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System Analysis and RF-Floodlight Exploitation of Short-Range GOTCHA Repeaters

Jose A. Montes de Oca
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System Analysis and RF-Floodlight Exploitation of Short-Range GOTCHA Repeaters

A thesis submitted in partial fulfillment of the requirements for the degree of Master of Science in Engineering

by

Jose A. Montes de Oca
B.S.M.E., Massachusetts Institute of Technology, 1998.

2006
Wright State University
I HEREBY RECOMMEND THAT THE THESIS PREPARED UNDER MY SUPERVISION BY Jose A. Montes de Oca ENTITLED System Analysis and RF-Floodlight Exploitation of Short-Range GOTCHA Repeaters BE ACCEPTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF Master of Science in Engineering.

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ABSTRACT

Montes de Oca, Jose. M.S.E., Department of Electrical Engineering, Wright State University, 2006. System Analysis and RF-Floodlight Exploitation of Short-Range GOTCHA Repeaters.

Emerging wide-area synthetic aperture radar (SAR) system concepts call for a single data collection platform to orbit a large (e.g., 20-km) spot at a nominal range of 40-km. The large standoff distance and desire for fine resolution, coupled with a need for persistent real-time sensing, pose a significant challenge in terms of clutter-to-noise ratio (CNR) performance and data processing. Increased CNR and reduced processing load can be achieved by decreasing the range of the SAR system and the size of the area of interest. Employing multiple cooperating SAR systems allows the same overall coverage area to be maintained with a patchwork of SAR footprints. This paper analyzes a high-level system architecture, for multiple SAR systems, that provides uninterrupted coverage over a wide area. Bistatic receivers are also considered to collect through-wall signals and indirect path signals. Three different wall constructions are considered for the through-wall study. Three different exterior wall coverings are used in the multi-path study. System analysis includes eclipsing diagrams, CNR performance, mutual interference issues, through-wall, and multi-path modeling.
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Finally, I would like to thank my family and friends for their enthusiastic encouragement.
This is dedicated to my wife Cecilia.
Introduction

Recent events have brought into focus the need for persistent wide-area surveillance in urban military operations. The complex and heavily trafficked environment of an urban setting dramatically complicates surveillance and reconnaissance efforts. Buildings and other man-made structures are highly effective barriers for many sensing modalities (e.g., infrared, optical), and one must also allow for the presence of multiple moving and stationary targets. A loitering aerial asset operating at RF wavelengths may provide coverage of a region of interest by flooding that sector with sufficient RF radiation such that all targets of interest are at least intermittently illuminated. The GOTCHA system concept involves a high-value RF-ISR platform maintaining persistent coverage of a large area (e.g., a 20-km spot) from a nominally safe range (e.g., 40-km slant range) using waveforms with fine resolution in range and Doppler. Through innovative combinations of SAR and GMTI methodologies, moving and stationary targets will be continuously monitored for purposes of classification, recognition, identification, and tracking [1].

While the strategic advantages of such a system are readily apparent, developing this architecture poses significant technical challenges. In particular, the size of the area to be illuminated prevents use of convenient waveforms such as linearly frequency modulated (LFM) stretch [2]. The large standoff distance limits the rate at which information-rich, wide-angle data might be collected, and range also limits the strength of returns from low radar cross section targets. As an alternative, one may partition the large area into smaller sectors to be interrogated by a team of low-cost repeaters that cooperate with the high-value ISR platform but operate at much shorter range. seen in Fig. 1.1 shows the smaller systems surveying the same area as the larger SAR system. By operating at shorter ranges,
the repeaters will observe smaller spot sizes and will be able to form fine-resolution SAR images with shorter coherent integration intervals. More importantly, they will be capable of quickly collecting wide-angle data for high-quality imaging, tracking, and identification.

Various studies have reported on designs of airborne SAR systems. Damini describes the testing of a single multi-mode X-band system mounted on a Convair 580 aircraft [3]. Similarly, Horn describes an airborne SAR system that uses a Dornier DO 228 aircraft, and it has the capability of using various RF-bands [4]. Dreuillet provides an update regarding a SAR system designed for foliage penetration (FOPEN), MTI, and airborne bistatic mode, and the two platforms considered are larger aircraft, such as the C-130, and a UAV [5]. The Dreuillet paper indicates that they tested their airborne bistatic setup with a 2-km distance between transmitter and receiver. With the exception of this last study, the other airborne radar systems studies were implemented on single-aircraft, non-UAV platforms. These papers demonstrated the successful testing of airborne SAR systems, but did not involve multiple transmitters, in which we are interested.

