3

Figure 4-17. 11x3 Window for All Test Cells

Using the 11x3 window, a total of 37 false alarms was generated. The performance of this window in the altitude line region is far superior to that shown in Figure 4-14. As expected, the performance of this window in the clutter hook region is poorer due to its bias toward averaging across Doppler bins.

In order to better deal with the strong SLC region (Region 2), a reference window that was long in range and short in Doppler was devised. Initial testing showed that while some advantage was gained in using such a window, large windows (16 cells or more) would not perform significantly better than the traditional square window. Since the performance from the square windows was unacceptable, the 3x5 reference window shown in Figure 4-18 was used, for a total of 6 cells in the CFAR average. The width of the clutter hook in Figure 4-8 suggests that Region 2 should be

at least 10 Doppler bins wide on each side of the map. A width of 10 Doppler bins was used for the initial test. The result is shown in Figure 4-19.

Figure 4-18. 3x5 Window Used in Region 2

Figure 4-19. 3x5 Window Used in Strong SLC Region

Using the 3x5 window in the Strong SLC region resulted in a total of 38 false alarms. Several of the false alarms in the strong SLC were cleared using this window (compare with Figure 4-14), but the window generated even more false alarms in the altitude region. Again, this would be expected in this region due to this window's bias toward averaging across range gates.

The CFAR techniques described above were then applied only to the regions for which they were designed, and the width of each region was adjusted to minimize the total number of false alarms. The 7x7 reference window actually produced fewer false alarms than the 5x5 window when restricted to Region 1, so the 7x7 window was used in this region. The final result is shown in Figure 4-20, with 15 false alarms. The

optimum combination of regions is shown in Figure 4-21 for the reference windows described above (for the 3,000 m case).

It is important to note that these 15 false alarms occupy each of the regions fairly evenly; no one region appears to be performing more poorly than the others with the given reference windows. This suggests that to realize a lower actual P_{fa} , the threshold multipliers will have to be increased slightly for all regions. It was stated earlier that a higher threshold might be needed in Regions 2 and 3. In a sense, this has been accomplished by using fewer cells to calculate T and Z ($N=40$, 6, and 16 for Regions 1, 2, and 3 respectively). Regions 2 and 3 realize a lower P_d than Region 1, but achieve a similar actual P_{fa} . This is advantageous for an operational system, as the thresholds for the entire map can be raised slightly if there are too many false alarms, avoiding the need to investigate each region separately.

Figure 4-20. False Alarms for Optimum Combination of Regions, 3,000 m Test Case

Figure 4-21. Optimum Regions for 3,000 m Test Case

In this simulation, the CFAR algorithm was given only radar altimeter data, antenna pointing angle, mainlobe width, and ownship velocity. The CFAR algorithm calculated the altitude line range gate based on altimeter data, and added another ten range gates to Region 3 (increasing in range). Because the characteristics for the altitude line region extend across the Doppler spectrum, Region 2 extends up to the MLC region. The width of the MLC was calculated based on ownship velocity and antenna look angle. The strong SLC region was then taken as 21 Doppler bins beside the MLC.

The width of the regions shown in Figure 4-21 can actually vary slightly and still maintain the same number of false alarms. Region 3 can extend out in range to a width of 12 range gates without incurring additional false alarms, but may not extend further below the altitude line itself. Each side of Region 2 can vary between 21 and 23 Doppler bins in width without incurring any additional false alarms.

Fifteen false alarms out of a total of 9,760 cells yields an actual P_{fa} of approximately 1.5×10^{-3} for a design P_{fa} of 10^{-3} . This process was repeated for the design probabilities of false alarm listed in Table 4-4. The entire procedure was then repeated for elevation angles of minus 10° and minus 5° for this scenario. The number of false alarms from a single simulation of minus 10° and minus 5° are listed in Table 4-4.

Table 4-4 Number of False Alarms for 3,000 m Scenarios

Design P_{fa}	Number of False Alarms (out of 9,760 test cells)		
	3,000 m, el -15°	3,000 m, el -10°	3,000 m, el -5°
1×10^{-3}	15	21	16
5×10^{-4}	10	13	9
1×10^{-4}	4	9	6
5×10^{-5}	3	7	4
1×10^{-5}	1	0	2

Thus, in this case, it appears that the design P_{fa} must be lower than 5×10^{-4} to achieve the desired number of false alarms (10) for all cases. Certainly, using a design

P_{fa} of 1×10^{-4} achieves better than the desired number of false alarms for all three of these scenarios.

4.3.3.1 Width of Region 2

Because a lower P_d is expected for the reference window designed for Region 2, it is desirable to make Region 2 as narrow as possible. The width of Region 3 was set to 12 range gates, which appeared to be close to optimum for all of the scenarios. The width of Region 2 was varied for four different altitudes to investigate the optimum width of this region. Figure 4-22 shows the total number of false alarms vs. the width of Region 2 in Doppler bins.

Figure 4-22. Number of False Alarms vs. Width of Region 2

There is an initial monotonic decreasing pattern for the 1,000 m, 3,000 m and 5,000 m cases as the width of Region 2 grows from zero Doppler bins. The 3,000 m case is optimum with a width of 21 bins ($\pm 1,747$ Hz), while the 1,000 m case has the

fewest false alarms with 8 Doppler bins (± 665 Hz). There is no one width that optimizes Region 2 for all cases. Indeed, the 6,000 m case does not appear to benefit at all from the special treatment of a strong sidelobe region for this particular simulated data set. A width greater than four Doppler bins actually causes more false alarms for the 6,000 m case. This is due to the fact that at 6,000 m, the aircraft altitude is high enough that the clutter drops off rapidly from the mainlobe peak, resulting in a negligible strong SLC region.

The false alarms caused when the width of Region 2 is assumed to be zero for the 3,000 m and 6,000 m cases are plotted in Figures 4-23 and 4-24, respectively. The upper plot in each case shows the Doppler bins that contain clutter from the sidelobes that are higher in elevation than the mainlobe. The lower plot in each case shows the Doppler bins that contain the sidelobe clutter hook. Note that Figure 4-23 shows the same false alarm pattern shown in Figure 4-15 (7x7 window applied to entire clutter map), but has had most of the false alarms near the altitude line cleared by Region 3. As expected, more false alarms occur near the sidelobe clutter hook at both altitudes. At 6,000 m, however, the assumption of a zero width for Region 2 causes many fewer false alarms than for the 3,000 m case. Indeed, there are no false alarms for the 6,000 m case in Doppler bins 4 to 20 (recall that the first three are blanked due to MLC). When the reference window chosen for Region 2 (3x5 window) is introduced, the five false alarms shown in the lower plot of Figure 4-24 are eliminated. There are, however, additional false alarms caused by the 3x5 reference window in both the upper and lower plots, resulting in the net increase in the total number of false alarms as Region 2 is widened (as seen in Figure 4-22).

Figure 4-23. False Alarms with 7x7 Window for 3,000 m

Figure 4-24. False Alarms with 7x7 Window for 6,000 m

Figure 4-25 demonstrates the benefit of applying the 3x5 reference window only to the half of Region 2 that is occupied by the sidelobe clutter hook. Clearly, restricting the use of the 3x5 reference window to one half of Region 2 is more effective for this 6,000 m scenario.

Figure 4-25. 3x5 Window Applied Only to Sidelobe Clutter Hook

Additional tuning of the width of Region 2 could be attempted with additional data sets, but would provide perhaps little benefit to the operational air-to-air modes. Empirical results from the simulated data indicate, however, that the special treatment of a strong SLC region is valid. When flight data become available for the air-to-air modes on the CP-140, the CFAR technique should be further tuned to provide optimum overall performance. In the meantime, for all 6,000 m scenarios, the 3x5 window is applied to only one half of Region 2.

From Figures 4-22 and 4-25, a width of fifteen Doppler bins was chosen as the final width of Region 2 to try to provide some benefit for all cases. This represents a Doppler frequency width of $\pm 1,248$ Hz.

4.3.3.2 Width of Region 3

Given a width of 15 Doppler bins for Region 2, the number of range gates in Region 3 was varied. Figure 4-26 shows the total number of false alarms vs. the width of Region 3 in range gates. Again, there is no single width that is optimum for the four cases. A width of 12 range gates was chosen to provide some benefit overall.

Note that the altitude range gate itself is eclipsed for the 1,000 m case. The width of 12 range gates reported for this case is the number of gates starting with the altitude gate. In the 1,000 m altitude case, the first five gates in the range-Doppler map are treated as Region 3 by the CFAR algorithm (i.e., the altitude gate calculated by the algorithm is range gate minus 6). In the 500 m altitude case, only the first gate in the range-Doppler map is treated as Region 3 (i.e., the altitude gate calculated by the algorithm is range gate minus 10).

Figure 4-26. Number of False Alarms vs. Width of Region 3

4.3.3.3 Final Results

The final number of false alarms for this regional CFAR technique using widths of 15 Doppler bins and 12 range gates for Regions 2 and 3 respectively is shown in Table 4-5 for four different altitudes at a 15° downlook angle. The final regional map is shown in Figure 4-27. The map of 16 false alarms for the 3,000 m case with a design $P_{fa} = 10^{-3}$ is shown in Figure 4-28.

Table 4-5 Final Number of False Alarms for Varying Altitudes

Design P_{fa}	Number of False Alarms (out of 9,760 cells)			
	1,000 m	3,000 m	5,000 m	6,000 m
1×10^{-3}	12	16	24	23
5×10^{-4}	5	11	13	14
1×10^{-4}	0	4	3	5
5×10^{-5}	0	3	2	5
1×10^{-5}	0	1	1	2

Figure 4-27. Final Regional Map

Figure 4-28. Final False Alarms for 3,000 m Case, Design $P_{fa} = 10^{-3}$

The CFAR threshold (TZ) is plotted in Figure 4-29 for the 3,000 m case to demonstrate the amount by which different design values for P_{fa} raise the threshold. Figure 4-29 shows the thresholds for range gate 22 for each of the five probabilities of false alarm shown in Table 4-5 along with the range gate data. Note that the data exceeds the threshold in Doppler bin 105 for the highest P_{fa} setting, which results in a false alarm.

Figure 4-29. CFAR Thresholds for Range Gate 22

Note that it may be beneficial to alter the width of each region depending on the current flight scenario. If a regional CFAR algorithm is to be used with the CP-140, the width of the regions should eventually be fine-tuned with real flight data. In the meantime, the widths chosen represent optimization for normal cruise speeds for the CP-140 and for a 0° boresight azimuth.

Comparing the 50 false alarms that were generated using a global 7x7 window for the 3,000 m case (Figure 4-15) and the 16 false alarms using regional CFAR (Figure 4-28), the advantage of the regional CFAR technique becomes clear. With the final reference windows shown in Figure 4-30, the probability of detection achieved with a 7x7 window has been maintained for Region 1. The large number of false alarms caused by the 7x7 window in Regions 2 and 3 have been avoided, while retaining some detection sensitivity for these regions.

Figure 4-30. Final Regional CFAR Reference Windows

The results in Table 4-5 help demonstrate the increase in the threshold required to achieve the design false alarm rate. A homogeneous, exponentially distributed background would, on average, have produced 10 false alarms for a design P_{fa} of 10^{-3} . In this simulation, the design P_{fa} had to be between 5×10^{-4} and 1×10^{-4} to realize an actual false alarm probability of 10^{-3} for all altitudes. While the results for the design P_{fa} of 10^{-5} are less meaningful with only 9,760 cells tested, it is still worthwhile to look at the trend they suggest. For an actual P_{fa} of 10^{-4} and 9,760 cells tested, only a single false alarm should be seen. The design P_{fa} had to be slightly lower than 10^{-5} to achieve these results for all cases. The preceding discussion suggests that a difference of approximately an order of magnitude is required in the

design P_{fa} to achieve the actual P_{fa} for the non-homogeneous clutter map. This is investigated further in the next chapter.

4.3.3.4 Alternate Reference Windows

Different types of CFAR windows were actually tested in each region to arrive at the reference windows chosen above. The reference windows shown in Figure 4-30 were those that appeared to achieve the best results. A final confirmation was done based on the regions shown in Figure 4-27, altering the reference window used in each region while leaving the reference windows in the other regions intact. The different patterns used were a 9x9 window in Region 1, and the patterns shown in Figures 4-31 and 4-32 for Regions 2 and 3 (compare these to Figures 4-16 and 4-18).

Doppler							
		X	X	G	G	G	X
		X	X	G	Y	G	X
		X	X	G	G	G	X
Range							

Figure 4-31. Alternate Window for Region 2

Doppler				X			
			X	X	X		
			X	X	X		
			X	X	X		
			G	G	G		
			G	Y	G		
			G	G	G		
			X	X	X		
			X	X	X		
			X	X	X		
Range							

Figure 4-32. Alternate Window for Region 3

Each alternate window was applied to the 3,000 m case separately, and the number of false alarms was recorded. The 3,000 m column in Table 4-5 is attached for comparison.

Table 4-6 False Alarms Using Alternate Windows

Number of False Alarms for 3,000 m Case				
Design P_{fa}	Altered R1	Altered R2	Altered R3	3,000 m from Table 4-5
1×10^{-3}	20	24	19	16
5×10^{-4}	15	14	13	11
1×10^{-4}	4	5	6	4
5×10^{-5}	2	4	4	3
1×10^{-5}	1	2	1	1

While this is by no means an exhaustive search of all possible reference windows, this exercise helps to demonstrate the effectiveness of the windows that have been chosen.

4.3.3.5 45° and 90° Azimuths

The regional CFAR shown in Figure 4-27 was then applied to three other scenarios to aid in determining its overall performance. In the first two, the elevation was kept at an angle of minus 15°, but the azimuth angle was altered to center the fourth PRF at 45° and 90°, respectively. The range-Doppler maps for these two scenarios at 3,000 m are shown in Figures 4-33 and 4-34.

Figure 4-33. 3000 m, el -15, az 45

Figure 4-34. 3000 m, el -15, az 90

As expected, the width of the mainlobe clutter is wider in Doppler (compare with Figure 4-8). More striking, though, are the effects of the antenna sidelobes. In these simulations, the effects of the large azimuthal sidelobes in the AN/APS-506 pattern are more pronounced than those seen in the zero azimuth case. These figures highlight again the fact that the CP-140 MPRF RWS processing algorithms will not be able to assume homogeneous, exponentially distributed clutter.

These range-Doppler maps were tested with the regional CFAR technique using the same regional widths as those shown in Figure 4-27. The 3-dB width calculated by the algorithm for each half of the MLC notch has widened by one bin to a total of four bins on each side for both cases. Thus, there were only 9,600 cells tested with the CFAR technique. Note that while four bins are notched out of each side of the map, the region around the MLC (due to the pattern outside the 3-dB width) extends much further in Doppler.

The results shown in Table 4-7 are perhaps better than expected based on the non-homogeneity of the clutter. The regional CFAR technique handled the sidelobe

clutter well, with the number of false alarms similar to that seen with the zero azimuth case. The false alarms for the 90° azimuth scenario are plotted for a design P_{fa} of 10^{-3} in Figure 4-35.

Table 4-7 45° and 90° Azimuth Cases

Number of False Alarms (out of 9,600 cells)		
Design P_{fa}	45° azimuth	90° azimuth
1×10^{-3}	20	13
5×10^{-4}	13	8
1×10^{-4}	5	1
5×10^{-5}	5	0
1×10^{-5}	2	0

Figure 4-35. False Alarms for 90° Azimuth

Note that there are still false alarms scattered about the map, without a high concentration in the areas where the sidelobe clutter is higher. It is advantageous that the sidelobe clutter appears in ridges rather than as single peaks; this results in the surrounding cells in the reference window containing the same type of clutter, which permits the CFAR technique to compensate more adequately for the clutter.

4.3.3.6 Slow Ownship Speed

In the final test scenario, the ownship speed was slowed to 50 m/s (97 kts). This opens up a noise-limited region in the range-Doppler map. While this is a somewhat unrealistic speed for the aircraft in most operational scenarios, it demonstrates another possibility that the CFAR technique must be able to handle. Certainly, when the eight-PRF cycle reaches the 20 kHz PRF, there will quite possibly be a noise-limited region of the spectrum.

This scenario results in a maximum sidelobe clutter frequency of $\pm 3,236$ Hz. Thus, some of the Doppler spectrum (total width of 10,647 Hz) contains thermal noise only. The range-Doppler map for this scenario is plotted in Figure 4-36. An antenna azimuth angle of 0° , an elevation angle of minus 15° , and an altitude of 3,000 m were used. Note that the figure is plotted slightly brighter than others in this thesis to show the noise-limited range-Doppler cells as well.

Figure 4-36. Range-Doppler Map for $V_o = 50$ m/s

To deal with this slow ownship speed, the noise-limited region of the map was made a separate region (Region 4). The resulting regional map is shown in Figure 4-37. Note that in the thermal noise-limited region, the threshold multiplier can be calculated using the design P_{fa} , as it is this type of background interference for which the theoretical calculation was designed.

Based on the ownship parameters, the width of the MLC notch narrowed to two Doppler bins on each side. Recall that the algorithm is programmed to round up and add another bin; half the width of the MLC was actually less than one Doppler bin at this speed. As a result of the slower ownship speed and the 0° antenna azimuth angle, all sidelobe clutter above f_{mlc} was eliminated as well in the two-bin notch. Thus, the noise-limited region extends right up to the edge of the MLC, and one side of Region 2 is eliminated altogether.

Figure 4-37. Regional CFAR for $V_o = 50$ m/s

Unlike the other regions, overlap of the reference windows was not allowed for the noise-limited region. No benefit would be gained in the noise-limited region by allowing the reference window to include clutter-limited cells, as the noise-limited threshold would be raised by those cells. Likewise, allowing Regions 1, 2, or 3 to use cells from the noise-limited region would introduce an unwanted clutter edge.

The regional CFAR technique was tested on the regions of Figure 4-37, with the 7×7 reference window used in Region 4 (i.e., the same reference window that is used in Region 3). The results of the tests are summarized in Table 4-8. Despite the fact that the widths of Regions 2 and 3 were designed for normal cruise speed, the regional CFAR technique performs well at $V_o = 50$ m/s. This suggests that perhaps the

current width of the regions can be used as is, alleviating the need to adjust them during the flight scenario.

Table 4-8 False Alarms for $V_o = 50$ m/s

Design P_{fa}	Number of false alarms (out of 9920 cells)
1×10^{-3}	20
5×10^{-4}	13
1×10^{-4}	4
5×10^{-5}	2
1×10^{-5}	2

A final note should be made about the extra logic required in a CFAR processor to implement regional CFAR. At first glance, the splitting of the range-Doppler map into regions may appear to add excessive computation. In reality, however, each test cell has to undergo some sort of CFAR anyway; regional CFAR just determines the type of reference window to be used. The location of the regions is a function of the flight scenario, and will normally change relatively slowly. The location of each region need only be monitored and altered outside the real-time processing that must take place. If the width and location of each region are determined beforehand, there is little additional computation required to implement this technique. The use of smaller reference windows in Regions 2 and 3 may actually reduce the total computational overhead. The use of a cell-averaging technique appears to make the performance of the algorithm fairly robust even if the regions are not perfectly matched to the current scenario.