This thesis presents a study of the viability of wide-area SAR surveillance with a contiguous patchwork of observation footprints and the use of bistatic non-airborne receivers to collect wide angle data. The first section begins with the analysis of a single monostatic SAR system. The equation used to estimate the two-way propagation of an electromagnetic
wave \[6, 7, 8, 9\] is used to create an eclipsing diagram. The eclipsing diagram is used to trade-off between choice of pulse repetition frequency (PRF) and range for a given pulse width. Next, a link budget analysis is performed using the clutter-to-noise ratio (CNR) equation \[6, 7, 8, 9\] to determine appropriate system parameters for the given mode of operation. Multiple sources are used to find reasonable parameters for use within the CNR equation \[8, 10, 11\]. A second SAR sensor, surveying an adjacent area, is then added to the scenario. The interaction between the two systems is studied in terms of mutual eclipsing and clutter-to-interference plus noise ratios (CINR). Coded waveforms are considered in place of the traditional LFM pulse in order to mitigate some losses due to mutual interference.

The next chapter presents the results of a bistatic receiver model. The X-Band airborne transmitter power is now collected by a receiver located within the illumination scene. Two different scenarios are considered during the bistatic receiver study. The first has the target located behind a single wall. The small wavelength of the X-band system is attenuated dramatically, so having the bistatic receiver at a smaller range than the transmitter allows it to capture the faint through-wall signal. However, since the receiver is located within the region of illumination it must have a low side-lobe ratio to reduce direct path interference. Three different wall constructions are explored, and the signal-to-noise ratio (SNR) results are presented. Wall loss measurements are used to modify the SNR equation \[12\]. The second bistatic receiver scenario involves a multi-path analysis. The bistatic receiver is used to engage a target via an indirect path. Although the target is not located behind a wall, the transmitter does not have a line of sight (LOS) to the target. The radar signal reaches the target after multiple reflections off of external walls. Again, different wall materials are explored \[12, 13, 14\], and SNR values are presented.
2.1 Monostatic SAR System Analysis

To establish a baseline for performance, we first consider the case of a single monostatic SAR observing an area with a diameter of 5 km at a range of 20 km. The system is assumed to employ a LFM waveform capable of supporting one-foot range resolution (i.e., \( B \approx 500 \) MHz bandwidth) with a 40-\( \mu s \) pulse width and an X-band RF frequency. The sensor should unambiguously capture the Doppler bandwidth of the illuminated spot, and the return from the lowest clutter reflectivity in the scene should have a peak CNR of 0 dB.

2.1.1 Eclipsing

At a nominal range of 20 km and with a pulse width of 40 \( \mu s \), the PRF \( (F_p) \) of the radar must be chosen to not only support the expected Doppler bandwidth of the scene but it must also prevent eclipsing of returns. Eclipsing occurs when pulses reflected from the scene of interest are received by the system at the same time that a new pulse is being transmitted. In most current systems, sufficient energy leaks from the transmitter into the receive chain to saturate the receiver and blind the system to any returns at that time. Selecting an appropriate PRF will prevent such overlap of outgoing and incoming pulses [6].

Before considering eclipsing, one must first determine the minimum PRF required to unambiguously sample the band of Doppler frequencies expected from the scene. This requirement is derived based on exceeding the Nyquist sampling rate for the scene’s aggregate Doppler response. Figure 2.1 is used to derive the Doppler bandwidth of the stationary
objects within a SAR scene [7]. Figure 2.1 shows that $\phi_s$ is the squint angle of the antenna beam relative to the aircraft velocity vector, and $\Delta \theta$ is the azimuth width of the antenna beam. Point 1 is located at the leading edge of the SAR beam, and it has an angle of $\phi_s - \frac{\Delta \theta}{2}$. Similarly, Point 2 is located at the trailing edge of the SAR beam, and it has an angle of $\phi_s + \frac{\Delta \theta}{2}$. The Doppler shift at Points 1 and 2 is defined as

$$f_{d1} = \frac{2v_a}{\lambda} \cos \left( \phi_s - \frac{\Delta \theta}{2} \right)$$

(2.1)

$$f_{d2} = \frac{2v_a}{\lambda} \cos \left( \phi_s + \frac{\Delta \theta}{2} \right)$$

(2.2)

The difference between (2.1) and (2.2) is then taken as

$$B_d = f_{d1} - f_{d2} = \frac{2v_a}{\lambda} \left\{ \cos \left( \phi_s - \frac{\Delta \theta}{2} \right) - \cos \left( \phi_s + \frac{\Delta \theta}{2} \right) \right\}$$