4.3.4 Regional TM-CFAR

With the regional CA-CFAR technique in place, the trimming values of the TM-CFAR were investigated for the baseline scenario (3,000 m altitude, 15° downlook angle). Ghandi and Kassam suggest that in a homogeneous background, the value of T_1 should be kept small to attain good detection performance [23]. The value of T_2 , on the other hand, depends on the maximum number of interfering targets expected in the reference window. Because of the large standard deviation of amplitudes expected for Regions 2 and 3, it is perhaps prudent to leave the CA-CFAR technique intact in these regions. Setting $T_2 = 0$ for the calculation of Z degrades the performance of the CFAR processor when an interfering target is present, but is likely to aid in preventing false alarms under normal operating conditions. In addition, the relatively small number of cells in each window (six for Region 2, sixteen for Region 3) means that any trimming severely affects the potential P_d for each test cell. Thus, TM-CFAR was only tested for Region 1.

A reasonable maximum number of interfering targets (within one 7x7 reference window) for use with the CP-140 is two, leading to a selected value of $T_2 = 2$. Indeed, multiple targets are more likely to be more widely spread in the range-Doppler map, except in formation flying situations. Even in formation flying, however, multiple targets would have to be separated by a significant distance in range to appear in separate range cells after pulse-Doppler processing and range centroiding. Nevertheless, there are possible situations where targets with completely different velocities and trajectories will be aliased back to nearby range-Doppler cells, and will become interfering targets.

With $T_2 = 2$, a value for T_1 must be chosen. The 3,000 m baseline scenario was repeated for $T_1 = 2, 3$, and 4. The same number of false alarms occurred for each of these trimming values. The results are shown in Table 4-9.

Table 4-9 Number of False Alarms for Various Trimming Values

	Number of False Alarms (out of 9,760 test cells)	
Design P_{fa}	$T_1 = 2, 3, 4$	$T_2 = 2$
1×10^{-3}	20	
5×10^{-4}	10	
1×10^{-4}	4	
5×10^{-5}	3	
1×10^{-5}	1	

Table 4-9 shows a slightly higher number of false alarms than for the baseline scenario, but the results are still similar to those seen with the regional CA-CFAR technique. As there is no difference in the results for the T_1 trimming values of 2, 3, and 4 for this particular case, a value of $T_1 = 2$ is used in the performance analysis in the next chapter. The trimming value of $T_2 = 2$ allows for two interfering targets with no performance degradation.

Once real flight data are available, the regional CFAR technique should be fine-tuned with larger data sets. The preceding analysis demonstrates the advantages that can be gained using regional CFAR, and provides empirical validation of the technique.

CHAPTER 5: PERFORMANCE PREDICTION

Evaluation of airborne pulse Doppler radar performance must consider the interrelated effects of target fluctuation, clutter characteristics, radar parameters, and target-radar encounter geometry. The next section outlines the system parameters to be used in all evaluations, based on the design of the air-to-air modes. Losses are also tabulated for each of the modes. Section 5.2 provides estimates of the CP-140 air-to-air mode performance based on theoretical calculations. Section 5.3 describes simulation techniques that were used to model this radar, and the simulation performance prediction.

5.1 SYSTEM PARAMETERS

Based on the design parameters of Chapter 3, Table 5-1 outlines pertinent features of the VS and RWS modes.

Table 5-1 Radar parameters

<i>Parameter</i>	<i>VS</i>	<i>HPRF RWS</i>	<i>MPRF RWS</i>
PRF (kHz)	170, 171	135, 136.154	8-20*
Peak power (Watts)	1550	1550	1550
Average power (Watts)	465	465	161-403*
Pulsewidth (μ s)	1.765, 1.754	2.222, 2.203	13
Pulse compression ratio	1	1	13
Antenna half-power beamwidth	2.5° az, 4.5° el	2.5° az, 4.5° el	2.5° az, 4.5° el

Antenna peak gain (dB)	33	33	33
Doppler filter bandwidth (Hz)	21	33	63-156*
Number of range gates	3	3	36-111*
Transmit frequency (GHz)	9.535-9.870	9.535-9.870	9.535-9.870
Scan rate (degrees/sec)	24	24	24
Time-on-target for each burst (ms)	0.052	0.035	0.009-0.018*
Number of coherently integrated pulses per burst	8192	4096	128

* numbers shown are burst dependent

In order to predict radar performance, an estimate must be made of system losses and other factors that lower the signal-to-interference ratio (SIR). These include:

- a. receiver noise,
- b. atmospheric attenuation, L_a ,
- c. target fluctuation, L_f ,
- d. range gate straddling and eclipsing,
- e. CFAR loss,
- f. antenna beamshape,
- g. amplitude weighting,

- h. quantization, and
- i. Doppler filter straddle.

It is convenient to summarize the effect of the receiver noise in terms of a single source at the antenna output terminal, represented by an input termination resistor with a effective temperature T_s [1]. This results in a receiver noise with spectral density $N_0 = kT_s$, where k is Boltzmann's constant. The value of T_s normally takes into account the contributions of the antenna, the RF components, and that of the receiver itself. In SAFIRE, these are combined in a single equation, resulting in $T_s = 290 F_n$, where F_n is the noise figure. The default value used for the noise figure in SAFIRE has been used here: $F_n = 6$ dB. Once the CP-140 receiver to be used with the air-to-air modes is constructed, the actual receiver noise figure should be measured and incorporated in the theoretical performance prediction.

Attenuation of radar waves in clear air is negligible at the lowest radar frequencies, but becomes significant in the microwave band. At sea level, the atmospheric attenuation coefficient for X-band, L_a , is 0.024 dB/km [1]. A lower atmospheric density results in less atmospheric attenuation of the signal. The clear-air attenuation can be significant for the air-to-air modes under consideration, and must be included in the analysis. Note that different types of weather will attenuate the radar signal even further. Of all weather-related effects, rain and wet snow attenuate the radar signal the most, while dry snow, clouds, and fog have lesser effects. The additional effects of weather are not considered in this analysis.

The fluctuation loss for a Swerling Case 1 ("slow" fluctuation, Rayleigh distribution) target varies for each mode. When results are reported for a Case 1 target, the fluctuation loss, L_f , is included.

Note that two loss factors are not explicitly calculated: the CFAR loss, L_{CFAR} , and the range gate straddling/eclipsing loss, L_{rg} . These two loss factors are incorporated in the detection model described in the next section. The other losses, including antenna beamshape loss, amplitude weighting loss, quantization loss, and Doppler filter straddle loss, are incorporated into a single system loss, L_s . Following a similar design in [8], $L_s = 2.8$ dB. Black [34] uses a total loss factor of 10 dB to account for all loss factors that lower the SIR. He notes that experience has shown that a total loss of between 8 to 12 dB will exist in any practical radar implementation. With $F_n = 6$ dB, $L_s = 2.8$ dB, and range gate straddling and CFAR losses included later, a similar total loss factor is used in the calculations in this thesis.

Using the same definitions as Chapter 3 and including the effects of atmospheric attenuation and target fluctuation, Equation 3-1 can be rewritten as

$$\frac{S}{N} = \frac{PG^2\lambda^2\sigma_i T_i f_p \tau}{(4\pi)^3 R^4 (kT_s) L_s L_a L_f} \quad (5-1)$$

The quantity $T_i f_p$ represents the number of integrated pulses for the on-target illumination time of T_i . The inclusion of this term in this form of the radar range equations accounts for the SNR gain achieved by coherent integration.

Equation 5-1 applies directly to the HPRF modes, but is missing the clutter contribution that is present in MPRF RWS mode. Section 4.3 showed that in MPRF

RWS mode, detection is normally clutter-limited rather than noise-limited. The question arises as to what estimate of signal-to-clutter ratio should be used in predicting the performance of the MPRF RWS mode. Clutter and noise returns in Region 1 (noise/weak clutter) were averaged at five different altitudes for the baseline scenario (500 m, 1000 m, 3000 m, 5000 m, 6000 m). The values for the average CNR in Region 1 varied from a high of 8.17 dB to a low of 6.78 dB, with an average of 7.65 dB. Since $\text{SNR} = \text{SCR} + \text{CNR}$ in terms of dBs, the average SCR can be estimated as $\text{SNR} - 7.65$ dB. To calculate detection ranges, then, the SNR from Equation 5-1 was reduced by a factor of 5.82 for MPRF RWS mode to account for average SCR. For simplicity, this factor is included as an additional loss factor for sidelobe clutter, L_{sc} , in Equation 5-2.

One final factor must be taken into account for MPRF RWS. In the previous chapter, it was found that the design P_{fa} had to be approximately an order of magnitude smaller for MPRF RWS to achieve the actual P_{fa} . This loss, L_{nr} , is due to the non-Rayleigh characteristics of the sidelobe clutter. This is compensated for in the theoretical calculations by using a design P_{fa} of 10^{-7} when reporting results for a P_{fa} of 10^{-6} .

Note that the compensation for clutter has been determined here in an empirical manner, whereas the contribution due to clutter is normally calculated deterministically based on a particular distribution [5]. If, for instance, it were known that the clutter actually exhibited Weibull characteristics, the larger required SCR due to non-Rayleigh characteristics could be calculated directly. An example given in [5] shows that the SCR must be 4.2 dB higher to detect the same target with the same P_{fa} in Weibull-distributed (as opposed to Rayleigh-distributed) clutter. The use of L_{nr} in addition to

the average CNR from SAFIRE should provide a more realistic performance prediction for the CP-140 than if a specific clutter distribution were assumed.

Recall that the CFAR loss and the range gate straddling and eclipsing losses are incorporated in the detection model in the next section for the HPRF RWS and VS modes. Similarly, the CFAR loss and the range gate eclipsing loss are incorporated in the detection model for MPRF RWS mode. For simplicity, however, the range gate straddling loss is incorporated as a single value, L_{rgs} , for MPRF RWS. This allows use of a single P_d vs. Range curve to represent the overall performance of the eight PRFs in MPRF RWS mode. Following Aronoff and Greenblatt, a range gate straddling loss of 0.6 dB is used [8]. Equation 5-2 is used, then, for MPRF RWS mode prediction:

$$\frac{S}{C} = \frac{PG^2\lambda^2\sigma_i T_i f_p \tau}{(4\pi)^3 R^4 (kT_s) L_s L_a L_f L_{sc} L_{nr} L_{rgs}} \quad (5-2)$$

All loss factors are tabulated in Table 5-2.

Table 5-2 Loss Factors

<i>Loss Factor</i>	<i>VS</i>	<i>HPRF RWS</i>	<i>MPRF RWS</i>
F_n , noise figure	6.0 dB	6.0 dB	6.0 dB
L_a , atmospheric losses	0.024 dB/km	0.024 dB/km	0.024 dB/km
L_f , target fluctuation loss	Calculated for each scenario and included for Case 1 target	Calculated for each scenario and included for Case 1 target	Calculated for each scenario and included for a Case 1 target
L_s , system losses	2.8 dB	2.8 dB	2.8 dB

L_{sc} , loss factor for sidelobe clutter	N/A	N/A	7.65 dB
L_{nr} , loss due to non-Rayleigh characteristics of sidelobe clutter	N/A	N/A	Taken into account by designing for an extra order of magnitude in P_{fa}
L_{CFAR} , CFAR loss, and L_{rg} , loss due to range gate straddling and eclipsing	Included in detection model, next section	Included in detection model, next section	CFAR loss and loss due to range gate eclipsing included in detection model, next section. Loss due to range gate straddling: 0.6 dB

5.2 THEORETICAL RESULTS

5.2.1 Single-Burst Detection

As described in the previous chapter, the envelope of signal plus noise, at the time of peak signal output, has a Rician pdf. Based on this distribution, a probability of detection can be calculated for a given probability of false alarm and signal-to-noise ratio. The resulting P_d is often plotted versus SNR as a family of curves for each P_{fa} of interest and each Swerling case of interest [1, 4, 7, 11, 21]. These are sometimes referred to as "Meyer" plots after one of the original authors [21]. Examples of the SNR required to achieve a specific P_d for a given design P_{fa} are shown in Table 5-3 for three probabilities of common interest for Swerling Cases 0 and 1. These SNR values, which are obtained directly from the Meyer plots, can be interpreted to be the required SNR after coherent integration.

Table 5-3 Required SNR for a Single PRF Burst

	$P_{fa} = 10^{-6}$		$P_{fa} = 10^{-7}$		$P_{fa} = 10^{-8}$	
	Case 0 (dB)	Case 1 (dB)	Case 0 (dB)	Case 1 (dB)	Case 0 (dB)	Case 1 (dB)
$P_d = 0.50$	11.2	12.8	11.9	13.5	12.5	14.0
$P_d = 0.85$	12.8	19.3	13.4	20.0	13.9	20.5
$P_d = 0.90$	13.2	21.4	13.8	22.0	14.2	22.4

Note the sometimes significant difference between the required SNR for the two types of targets. The higher SNR for the Case 1 target reflects a target fluctuation loss that is not present for the Case 0 target.

5.2.2 Cumulative Detection

Operational detection requirements are often defined in terms of a cumulative detection range [4]. Cumulative detection may be considered to be the binary integration of target returns from multiple radar "looks". There are two different time frames for which cumulative detection can be calculated. The first is the cumulative detection that is available in a single scan, and the second is the cumulative detection that is available with multiple scans. Only single-scan cumulative detection is discussed here, because it is considered to be the most meaningful.

It is possible to report cumulative detection performance in a single scan for modes that illuminate the target with more than one burst in a single scan. Recall that in MPRF RWS mode the target is illuminated with eight bursts at different PRFs in a single scan. Detection is required on at least three PRFs to resolve range and Doppler

ambiguities. If the target is illuminated by more than three PRFs, the cumulative detection range is calculated based on the binomial distribution [35]. Following Hovanessian, the probability of obtaining exactly r observations out of v attempts is:

$$P_r = \binom{v}{r} \theta^r (1-\theta)^{v-r} \quad , \quad (5-3)$$

where θ is the probability of detection for each PRF burst [35]. The probability of obtaining at least three out of eight detections can be found by computing P_0 , P_1 , and P_2 , and subtracting their sum from unity. For example, the value of θ resulting in an 85% probability that a detection will be made on at least three out of eight bursts is $\theta = 49.7\%$. Thus, if there are actually eight PRFs illuminating the target in a scan (i.e., the target is eclipsed in none of the PRFs), then a single-burst probability of detection of only 49.7% is required to achieve an overall P_d of 85% for MPRF RWS.

Note, however, that in [35] the author does not mention the possibility that the target will be eclipsed on at least one of the eight PRFs. Indeed, from Section 3.3, there are an average of 6.6 PRFs covering each point in the range-Doppler space of interest. Rather than mapping out θ for the entire range-Doppler space, it is reasonable to consider the cumulative detection probability of detecting the target on at least three of six PRFs rather than three of eight. For MPRF mode, for a P_d of 85%, θ is equal to 62.4%. This represents the overall performance of the mode while still taking into account the fact that the target is not likely to appear on all eight PRFs. The use of six PRFs (when there is actually an average of 6.6 PRFs illuminating the target) gives a conservative performance estimate. The values of θ for the three detection probabilities of interest for MPRF RWS are listed in Table 5-4.

In VS mode, on the other hand, there are two different PRFs transmitted during the TOT. Thus, the single-scan probability of detection, P_d , is simply the cumulative probability of detecting the target on at least one of the two PRFs. Based on Equation 5-3, P_d is given by

$$P_d = \binom{2}{1}\theta(1-\theta) + \binom{2}{2}\theta^2 \quad (5-4)$$

for the VS mode. For an 85% cumulative probability of detecting the target when the target is not blanked on either of the PRFs, the single-burst probability of detection θ is only 61.4%. The values of θ for the three single-scan detection probabilities of interest for VS mode are listed in Table 5-4.

For HPRF RWS, detection is required on all three PRFs to perform range resolving. Thus, the single-scan probability of detection is given by:

$$P_d = \binom{3}{3}\theta^3 \quad (5-5)$$

In order to achieve an overall probability of detection of 90%, there must be a single-burst probability of detection of 96.6% on each stage. Thus, the SNR required to obtain a given single-scan P_d is higher than that required to obtain the same value for a single-burst P_d . The values of θ for each mode and single-scan P_d of interest are shown in Table 5-4. Based on these values of θ , the corresponding required SNRs obtained from the Meyer plots for a P_{fa} of 10^{-6} are shown for the three modes in Table 5-5.

Note that in an operational system, the threshold multiplier settings are based on an expected time interval between false alarms. As will be seen in Subsection 5.2.5, a common time interval for each mode requires a different P_{fa} setting for each mode. A common P_{fa} of 10^{-6} is used, however, in this initial discussion in order to readily compare the performance of each of the modes. Appropriate operational threshold multiplier values for each mode are discussed in Subsection 5.2.6.

Table 5-4 Single Scan Cumulative Detection

	θ for VS	θ for HPRF RWS	θ for MPRF RWS
$P_d = 0.50$	29.4%	79.4%	42.3%
$P_d = 0.85$	61.4%	94.7%	62.4%
$P_d = 0.90$	68.5%	96.6%	66.8%

Table 5-5 Required SNR for Single Scan Detection

	VS Mode		HPRF RWS Mode		MPRF RWS Mode	
	Case 0 (dB)	Case 1 (dB)	Case 0 (dB)	Case 1 (dB)	Case 0 (dB)	Case 1 (dB)
$P_d = 0.50$	10.2	10.1	12.5	17.9	11.6	12.5
$P_d = 0.85$	11.8	14.4	13.6	24.0	12.4	15.2
$P_d = 0.90$	12.0	15.6	13.8	25.0	12.6	16.0

For VS mode where the single-scan $P_d = 0.50$, the Case 1 target actually requires a smaller SNR than for the Case 0 target. This is characteristic of the pdfs describing the fluctuating targets; as mentioned in Chapter 4, Case 0 requires a smaller SNR than the

others only for probabilities of detection greater than 30%. In this case, a single-PRF P_d of only 29.4% is required to achieve the single-scan $P_d = 0.50$.

5.2.3 Range Performance

At this point it is useful to comment on the method used here to report expected radar range performance. There are many different factors that should be taken into consideration to give realistic performance predictions, yet they are sometimes excluded or manipulated to exaggerate the performance of the radar. These factors include:

- a) range and Doppler blind zones,
- b) target models, and
- c) cumulative detection performance.

If radar range performance is reported without commenting on these three factors, it is possible that the report ignores range and Doppler blind zones, employs a constant RCS model, and/or uses multiple scans for cumulative detection. Reporting in this manner is justified if a radar range prediction is desired for this type of situation, but the reader must note that performance in this situation is generally significantly better than, say, for the single-scan probability for a Swerling case 1 target with range and Doppler blind zones taken into account.