(2.3)

The following trigonometry identity is used to simplify (2.3):

$$\cos(A) - \cos(B) = -2 \sin \left( \frac{A + B}{2} \right) \sin \left( \frac{A - B}{2} \right)$$

(2.4)

The Doppler bandwidth is simplified to

$$B_d = \frac{4v_a}{\lambda} \sin \phi_s \sin \frac{\Delta \theta}{2}.$$  

(2.5)

Assuming an aircraft velocity of $v_a = 150$ m/s and X-band operation ($\lambda = 3$ cm), the Doppler bandwidth is $B_d = 2.48$ kHz for the case of broadside observation ($\phi_s = 90^\circ$).
Figure 2.2: Outgoing pulses eclipse incoming target returns.

of a 5-km spot at a range of $R_a = 20$ km ($\Delta \theta = 14.25^\circ$). To satisfy Nyquist for complex sampling, the PRF must satisfy the condition $F_p > B_d$.

For a given PRF and pulse width ($\tau$), we now seek to determine what target ranges will be eclipsed by transmissions of subsequent pulses. The electromagnetic waves emitted by the radar system travel at the speed of light, $c = 3 \times 10^8$ m/s. They reach the target at a certain distance or range (R) and return during an elapsed time ($\Delta t$):

$$\Delta t = \frac{2R}{c}$$  \hspace{1cm} (2.6)

Assuming that the first pulse is transmitted at time $t = 0$ seconds, the $k$th subsequent pulse begins transmission at time $t_k = k/F_p = kT_p$ – where $T_p$ is the inter-pulse period – and ends transmission $\tau$ seconds later. Therefore, target returns that overlap the time interval $[t_k - \tau, t_k + \tau]$ will be at least partially eclipsed by the $k$th pulse. The return from a target at range $R_k$ falls in that interval if

$$t_k - \tau \leq \frac{2R_k}{c} \leq t_k + \tau.$$  \hspace{1cm} (2.7)

Figure 2.2 illustrates that this interval includes the overlap of the leading edge of returning pulses with the trailing edge of outgoing pulses and vice versa. Equation (2.7) thus implies that the ranges eclipsed by the $k$th pulse for a given PRF and pulse width are

$$\frac{c}{2} (kT_p - \tau) \leq R_k \leq \frac{c}{2} (kT_p + \tau).$$  \hspace{1cm} (2.8)
The interval of eclipsed ranges (2.8) can be computed for \( k = 0, 1, 2, 3, \ldots, n \) and for a range of pulse repetition frequencies. Eclipsing as a function of range and PRF may then be represented as shown in Fig. 2.3, where the shaded bands indicate combinations of range and PRF that result in eclipsing. The vertical band corresponds to very short ranges which are eclipsed because the entire pulse has not left the transceiver \[8\]. The open region of operation directly to the right of this band is commonly referred to as range unambiguous operation, and subsequent open bands of operation are said to be range ambiguous. The horizontal line centered at 20 km indicates that operation at a PRF of \( F_p = 4 \text{ kHz} \) will insure that the entire 5-km range swath interest will be observable without eclipsing.

### 2.1.2 Clutter-to-Noise Ratio

Having defined the key system requirements for the monostatic SAR, we may now continue with a radar range equation analysis, therein seeking to determine a set of system parameters that will provide the desired nominal system performance. The CNR of the system will be predicted using \[7, 10, 9\]

\[
CNR = \frac{P_{ave} G_t^2}{(4\pi)^3 R^4} \cdot A_{res} \cdot \sigma_0 \cdot \lambda^2 \cdot \frac{1}{L} \cdot \frac{1}{kT_o F_n} \cdot T_{coh} \tag{2.9}
\]
where \( P_{ave} \) is the average transmitted power, \( G_t \) is the antenna gain, \( R \) is the range to the target, \( A_{res} \) is the ground resolution area, \( \sigma_0 \) is the groundback scatter coefficient, \( L \) is the system losses, \( k \) is the Boltzmann’s constant \([1.38 \times 10^{-23} \text{ W-sec/K}]\), \( T_o \) is the receive temperature \([290 \text{ K}]\), \( F_n \) is the receiver noise figure, and \( T_{coh} \) is the coherent integration time. Note that (2.9) assumes that the sampling frequency \((F_S)\) is equal to the receiver bandwidth \((B)\). The parameters contained in (2.9) will be adjusted to meet the aforementioned system requirements and to deliver a peak CNR of 0 dB. Many of the parameters in (2.9) are predetermined by system requirements. For example, the range has already been chosen to be \( R = 20 \text{ km} \), and producing square resolution cells will dictate \( A_{res} = 1 \text{ ft} \times 1 \text{ ft} = 0.093 \text{ m}^2 \). The crossrange resolution is related to the coherent integration time via [7]