Consider an example given by Belcher in [4] for a required detection range of 250 km. His example uses a Swerling 1 target, closing with a velocity of 500 m/s, $P_{fa} = 10^{-6}$, $P_d = 0.9$, and two revisits within the allotted time for cumulative detection. The required SNR for a single-pulse detection of this target is 22 dB, whereas the

required SNR for the cumulative detection of this target is only 9 dB. Thus, the radar specifications could quote a cumulative detection range of 250 km with a probability of detection of 90% as long as the SNR was at least 9 dB at 250 km. The Meyer plots show very different performance predictions for this radar if cumulative detection is not used. If the SNR were 9 dB, the Meyer plots show that a single-pulse probability of detection of only 20% would be achieved at 250 km. This example highlights the need to look closely at whether the reported range performance is based on single-scan or cumulative detection performance for any radar system.

Other factors that are often left out of range performance calculations are the range and Doppler blind zones. To provide a fair estimate of performance, these factors should be taken into consideration. Because detections are only made in the noise-limited region for HPRF modes, it must be stated that targets with relative closing or receding velocities within $\pm V_0$ will not be detected. The range blind zones in HPRF often encompass a large percentage of the total range. As was shown in Sections 3.6 and 3.7, the use of a second PRF enhances the range coverage of the radar for HPRF modes.

When range performance is quoted for a given radar in a particular engagement scenario, the target could quite possibly lie completely within a range gate, such that range eclipsing and range gate straddling do not have to be considered. If an overall range estimate is given for a certain mode, however, it should take into account the probability that a target will not be seen at any one point in time due to range eclipsing and Doppler blanking. The cumulative probability of detection could be used to make the range performance seem better, since a target that is eclipsed in the first scan is likely to move to another range and/or Doppler bin for a second scan. Once again, it

is not inaccurate to report the radar range performance using any of these techniques, but the reader must take the method of calculation into consideration. In reporting here the cumulative detection performance for the CP-140, the single-scan probability of detection is reported rather than a multi-scan cumulative probability of detection.

The discussion of expected range performance in Subsections 5.2.3.1 to 5.2.3.3 assumes target detection in Region 1 of the range-Doppler map. A discussion of expected range performance in Regions 2 and 3 is included at the end of Subsection 5.2.3.4.

5.2.3.1 Single-Burst Performance for VS Mode

An aspect that has not yet been taken into consideration for the VS and HPRF RWS modes is range gate straddling and eclipsing. For the VS and HPRF RWS modes, target eclipsing will be calculated in conjunction with the final performance factor, the CFAR loss. Equations 4-17 and 4-24 represent the predicted P_d including the CFAR loss for CA and TM-CFAR. As TM-CFAR is being recommended for use on the CP-140, TM-CFAR was used to calculate range performance. It was assumed that 40 cells are used in the reference window for the noise-limited detection in the HPRF modes. As well, the range gate responses for HPRF RWS and VS (from Figures 3-16 and 3-19) were incorporated using the range coverage of both PRFs in the PRF set. The SNR was calculated using Equation 5-1 using range increments of 100 m, and then substituted into Equation 4-24. SNR vs. Range and P_d vs. Range curves for VS mode are shown in Figures 5-1 and 5-2 for a P_{fa} of 10^{-6} and a constant target RCS of 5 m^2 . Note that because the SNR vs. Range is plotted with a dB scale, range intervals where the $\text{SNR}=0$ due to full eclipsing are not displayed. These range intervals are shown with $P_d=0$, however, in the P_d vs. Range curve.

Note that using this method to account for the range gate straddling and eclipsing losses has not been found in the literature, but is more representative of the true radar performance than including a single range gate straddle loss.

Figure 5-1. SNR vs. Range for VS Mode for a Single Burst

Figure 5-2. P_d vs. Range for VS Mode for a Single Burst

There are several interesting features to note from Figures 5-1 and 5-2. The intervals near zero ranges that are fully eclipsed on both PRFs have a $P_d = 0$. Because the radar has sensitivity to targets that are partially eclipsed, the regions where $P_d = 0$ are very limited. Note that as soon as a fraction of the received pulse from a close-in target lies within a range gate, the high SNR associated with the short range means that there is once again a very high probability of detection. Recall that R_E was chosen as the distance at which it is acceptable for the targets of interest to be eclipsed on both PRFs. In VS mode, the two PRFs were chosen such that the eclipsed regions match up again at $R_E = 150$ km. Near 150 km in range, there are once again intervals where the $P_d = 0$. System level performance is often specified in terms of detection range R_n , where n represents the percent probability of detection at that range for a target of specified RCS. From Figure 5-2, $R_{85} = 83.2$ km for a P_{fa} of 10^{-6} and a constant target RCS of 5 m^2 .

To demonstrate the effect of the choice of R_E and f_E , the P_d vs. Range is plotted in Figure 5-3 for $R_E = 120$ km. Notice that there are fewer eclipsed portions near zero range, and that the overlap of the eclipsed portions occurs at 120 km instead of 150 km. Because the P_d may still be above 40% for a 5 m^2 target at 120 km, it is advantageous to have a longer R_E to minimize eclipsing. At 150 km, the P_d has dropped well below 20%, so eclipsing of this target does not significantly degrade the overall performance. R_E could be pushed out even further, but the close-in eclipsed regions would mask close-in targets even more.

Figure 5-3. P_d vs. Range for VS Mode for a Single Burst, $R_E = 120$ km

5.2.3.2 Single-Burst Performance for HPRF RWS Mode

SNR vs. Range and P_d vs. Range curves for HPRF RWS mode are shown in Figures 5-4 and 5-5. The plots for HPRF RWS mode are similar to those shown for VS mode. In this case, however, the PRFs were chosen such that targets of interest were again eclipsed at 130 km. When the CFAR loss and pulse eclipsing are taken into consideration, $R_{g5} = 73.0$ km for a P_{fa} of 10^{-6} and a constant target RCS of 5 m^2 .

Figure 5-4. SNR vs. Range for HPRF RWS Mode for a Single Burst

Figure 5-5. P_d vs. Range for HPRF RWS Mode for a Single Burst

5.2.3.3 Single-Burst Performance for MPRF RWS Mode

As a target will be uneclipsed on an average of 6.6 PRFs, it is reasonable to investigate the single-burst performance of the MPRF RWS mode with no additional eclipsing loss. Target eclipsing was somewhat compensated for in the single-scan performance of MPRF RWS in the values shown in Table 5-5, as it was assumed that the target was covered by only six of the eight PRFs. As the true value was 6.6 PRFs, no further eclipsing is added to the MPRF RWS single-scan performance prediction. The loss of 0.6 dB mentioned earlier was incorporated to account for range gate straddling for both single-burst and single-scan performance.

The SCR was calculated using Equation 5-2 using range increments of 100 m, and then substituted into Equation 4-24. SCR vs. Range and P_d vs. Range curves for MPRF RWS mode are shown in Figures 5-6 and 5-7 for a target detected on a single PRF burst. The loss due to the non-Rayleigh clutter characteristics, L_{nr} , was calculated from Table 5-3 by subtracting the listed SNR value for $P_{fa} = 10^{-7}$ from the SNR value given for P_{fa} of 10^{-6} , for the Case 0 target and the same P_d . This calculation, which reflects the requirement to use a design P_{fa} of 10^{-7} to achieve an actual P_{fa} of 10^{-6} , yields $L_{nr} \approx 0.6$ dB.

Figure 5-6. SCR vs. Range for MPRF RWS Mode for a Single Burst

Figure 5-7. P_d vs. Range for MPRF RWS Mode for a Single Burst

Note that target detection is not normally feasible for SNR or SCR values less than 0 dB. The SCR falls below 0 dB at around 90 km for MPRF RWS. The relatively short-range performance of the MPRF RWS mode (compared to the HPRF modes) is due to several factors: the shorter coherent integration time, the clutter-limited (as opposed to noise-limited) detection, and the requirement to design for a P_{fa} of 10^{-7} to achieve an actual P_{fa} of 10^{-6} . From Figure 5-7, $R_{85} = 30.8$ km for an actual P_{fa} of 10^{-6} and a constant target RCS of 5 m^2 .

5.2.3.4 Single-Scan Performance

This process was repeated to predict the single-scan detection probabilities for an actual P_{fa} of 10^{-6} and a constant target RCS of 5 m^2 . The results are shown in Figures 5-8 to 5-10. The change in P_d due to detection on multiple PRFs was approximated by adding or subtracting the difference in SNR between Tables 5-3 and 5-5 to S in Equation 4-24. Note that the difference in SNR is not the same for all probabilities of detection (i.e., the difference is 1.0 dB for R_{50} , 1.0 dB for R_{85} , and 1.2 dB for R_{90} for Case 0 targets). Figures 5-8 to 5-10 assume the difference in SNR for R_{85} . The values listed in Table 5-6, however, use the correct difference in SNR for each P_d for the two types of targets.

Figure 5-8. P_d vs. Range for VS Mode for Single Scan

Figure 5-9. P_d vs. Range for HPRF RWS, Single Scan

Figure 5-10. P_d vs. Range for MPRF RWS, Single Scan

The single-scan detection ranges for R_{50} , R_{85} , and R_{90} are shown in Table 5-6 for Case 0 and Case 1 targets. The change in P_d due to increase in required SNR for a Case 1 target was approximated using the same method (i.e., by subtracting the difference in SNR from S in Equation 4-24).