\[
\Delta_y = \frac{\lambda R}{2v_a T_{coh}} \rightarrow T_{coh} = \frac{\lambda R}{2v_a \Delta y},
\]

which indicates an integration interval of \( T_{coh} = 6.56 \text{ seconds} \) in order to provide 1-foot azimuth resolution.

The transmit and receive gain at the bore sight depends upon the effective antenna aperture size, and it is calculated as [7, 8, 10, 9],

\[
G_t = \frac{4\pi A_e}{\lambda^2}.
\]

The effective aperture size is largely dictated by the spot size to be observed. Assuming that the antenna (3-dB) beam width in radians for a given dimension is well approximated by [7]

\[
\Delta \theta = 0.88 \frac{\lambda}{D}.
\]

A horizontal aperture dimension of 10.6 cm is required to provide a 14.25° azimuth beam. With a slant range of \( R = 20 \text{ km} \) and an assumed aircraft altitude of 10 km, an elevational beam width of 7.2° is necessary to cover the 5-km ground swath. This then indicates a vertical aperture dimension of 21 cm, thus giving a total effective aperture area of 222.6 cm\(^2\) and thus a peak gain of approximately \( G_t = 25 \text{ dB} \). For the remainder of the scene, the main lobe of the beam was modeled as a two-dimensional Gaussian with specified 3-dB widths:

\[
G_t = \frac{4\pi lh}{\lambda^2} \exp \left( \frac{4}{\Delta \theta_A^2} \log_{10}(0.5) \cdot \theta_A^2 \right) \cdot \exp \left( \frac{4}{\Delta \theta_E^2} \log_{10}(0.5) \cdot \theta_E^2 \right) \]

8
where \( l \) is the horizontal aperture dimension (10.6 cm), \( h \) is the vertical aperture dimension (21 cm), \( \Delta \theta_A = \frac{0.88 \lambda}{l} \) is the 3-dB azimuth beam width, \( \theta_A \) is the azimuth angle, \( \Delta \theta_E = \frac{0.88 \lambda}{h} \) is the 3-dB elevation beam width, and \( \theta_E \) is the elevation angle. Figure 2.4 is used to illustrate the two vectors that are used to compute \( \theta_A \) and \( \theta_E \) in (2.13). The bore sight antenna direction is defined as

\[
\vec{v}_b = (x_t - x_b, y_t - y_b, z_t - z_b),
\]

and the vector to a point \((x_k, y_k, z_k)\) in the scene is defined as

\[
\vec{v}_k = (x_t - x_k, y_t - y_k, z_t - z_k).
\]
The azimuth angle ($\theta_A$) to any point in the scene is calculated as

$$\theta_A = \cos^{-1}\left(\frac{\overrightarrow{v_{bA}} \cdot \overrightarrow{v_{kA}}}{\|\overrightarrow{v_{bA}}\|\|\overrightarrow{v_{kA}}\|}\right)$$

$$= \cos^{-1}\left\{\frac{(x_t - x_b, y_t - y_b) \cdot (x_t - x_k, y_t - y_k)}{\|(x_t - x_b, y_t - y_b)\|\|(x_t - x_k, y_t - y_k)\|}\right\}$$

(2.16)

where $\overrightarrow{v_{bA}}^T$ is transpose vector of $\overrightarrow{v_{b}}$ projected onto the horizontal ground plane $(x_t - x_b, y_t - y_b)$ and $\overrightarrow{v_{kA}}$ is vector $\overrightarrow{v_{k}}$ projected onto the horizontal ground plane $(x_t - x_k, y_t - y_k)$.

Figure 2.5 has the region of interest with $\theta_A$ contours at $1^\circ$ increments. These angles are defined with respect to the bore sight marked—shown with an asterisk at 1.75 km x 2.5 km. Since the monostatic radar system is located to the right of this area the constant angle contours are nearly constant along the range dimension. Similarly, the elevation angle ($\theta_E$) is calculated as

$$\theta_E = \cos^{-1}\left(\frac{\overrightarrow{v_{bE}}^T \overrightarrow{v_{kE}}}{\|\overrightarrow{v_{bE}}\|\|\overrightarrow{v_{kE}}\|}\right)$$

$$= \cos^{-1}\left\{\frac{\left(\sqrt{(x_t - x_b)^2 + (y_t - y_b)^2}, (z_t - z_b)\right)}{\left\|\sqrt{(x_t - x_b)^2 + (y_t - y_b)^2}, (z_t - z_b)\right\|}\right\}$$

$$\ldots$$

(2.17)

$$\ldots$$

where $\overrightarrow{v_{bE}}^T$ is transpose vector of $\overrightarrow{v_{b}}$ projected onto the vertical plane $\sqrt{(x_t - x_b)^2 + (y_t - y_b)^2}$.
Figure 2.6: Elevation angle.

\((z_t - z_b)\) and \(\vec{v}_{kE}\) is vector \(\vec{v}_k\) projected onto the vertical plane \(\sqrt{(x_t - x_k)^2 + (y_t - y_k)^2}\), 
\((z_t - z_k)\). Figure 2.6 has the region of interest with \(\theta_E\) contours at 1° increments. These angles are defined with respect to the bore sight marked—shown with an with the asterisks at 1.75 km x 2.5 km. Since the monostatic radar system is located to the right of this area

Figure 2.7: Two way antenna gain.
the constant angle contours are nearly constant along the cross-range dimension. Figure 2.7 has the results of (2.13) squared to provide the two-way antenna gain that is used in (2.9). The two-way gain at the bore sight is 49.8 dB, and most of the region of interest has a two-way antenna gain of 43 dB.

Two angles are used to define the groundback scatter coefficient ($\sigma_0$). Figure 2.8 shows that the angle of incidence is defined as

$$\theta_I = \sin^{-1}\left(\frac{Z_t}{R}\right)$$

(2.18)

where $Z_t$ is the altitude of the transmitter and $R$ is the range of the transmitter. Note that in monostatic systems the incident angle equals the scatter angle ($\theta_S$) because the transmitter and receiver are located on the same platform. Figure 2.9 has contours for $\theta_I$. As expected the angles are constant along the cross-range direction, and they vary between $28^\circ$ to $35^\circ$ along the range. Figure 2.8 also shows that the monostatic angle is defined as

$$\psi_m = \cos^{-1}\left(\frac{-v_x^2 - v_y^2 + v_z^2}{R^2}\right)$$

(2.19)

where $v_x$, $v_y$, and $v_z$ are the incoming vector directions. Note that the dot product has been taken between the incoming vector and the same vector but with negative x and y
Figure 2.9: Angle of incidence ($\theta_I = \theta_S$).

Figure 2.10: Monostatic angle ($\psi_m$).
components. The angle contours are shown in Fig. 2.10. The contours are constant along the cross-range direction, and they vary between 110° to 124° along the range. If the monostatic angle is added to twice the incident angle those angles equal 180°.

The monostatic back scatter coefficient combines two equations found in [10], and it is defined as

\[ \sigma_o = \Gamma \sin \theta_i \cdot \left( 0.003 + \exp \left( \frac{-\psi_m^2}{0.0289} \right) \right) \]  

(2.20)

where \( \Gamma \) is the normalized reflectivity parameter which was set equal to 0.6 dB. The radar cross section value of about \( \sigma_0 = -28 \) dB was computed throughout the region of interest. This value corresponds to the minimum reflectivity observed in an urban setting [11] (e.g., pavement, asphalt). It also matches well with the numbers reported in [10].

The remaining parameters in (2.9) were set based on typical system values [8, 15], and are listed along with a summary of the other parameters in Table 2.1. The average transmit power was chosen last. It was adjusted in order to give a peak image CNR of 0 dB. Figure 2.11 shows the predicted scene CNR as a function of range and azimuth. Contour lines are drawn in 1-dB increments. We also see that most of the region of interest has a CNR between -3 dB and 0 dB.

![Figure 2.11: Predicted CNR performance for an isolated monostatic SAR.](image-url)
Table 2.1: Parameters used in the baseline monostatic SAR CNR analysis (see Fig. 2.11)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
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<td>$\phi_s$</td>
<td>90°</td>
<td>$\Delta\theta$</td>
<td>14.25°</td>
</tr>
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<td>$A_e$</td>
<td>10.6 cm × 21 cm</td>
<td>$R$</td>
<td>20 km</td>
<td>$\sigma_0$</td>
<td>-28 dB</td>
</tr>
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<td>$T_{coh}$</td>
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<td>$F_n$</td>
<td>3 dB</td>
<td>$G_t$</td>
<td>25 dB</td>
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<td>$L$</td>
<td>5 dB</td>
<td>$k$</td>
<td>$1.38 \times 10^{-23}$ W-s/K</td>
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<td>3 cm</td>
<td>$T_o$</td>
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<tr>
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<td>$\tau$</td>
<td>40 µs</td>
<td>$F_P$</td>
<td>4 kHz</td>
</tr>
</tbody>
</table>