Table 5-6 Detection Ranges for a Single Scan With All Losses Incorporated, $P_{fa} = 10^{-6}$

	VS Mode		HPRF RWS Mode		MPRF RWS Mode	
	Case 0	Case 1	Case 0	Case 1	Case 0	Case 1
R_{50} (km)	121.3	121.6	96.2	71.9	46.2	44.0
R_{85} (km)	87.4	76.0	69.7	40.3	32.4	27.8
R_{90} (km)	79.1	65.5	64.7	36.2	29.2	24.2

Based on the values in Table 5-6, the detection ranges for each mode are plotted in Figure 5-11 for a 5 m² Swerling Case 1 target. The outer ring shown for each mode represents R_{95} for a single scan. Note that each of the modes is shown scanning a different sector simply to demonstrate that only one mode is active at a time, and that the CP-140 is not drawn to scale. In normal operation, each would scan a full azimuthal sweep around the aircraft. Note also that target detection is not limited to the ranges between the inner and outer rings for HPRF RWS mode and VS mode. The inner ring shown here merely helps to demonstrate the typical usage of each mode. VS mode is commonly used as the initial search mode as it provides the longest detection range. When a target moves within the inner ring shown for VS, it is advantageous to switch to HPRF RWS, as HPRF RWS provides both target range and velocity information. When a target is within the inner ring shown for HPRF RWS, it is better to switch to MPRF RWS, since MPRF RWS provides more accurate target range, and detection is not restricted to high closing/receding velocity targets.

Figure 5-11. CP-140 Air-to-Air Mode Detection Ranges for a 5 m² Case 1 Target,
 $P_{fa} = 10^{-6}$, Single-Scan $P_d = 85\%$

One final note should be made about the expected performance of the MPRF RWS mode for targets in Regions 2 and 3. Recall from Subsection 4.3.3.3 that the

reference window for Region 2 used $N=6$ cells in calculating the threshold, and that Region 3 used $N=16$. Recall from Subsection 4.3.4 that CA-CFAR was used in each due to the small number of cells in the reference windows. From Equation 4-17, $P_d = [1 + T/(1 + S)]^{-N}$. Measured using the baseline scenario, the CNR is typically 6.50 dB higher in Region 2 (average CNR 14.15 dB), and typically 7.25 dB higher in Region 3 (average CNR 14.90 dB). Using these average CNRs rather than the 7.65 dB assumed for Region 1, the P_d at 32.4 km (R_{q5} for a Case 0 target in Table 5-6) is equal to only 35% in Region 3 and 15% in Region 2. Thus, while there is some ability to detect targets lying in Regions 2 and 3, it is severely restricted by the clutter-to-noise level. This highlights the need for a low sidelobe antenna for good performance from a medium PRF mode.

5.2.4 Maximum Unambiguous Detection Range

The expected RCS of a large jet at X-band is typically 200 m². The size of such a target helps demonstrate the need for the long maximum unambiguous detection ranges allowed for in Chapter 3. Recall that fill pulses were used to allow for maximum unambiguous detection ranges of 300 km for MPRF RWS, 500 km for HPRF RWS, and 550 km for VS.

Consider the following "worst case scenario" for calculating maximum unambiguous detection ranges. Barton [1] shows that for an elevation angle of 10° at 10 GHz, the total two-way atmospheric attenuation is a constant 0.5 dB at slant ranges longer than 100 km. Also, the target may possibly lie in a portion of the range-Doppler map where the CNR is 1.50 dB, rather than the 7.65 dB used earlier as an average. Assume the following: 1) a constant RCS target, 2) a total atmospheric attenuation of 0.5 dB, and 3) for MPRF RWS, the target is detected where the CNR is

only 1.50 dB. The P_d for each mode at the design maximum unambiguous detection range for this situation is shown in Table 5-7.

Table 5-7 Probability of Detection at Maximum Unambiguous Detection Range

RCS (m ²)	P_d at 550 km in VS Mode (%)	P_d at 500 km in HPRF RWS Mode (%)	P_d at 300 km in MPRF RWS Mode (%)
100	3.22	1.62	1.41
200	14.03	8.82	8.01

Thus, while the probabilities that the largest targets will be detected at these ranges are small, they are still large enough to warrant the large design maximums. The use of the long maximum unambiguous detection ranges ensures that a large target is not erroneously reported at a much shorter, ambiguous range.

5.2.5 Threshold Multiplier

Given the expected performance of the air-to-air modes, a range of threshold multipliers must be selected for operational testing. The CFAR processor monitors the rate at which false alarms and/or detections occur, and must be allowed to alter the threshold multiplier, T , if the rate is too high or too low. This range of multipliers must be wide enough to allow maximum sensitivity in ideal background interference, while preventing an unmanageable number of false alarms in severe clutter.

The analysis to date has used an actual P_{fa} of 10^{-6} . The meaning of an actual P_{fa} of 10^{-6} in terms of time between false alarms is discussed in the following subsections for the different modes. Recall that the total time-on-target for each of the

modes is approximately 104 ms. In that time, MPRF RWS emits eight different PRF bursts, HPRF RWS emits three frequency-modulated bursts, and VS mode emits two different PRF bursts.

5.2.5.1 Threshold Multiplier for MPRF RWS Mode

At an average 14 kHz PRF, the MPRF RWS mode range-Doppler map consists of 58 range gates by 128 Doppler bins. If the MLC width from the baseline scenario is used, there are four Doppler bins blanked. Thus, there are 7,192 cells available for test. After eight PRFs, approximately 57,536 cells will have been tested by the CFAR technique. At an actual P_{fa} of 10^{-6} , there are an average of 0.007 false alarms on a single PRF burst, and an average of 0.058 false alarms after all eight PRFs have been investigated.

Recall, however, that a target must correlate on at least three of the eight PRFs to be declared. This suggests that the resulting P_{fa} would be lower after all eight bursts have been processed. The binomial distribution theorem can be used to calculate the probability of correlating at least three false alarms to produce a false target declaration. Based on a single-cell P_{fa} of 10^{-3} , the probability that a false alarm will be seen in the same cell on at least three of eight PRFs is:

$$P_{3of8} = \sum_{i=3}^8 \binom{8}{i} (10^{-3})^i (1 - 10^{-3})^{8-i} \quad (5-6)$$

There are 7,192 chances for this to occur, resulting in an average number of false alarms, P , over the eight bursts that is given by $P = (7192)P_{3of8}$. When the P_{fa} for each cell is 10^{-3} , $P = 5.58e-8$. Based on an average of $5.58e-8$ false alarms in 104 ms,

one false target declaration would be expected every 518 hours. This calculation is repeated for several different probabilities of false alarm in Table 5-8.

Table 5-8 MPRF RWS False Alarm Rates

Actual P_{fa} achieved with CFAR	Estimated design P_{fa} required	Average number of false alarms expected in a single range- Doppler map	Average number of false alarms after all eight bursts	Expected time between false target declarations
5×10^{-2}	5×10^{-3}	359.6	5.79×10^{-3}	17 seconds
3.27×10^{-2}	3.27×10^{-3}	235.2	1.73×10^{-3}	1 minute
1.48×10^{-2}	1.48×10^{-3}	106.4	1.72×10^{-4}	10 minutes
10^{-2}	10^{-3}	71.9	5.39×10^{-5}	32 minutes
5×10^{-3}	5×10^{-4}	36.0	6.87×10^{-6}	4 hours

Assuming that one minute is a reasonable length of time between false targets, T would be set to a starting value based on a design P_{fa} of 3.27×10^{-3} . The CFAR processor would then be programmed to alter T to attempt to maintain a one-minute interval between false alarms. Note that the threshold multiplier can be monitored during a flight to ensure that the values selected are reasonable. For instance, the CFAR processor should never need to use a design P_{fa} higher than 3.27×10^{-2} if a one-minute interval is desired between false alarms.

5.2.5.2 Threshold Multiplier for HPRF RWS Mode

Recall that the HPRF RWS mode uses a PRF of 135 kHz and has three range gates per PRI. There are three separate bursts emitted during one TOT, each with a different frequency modulation. The range-Doppler map from a single burst consists of 4,096 Doppler bins by three range gates, for a maximum of 12,288 cells. At a cruise velocity of 180 m/s, there is a Doppler notch of $\pm 11,650$ Hz, or a total of 708 Doppler bins per range gate. Thus, there is normally a total of 10,164 cells tested by the CFAR processor in each burst.

The same type of false alarm analysis can be carried out for the HPRF RWS mode as was done previously for MPRF RWS mode. Note that for the HPRF modes, the design P_{fa} is equal to the actual P_{fa} , as detection is carried out in the noise-limited (clutter-free) region. In HPRF RWS mode, as previously discussed, the detection must correlate on all three bursts to be declared a target. The probability, P_{3of3} , that a false alarm would be seen in the same range-Doppler bin on the all three PRFs, however, is

$$P_{3of3} = \binom{3}{3} P_{fa}^3 = P_{fa}^3 \quad (5-7)$$

The average number of false alarms, P , after processing all three bursts is then $P = (10164)P_{3of3}$. The value of P is calculated for several P_{fa} values in Table 5-9.