2.2 Analysis of Two Cooperating Monostatic SAR Systems

The first contribution of this thesis is a study of the viability of deploying multiple monostatic SAR sensors to cooperatively observe a large region on the ground. The previous section established a baseline system model upon which we can build further analysis. Here, deployment of a second SAR sensor with a footprint directly adjacent to that of the first SAR system will be considered. The second sensor is assumed to be identical to the first in terms of performance and operation. The models developed in Section 2.1 must therefore be modified to include the interaction between the two radars. Their direct path interference complicates the eclipsing diagrams, and their ground path interference affects the CNR analysis.

2.2.1 Eclipsing

For simplicity, it is assumed that the two SAR systems employ identical waveforms and will therefore interfere coherently with one another. The strongest component of this interference will potentially be due to direct path propagation. It is expected that the direct path interference observed by each platform will be sufficient to saturate their receivers or to at least degrade performance to an unacceptable level. The two cooperating systems should consequently synchronize their transmissions in order to minimize or eliminate this
degradation. The eclipsing due to a second transmitter at some range $R_0$ from the first may be incorporated into the previous eclipsing diagram by adding another set of eclipsed regions, derived with an offset equal to the direct-path time delay between the two platforms. For example, if the two aircraft are separated by 5-km, the time offset would be $t_0 = 5 \text{ km}/(3 \times 10^8 \text{ m/s}) = 16.7 \mu \text{s}$. Including the time offset in equation (2.8) to yield

$$\frac{c}{2}(kT_p - \tau + t_0) \leq R_k \leq \frac{c}{2}(kT_p + \tau + t_0),$$

(2.21)

simply adds a shifted copy of the original eclipsing bands, as shown in Fig. 2.12. For the current system model, which includes a $\tau = 40 \mu \text{s}$ pulse width and a $T_p = 250 \mu \text{s}$ interpulse period, an additional 16.7 $\mu \text{s}$ of eclipsing is negligible, as there are still significant regions of PRF-versus-range trade space in which to operate. Indeed, as the subsequent radar range equation analysis will show, choosing to place the two platforms 5 km apart, such that they fly parallel flight paths on the same side of the area of observation, will be a beneficial decision from a multipath interference perspective as well. However, in the case that more than two SAR sensors are cooperating, there would be additional eclipsing bands added to each platform’s trade space, and selection of an acceptable mode of operation would become more difficult.
2.2.2 Clutter-to-Interference Plus Noise

In the CNR analysis of the individual monostatic SAR sensor, the data collection platform was assumed to fly a northbound trajectory centered due east of the map center with an altitude of 10 km and a slant range to scene center of 20 km. In order to image an adjacent 5-km spot, a second platform is now assumed to be flying an identical trajectory 5 km due south of the first aircraft. The second SAR system, possessing identical system parameters, observes an equal sized region contiguous with the southern edge of the first region. The platforms are synchronized to transmit simultaneously, such that an operating point within the range-PRF trade space that is free of direct path interference and eclipsing can be selected. Having both systems transmitting in the same westerly direction will help to minimize the amount of backscatter interference energy observed due to overlap of antenna beams.

As such, the radar range equation analysis from the previous section must only be modified to include multi-path interference, which will be dominated by the overlap of the two systems’ beams on the ground. Including this new term yields the clutter-to-interference plus noise (CINR) model

\[
\text{CINR} = \frac{P_{\text{ave}} G^2}{(4\pi)^3 R^4} \cdot \frac{A_{\text{res}} \cdot \sigma_0 \cdot \lambda^2}{L} \cdot \frac{T_{\text{coh}} \cdot F_s}{kT_0 F_n B + \sum_{x,y} I_{MP}(x, y) \tau F_s N_{RG}}
\]

where the received power of the multi-path interference signal is represented by \( \sum_{x,y} I_{MP}(x, y) \). This is the sum of the interference returns from all of the illuminated resolution cells in both SAR scenes. The multi-path term also experiences a coherent integration gain of \( \tau F_s \), under the assumption that the two transceivers are employing identical waveforms and thus are coherent with one another. Pulse compression also serves to distribute the multi-path interference energy across range bins \( N_{RG} = 5 \text{ km}/0.304 \text{ m} = 16447 \text{ gates} \). Furthermore, it is assumed that the interference terms will not be coherent on a pulse-to-pulse basis, and thus do not realize a full \( T_{\text{coh}} F_s \) coherent integration gain, due to mismatches in Doppler processing.

All of the other terms in (2.22) were described in the single monostatic analysis except
Figure 2.13: (a) Transmitter 1 two-way system gain, (b) Transmitter 2 two-way system gain, and (c) Combined Transmitter 1 and Transmitter 2 system gain

for the multi-path interference term. The power of the interference signal is modeled as

\[
I_{MP}(x, y) = \frac{P_t G_2(x, y)}{4\pi R_2(x, y)^2} \cdot A_{res} \cdot \sigma_b(x, y) \cdot \frac{G_1(x, y)}{4\pi R_1(x, y)^2} \cdot \frac{\lambda^2}{4\pi L_{MP}}
\] (2.23)

where \( P_t = P_{avg} T_p/\tau \) is the peak transmitted power. The interference power is computed for each resolution cell (or pixel) in the two scenes, such that \( R_1(x, y) \) and \( R_2(x, y) \) are the ranges to each cell from the first and second platforms, respectively. Similarly, \( G_1(x, y) \) and \( G_2(x, y) \) are the antenna gains experienced by each pixel due to the first and second systems. Figure 2.13 has various two-way gain values. Figure 2.13a has the contours from transmitter 1. Figure 2.13b has the same contours but from platform 2. Figure 2.13c has the combined gain values from both transmitters. The antenna beams were modeled identically as two-dimensional Gaussian functions with the earlier specified 3-dB widths. The parameter \( L_{MP} = 5 \) dB is a miscellaneous multi-path loss factor.

A bistatic scattering model is required to represent the multi-path reflectivity \( \sigma_b(x, y) \) of each resolution cell. Using the angular quantities illustrated in Fig. 2.14, the ground backscatter coefficient is calculated as

\[
\sigma_b(x, y) = \Gamma \sqrt{\sin \theta_f(x, y) \sin \theta_s(x, y)} \cdot \left( 0.003 + \exp \left( \frac{-\psi(x, y)^2}{0.0289} \right) \right)
\] (2.24)
where $\Gamma = 20 \text{ dB}$ is an estimated normalized reflectivity parameter (based on [7] corresponding to urban terrain), $\theta_I$ is the incident angle, $\theta_S$ is the scatter angle, and $\psi$ is the angle between the forward scatter direction vector and the unit vector pointing from the pixel to the second platform. Figure 2.15a and 2.15.b shows the contours for the incident angles and scatter angles respectively. Both transmitters are located east of the region of interest, with transmitter 1 to the north of transmitter 2. The forward scatter angle corresponding to a given point in the scene is defined as

$$
\psi = \cos^{-1} \left( \frac{-v_{1x}v_{2x} - v_{1y}v_{2y} + v_{1z}v_{2z}}{R_1 R_2} \right)
$$

where subscript 1 corresponds to the vector component from transmitter 1 and subscript 2 corresponds to the vector component from transmitter 2. The angle contours for $\psi$ are shown in Fig. 2.15b. The $\psi$ angles vary from $104^o$ to $124^o$. Note $\psi$ is equal to zero in the case of forward scatter such that mutual interference is maximized when the two sensors view the scene from opposite sides. This expression is an amalgam of models [10] meant to estimate scattering coefficients for forward and backscatter cases. The $\sigma_b$ values vary between -9 dB and -7 dB. This was also expected because of the higher normalized coefficient used here which represents more reflective surfaces (e.g., buildings, vehicles, natural clutter).

Once the interference signal power $I_{MP}(x,y)$ from each resolution cell is calculated, they are summed and multiplied by $\tau$ and $F_S$ then divided by the number of range bins to yield the interference contribution to the noise term. Figure 2.16 shows the multi-path
interference contours which vary between -160 dB and -145 dB. The largest values are observed within the overlap region. This straightforward model yielded a peak CINR of -12.6 dB when all of the previous parameters were kept the same. The predicted CINR for one of the two systems is shown in the top half of Fig. 2.17. Figure 2.17 shows that the 5 km x 5 km area monitored by the first transmitter will have a CINR that ranges between -12.6
Figure 2.17: Predicted CINR for a monostatic SAR cooperating with an adjacent system (10.6 cm x 21 cm antenna).