Table 5-9 HPRF RWS False Alarm Rates

Actual P_{fa} achieved with CFAR	Average number of false alarms expected in a single range-Doppler map	Average number of false alarms after all three bursts	Expected time between false target declarations
10^{-2}	101.64	$1.01\text{e-}2$	10 seconds
5×10^{-3}	50.82	$1.27\text{e-}3$	1 minute
2.57×10^{-3}	26.12	$1.73\text{e-}4$	10 minutes
10^{-3}	10.16	$1.02\text{e-}5$	3 hours
5×10^{-4}	5.08	$1.27\text{e-}6$	23 hours

For HPRF RWS mode, then, it would be reasonable to set the threshold multiplier to a starting value based on a design P_{fa} of 5×10^{-3} . Because HPRF RWS involves noise-limited detection, there should be no reason to alter T unless a different time between false target declarations is desired.

5.2.5.3 Threshold Multiplier for VS Mode

Recall that the VS mode uses a PRF of 170 kHz and has three range gates per PRI. The range-Doppler map from a single burst consists of 8,192 Doppler bins by three range gates, for a maximum of 24,576 cells. At a cruise velocity of 180 m/s, there is a Doppler notch of $\pm 11,650$ Hz, or a total of 1,124 Doppler bins per range gate. Thus, there is normally a total of 21,204 cells tested by the CFAR processor. Two bursts of pulses are sent, but each is processed independently. A target does not need to appear on both to be declared. Thus, the false alarm rate in a single PRF burst

determines the overall false alarm rate for VS mode. Each burst takes approximately 0.05 s. The VS mode false alarm rates are shown in Table 5-10.

Table 5-10 VS Mode False Alarm Rates

Actual P_{fa} achieved with CFAR	Average number of false alarms expected in a single range-Doppler map	Expected time between false target declarations
5×10^{-7}	1.06e-2	5 seconds
10^{-7}	2.12e-3	24 seconds
5×10^{-8}	1.06e-3	47 seconds
3.93×10^{-8}	8.33e-4	1 minute
3.93×10^{-9}	8.33e-5	10 minutes

Notice that the P_{fa} required to achieve a desired time between false alarms is significantly lower when there is no correlation of targets required. Indeed, a P_{fa} of 3.93×10^{-8} is required just to achieve the one-minute interval. As with HPRF RWS, the only reason to alter T is when a different time between false alarms is desired.

5.2.6 One-Minute and Ten-Minute False Alarm Intervals

Based on the theoretical values calculated in the previous subsections to achieve a one-minute interval between false alarms, the detection ranges for a 5 m² target for each mode were calculated in the same fashion as in Section 5.2.3, where cumulative detection is assumed within a single scan. The results are shown in Table 5-11.

Table 5-11 Detection Ranges for a One-Minute False Alarm Interval

	VS Mode		HPRF RWS Mode		MPRF RWS Mode	
	Case 0	Case 1	Case 0	Case 1	Case 0	Case 1
R_{50} (km)	113.5	115.3	122.0	96.2	59.4	56.4
R_{85} (km)	82.0	71.0	89.5	56.9	43.2	41.4
R_{90} (km)	74.4	61.3	80.9	44.7	39.6	37.0

Notice that the detection ranges for the VS mode are slightly shorter than those shown in Table 5-6 due to the lower P_{fa} . The detection ranges for HPRF RWS mode, however, are much longer, as a higher false alarm rate can be tolerated.

Table 5-12 is included to demonstrate the difference in range if the theoretical values for a ten-minute interval between false alarms are used. Cumulative detection is assumed within a single scan.

Table 5-12 Detection Ranges for a Ten-Minute False Alarm Interval

	VS Mode		HPRF RWS Mode		MPRF RWS Mode	
	Case 0	Case 1	Case 0	Case 1	Case 0	Case 1
R_{50} (km)	109.9	110.0	118.4	93.0	57.2	54.4
R_{85} (km)	79.2	68.7	86.6	54.9	41.8	40.0
R_{90} (km)	71.7	59.4	78.7	42.7	38.2	35.8

The detection ranges for each mode are plotted in Figure 5-12. In each case, the outer ring shown for each mode represents R_{85} for a single scan for a 5 m² Swerling Case 1 target for a one-minute false alarm interval. Note that the modes are

shown scanning a different sector simply to demonstrate that only one mode is active at a time. In normal operation, each would scan a full azimuthal sweep around the aircraft.

Figure 5-12. Detection Ranges for One-Minute False Alarm Interval, 5 m², Case 1 Target

P_d vs. Range curves are included in Appendix A for several different target RCSs. Each is based on the one-minute false alarm interval and single-scan cumulative detection.

The preceding analysis assumes only basic radar signal processing. There are more sophisticated processing schemes such as *track-before-detect* (TBD) that can allow a much longer detection range for a given false alarm rate. By requiring multiple correlated detections before declaring a target, TBD can use a lower threshold while still maintaining the desired time between false alarms.

5.3 SIMULATIONS

Simulations were carried out using SAFIRE to predict the radar performance of the MPRF RWS mode on the CP-140. A brief description of SAFIRE was given in Subsection 4.3.2. Simulations were not conducted for the HPRF modes because detection is carried out in the noise-limited region, and the theoretical analysis is

expected to be accurate. The theoretical calculations for the MPRF RWS mode, however, were based on assumptions about the clutter that should be tested with simulated data. Note that the entire design of the CFAR processor and the need for L_{nr} were based on an analysis of the false alarm performance using simulated data from SAFIRE. Thus, target detection in the MPRF RWS mode is considered in this section.

The clutter simulation used in this section follows the parameters used in the previous chapter. The data are extracted after quantization with a 12-bit ADC. SAFIRE uses a backscatter coefficient, σ^0 , for vegetation clutter as shown in Figure 5-13. Also plotted for reference are data for farmland, heavy woods, and urban backscatter coefficients at X-band taken from [22].

Figure 5-13. Backscatter Coefficients

All losses are already accounted for in SAFIRE except L_a . To account for atmospheric attenuation, the expected loss for each target was added to SAFIRE's receiver noise figure for each scenario. For instance, when the target was placed at 40 km, $0.024 \text{ dB/km} \times 40 \text{ km} = 0.96 \text{ dB}$ of loss was added to the receiver noise figure. Losses such as L_{sc} , L_{nr} , and L_{CFAR} are not explicitly used in SAFIRE; rather they are inherently included when attempting to use a CFAR technique to detect a target.

5.3.1 Simulation Scenarios

Scenario #1. A target with an RCS of 50 m^2 was placed in the scenario shown in Figure 5-14 with $V_o = 18 \text{ m/s}$ and $V_T = 40 \text{ m/s}$. Note that this is slightly different from the baseline scenario discussed in the last chapter; the downlook angle is 3.87° rather than 15° , and V_o is much slower. This slow ownship velocity places the target return in the noise-limited region, between Doppler bins 31 and 32.

Figure 5-14. Scenario #1

The clutter map for this scenario is shown in Figure 5-15. A cross section in range through the target is shown in Figure 5-16.

Figure 5-15. Clutter Map for Scenario #1

Figure 5-16. Range Cross Section for Scenario #1

As expected with this short range and large target cross section, the signal appears well above the noise in range gate 66. Also apparent in both plots are the range sidelobes caused by pulse compressing the target.

Scenario #2. This scenario is the same as Scenario #1, but with a target RCS of 5 m^2 . The clutter map for this scenario is shown in Figure 5-17. A cross section in range through the target is shown in Figure 5-18. The signal strength is about 10 dB lower than that shown in Figure 5-16, but the target signal is still well above the noise floor.

The performance of the TM-CFAR technique was tested using the data for Scenario #2. Using threshold settings for a design P_{fa} of 10^{-3} , there were 14 false alarms and 7 hits clustered about the location of the target. With a design P_{fa} of 10^{-7} , there were no false alarms, and a total of four hits clustered around the target. The CFAR maps are shown in Figure 5-19.

Figure 5-17. Clutter Map for Scenario #2

Figure 5-18. Range Cross Section for Scenario #2

(a)

(b)

Figure 5-19. CFAR Map for Scenario #2, a) with design $P_{fa} = 10^{-3}$, and b) with design $P_{fa} = 10^{-7}$

Scenario #3. This scenario is the same as Scenario #2, but with an ownship speed of $V_o = 180$ m/s. In this case, the target is located in the sidelobe clutter region. The clutter map for this scenario is shown in Figure 5-20. A cross section in range through the target is shown in Figure 5-21. Note that the average CNR in Region 1 in

this range gate is only 4.43 dB, whereas an average of 7.65 dB was used in the previous section to predict overall performance. The regional TM-CFAR technique with a design P_{fa} of 10^{-7} results in three hits clustered about the target, and no false alarms.

Figure 5-20. Clutter Map for Scenario #3

Figure 5-21. Range Cross Section for Scenario #3

Figure 5-22. CFAR Map for Scenario #3

Scenario #4. For Scenario #4, a target was placed at a range of 55.0 km with the parameters shown in Figure 5-23 to test the outer detection limits of the MPRF RWS mode. In this case $0.024 \text{ dB/km} \times 55 \text{ km} = 1.32 \text{ dB}$ of loss was added to the receiver noise figure to account for atmospheric attenuation. The simulation was run fifty times, with each simulation using a different random number seed to produce the clutter data (note that the simulator takes about 16 hours to generate each range-Doppler clutter map). The clutter map for the first of these runs, Simulation #4a, is shown in Figure 5-24. A range cross section through the target in Doppler bin 32 is shown in Figure 5-25.

Figure 5-23. Scenario #4

Figure 5-24. Clutter Map for Simulation #4a

Figure 5-25. Range Cross Section for Simulation #4a

At a range of 55 km, the target straddles range gates 72 and 73. The regional TM-CFAR technique was run on the data from Simulation 4a using seven different design false alarm rates ($P_{fa} = 10^{-3}$, 5×10^{-4} , 10^{-4} , 5×10^{-5} , 10^{-5} , 10^{-6} , and 10^{-7}). Only the three highest probabilities of false alarm resulted in a target declaration. The threshold levels for these design values are shown in Figure 5-26.

Note from Figure 5-25 that the CNR tends to drop off in the outer range gates, so the SCR in these gates is higher. Because the target has been placed in one of the outer range gates to investigate maximum detection ranges, the target lies in a region where the CNR is approximately 4.5 dB lower than the 7.65 dB used as an average in the theoretical calculations. Note also that the signal in range gate 73 is derived from the pulse compression of only eight range gates rather than thirteen. When these factors are taken into account in the theoretical calculations, $R_{50} = 55.0$ km when a design P_{fa} of 5.79×10^{-6} ($T=0.44$) is used.

Figure 5-26. CFAR Thresholds for Range Gate 73, Simulation #4a

Scenario #4 was repeated using 50 different random number seeds to generate the clutter data. In each case, the maximum value of T that would still detect the target was calculated. Figure 5-28 shows these values and the median value, $T=0.48$. Figure 5-27 shows the corresponding design probabilities of false alarm, along with the median $P_{fa}=2.28e-6$. Thus, a similar design P_{fa} ($2.28e-6$ as compared to $5.79e-6$) for R_{50} was obtained from the simulated data, suggesting that the theoretical calculations are reasonable. Indeed, the simulated data suggest that the theoretical calculations are slightly pessimistic. The theoretical analysis suggested that $T=0.44$ would be needed

to achieve a 50% probability of detection. The results from the simulated data show that a 50% probability of detection was achieved with a slightly higher threshold multiplier (lower P_{fa}), $T=0.48$. In fact, for this particular set of simulated data, a threshold multiplier of $T=0.44$ would have resulted in a probability of detection of 64%.

Figure 5-27. Maximum Threshold Multiplier, T , Required to Detect Target

Figure 5-28. Minimum Design P_{fa} Required to Detect Target

Thus, the theoretical detection ranges listed earlier are a reasonable prediction of the expected performance of the CP-140 air-to-air modes. The modes are summarized and the performance of the modes is reviewed in the next chapter.

CHAPTER 6: CONCLUSIONS

6.1 SUMMARY

This thesis describes the high-level design of the velocity search (VS), high PRF range-while-search (HPRF RWS), and medium PRF range-while-search (MPRF RWS) air-to-air modes for the CP-140. A summary of the burst structure of each mode was given in Section 3.9. A regional trimmed mean constant false alarm rate (TM-CFAR) technique was developed in Chapter 4. This technique makes use of *a priori* knowledge about the expected range-Doppler clutter maps in MPRF RWS mode of the CP-140 application. TM-CFAR was used in the HPRF modes. The predicted detection range performance of the CP-140 air-to-air modes was calculated in Chapter 5. Both single-burst and single-scan detection ranges were investigated. Range gate straddling and CFAR losses were built into the detection model to give a more accurate representation of true detection range performance. The predicted performance of the MPRF RWS mode was then further validated using simulated data.

Based on the theoretical values calculated in Chapter 5 to achieve a one-minute interval between false alarms, the detection ranges from Table 5-11 for a 5 m² target are repeated here in Table 6-1. For each mode, performance against a Swerling Case 0 (constant RCS) and a Swerling Case 1 (fluctuating, Rayleigh-distributed RCS) are given. Cumulative detection during a single scan is assumed.

Table 6-1 Detection Ranges for a One-Minute False Alarm Interval

	VS Mode		HPRF RWS Mode		MPRF RWS Mode	
	Case 0	Case 1	Case 0	Case 1	Case 0	Case 1
R_{50} (km)	113.5	115.3	122.0	96.2	59.4	56.4
R_{85} (km)	82.0	71.0	89.5	56.9	43.2	41.4
R_{90} (km)	74.4	61.3	80.9	44.7	39.6	37.0

In HPRF RWS mode, three frequency-modulated stages are transmitted during one time on target (TOT). Target correlation is required in all three stages in order to perform range resolving. In MPRF RWS mode, target correlation is required in at least three of the eight PRFs transmitted during one TOT for range and Doppler ambiguity resolving. With the current design of VS mode, however, a target is declared after detection on a single burst. Because no further correlation is required, the design probability of false alarm (P_{fa}) has to be much lower for VS mode than for the other modes in order to achieve the one-minute false alarm interval. As shown in Chapter 3, reducing the P_{fa} reduces the probability of detection (P_d) for a given scenario. This reduction in detection range could be alleviated by adding a track-before-detect scheme. If a new antenna with lower sidelobes is implemented on the CP-140, the MPRF RWS mode on the CP-140 should provide significantly longer detection ranges than those listed in Table 6-1.

6.2 FUTURE CONSIDERATIONS

There are additional design considerations for the CP-140 air-to-air modes that have not been addressed in this thesis. Notably absent from this baseline design and

performance prediction is a tracking algorithm for use in the TWS mode. Other considerations are listed below.

The duty cycle achieved with the highest PRF in MPRF mode is 26.5%. Because there is no specific MPRF range resolution driving the design for the CP-140 application, it would be possible to alter the design to increase the duty cycle to the maximum allowed by the transmitter. Using the same 13-bit Barker code, a compressed pulse width of 1.1 μs would result in an uncompressed width of 14.3 μs . This would widen the range gates slightly, resulting in a range gate width of 165 m rather than 150 m. The duty cycle for the highest PRF would become 29.2%, resulting in an increase in the average transmitted power of 2.7%. While this would result in only minor improvement in range, there are no design criteria precluding the use of the higher duty cycle.

The CFAR algorithm must be able to accommodate spurious lines in the range-Doppler spectrum that are not caused by clutter. These spurious lines may be generated internally from components in the transmit and receive chains, or may be due to external sources such as jet engine modulation. Unlike target or clutter signatures, these types of spurious lines are likely to appear in the same range-Doppler cell from burst to burst. One means to control the false alarms from these spurious lines is to use separate threshold multipliers on cells that declare repeated detections.

Fill pulses were added to each mode to ensure that the full coherent processing gain was achieved for targets at the maximum unambiguous detection ranges. As presented in this thesis, the design makes no use of the data received while the fill pulses are being transmitted. If adequate signal processing power is available, however, there is no reason to exclude these data from the processing. This would

result in the use of the much less efficient DFT because the number of points in the FFT will no longer be an integral power of two, but will increase the SIR of targets at shorter ranges.

Further testing of the regional TM-CFAR technique is critical once flight data are available. Specifically, both the use of regional TM-CFAR and the choice of each CFAR reference window should be validated with the flight data. The widths of each region should then be optimized using Monte Carlo techniques. In addition, the dependence of the technique's performance on factors such as altitude and terrain type should be investigated.

The MPRF RWS mode requires target correlation on at least three of eight PRFs to resolve ambiguities in range and Doppler. HPRF RWS requires target correlation on all three FM stages to resolve the target in Doppler. To accomplish this, range and Doppler resolving algorithms are required for both of these modes. A clustering algorithm developed by Trunk and Brockett has been shown to be superior to the Chinese Remainder Theorem that is commonly used in operational systems [36]. The clustering algorithm was further combined with a maximum likelihood technique by Trunk and Kim for the multiple target problem [38]. These algorithms are recommended for use in the CP-140 application.

The effect of the high sidelobes of the current AN/APS-506 antenna has been discussed with respect to the performance degradation of the MPRF mode. Little mention was made of jamming. Jamming waveforms can also corrupt the system more easily when there are high antenna sidelobes. The potential effect of jamming waveforms should be investigated along with potential electronic counter-countermeasures for the CP-140 application.

The antenna measurements taken in [10] included a set of measurements without the radome and lightning strips. As expected, the sidelobes are much lower for these measurements than for the measurements which included the radome and lightning strips. If the air-to-air modes use the current antenna (i.e., if the antenna is not to be replaced), then the effect of modifying the position of the lightning strips should be investigated. In this manner, it may be possible to reduce the sidelobe levels of the current antenna installation. As seen throughout this thesis, any reduction in sidelobes would greatly improve the performance of the MPRF RWS mode.

The AN/APS-506 does not currently have a guard antenna to isolate false alarms due to sidelobe clutter and jamming. If the antenna were to be replaced and if it were feasible to implement a guard antenna, sidelobe discretes could be rejected more effectively. Note, however, that simply using an antenna with lower sidelobes will allow better rejection of sidelobe discretes even with the simple sidelobe blanking technique described in the thesis. Algorithms are not included with this thesis to provide sidelobe false alarm blanking.

The maximum/minimum elevation angles for the CP-140 application are currently $+10^{\circ}$ / -15° . An increase in these angles would allow the radar to search a larger volume, and would alleviate restrictions on the ownship-target geometry necessary for detection.

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APPENDIX A. PROBABILITY OF DETECTION VS. RANGE CURVES

Figure B-1. VS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-2. VS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-3. VS Mode, Case 1 Target, One Minute False Alarm Interval

Figure B-4. VS Mode, Case 1 Target, One Minute False Alarm Interval

Figure B-5. HPRF RWS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-6. HPRF RWS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-7. HPRF RWS Mode, Case 1 Target, One Minute False Alarm Interval

Figure B-8. HPRF RWS Mode, Case 1 Target, One Minute False Alarm Interval

Figure B-9. MPRF RWS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-10. MPRF RWS Mode, Case 0 Target, One Minute False Alarm Interval

Figure B-11. MPRF RWS Mode, Case 1 Target, One Minute False Alarm Interval

Figure B-12. MPRF RWS Mode, Case 1 Target, One Minute False Alarm Interval