dB and -18 dB. The reduced performance between the single system and this dual system scenario is due to the overlap between the two antenna beams. This overlap increases the $I_{MP}(x, y)$ which reduces the overall CINR values. Equation 2.12 indicates that increasing the antenna width would decrease the beam width. Increasing the antenna width from 10.6 cm to 21 cm (making antenna square 21 cm x 21 cm) increases the maximum CINR to 0.1 dB. Figure 2.18 shows that the reduced beam width decreases the -3 dB cross range coverage to 2.5 km. The second system would image the bottom half of the same figure.
2.3 Coded Waveforms

Use of two identical SAR systems to map adjacent regions led to a drastic degradation in system performance. This was driven by an assumed coherency between their transmitted pulses, under the assumption that identical LFM waveforms would be employed. As an alternative, to reduce the effect of the multi-path interference contribution, a different coded waveform [10, 9] could be employed by each SAR platform. As the large range swath already prevents use of stretch-LFM processing, the entire signal bandwidth must already be captured by the receiver, and a change to a coded waveform would not affect this requirement.

Use of a different coded waveform by each transceiver would eliminate the assumed coherency seen in pulse compression and would alter the CINR model to be

\[
CINR = \frac{P_{ave} G_t^2}{(4\pi)^2 R^4} \cdot A_{res} \cdot \sigma_0 \cdot \lambda^2 \cdot \frac{1}{L} \cdot \frac{1}{kT_0 F_n + \Sigma I_{MP}/F_s} \cdot T_{coh} \tag{2.26}
\]

where the pulse compression gain of \(\tau F_s\) has been removed from the interference term. The predicted CINR performance for coded waveforms is shown in Fig. 2.19, which again
Figure 2.19: Predicted CINR for a monostatic SAR cooperating with an adjacent system with coded waveform (10.6 cm x 21 cm antenna).

illustrates the performance of only one of the two platforms.

To completely eliminate the effects of mutual interference, orthogonally frequency modulated waveforms could be employed to place the two platform transceivers out of band with respect to one another [16, 17]. This should theoretically nullify the interference term $\Sigma I_{MP}/F_a$ in (2.26), thereby restoring performance to the baseline level, shown in Fig. 2.11 for a single isolated SAR sensor.
System-Level Architecture for Bistatic Receivers

The deployment of close-in bistatic receivers can be used to better engage obscured targets and to more quickly collect wide-angle data. The goal of this section is to once again solve for the signal-to-noise ratio of the receiver where it is not collocated with the transmitter. We will also limit the study to a single transmitter and receiver setup.

3.1 Analysis of Bistatic Receiver

Figure 3.1 shows one potential scenario where the receiver is immersed in the region of interest. The airborne transmitter ($T_x$) is illuminating both the receiver ($R_x$) and the target ($T$). The receiver is located at a range of $R_L$ from the transmitter. The transmitter-to-target range is abbreviated $R_T$, and the receiver-to-target range is $R_r$. All three range values are equal to the Euclidean distance between the points of interest. The electromagnetic energy emitted by the transmitter will act like a side-lobe jammer to the bistatic receiver. Hence, this inference is captured by a direct path interference term, and it is defined as

$$I_{DP} = \frac{P_{ave} G_{Tx} A_{RXL}}{4\pi R_L^2} T_{coh} F_s$$

where $P_{ave}$ is the average transmitter power, $G_{Tx}$ is the transmitter gain, $A_{RXL}$ is the effective area as seen from the side-lobe direction, $L$ is the overall system loss (assumed to be equal to 3 dB), $T_{coh}$ is the coherent time, and $F_s$ is the sampling frequency. The average transmitter power is maintained at 91.2 W and the transmitter’s 10.6 cm x 21 cm antenna...
Figure 3.1: Bistatic exploitation of monostatic repeaters.

has a main-lobe gain of 27.9 dB. The overall system loss is 5 dB. The sampling frequency is 493 MHz.

Equation 2.11 is used to define the relationship between $A_{RxSL}$, the side-lobe gain $G_{RxSL}$, in dB units, and the wavelength ($\lambda$)

$$G_{RxSL} = \frac{4\pi A_{RxSL}}{\lambda^2} \rightarrow A_{RxSL} = \frac{G_{RxSL} \lambda^2}{4\pi}.$$  \hspace{1cm} (3.2)

Also, the side-lobe gain is typically defined as an attenuation of the main-lobe gain ($G_{Rx}$),

$$G_{RxSL} = G_{Rx} \cdot 10^{G_{SL}/10},$$  \hspace{1cm} (3.3)

and the receiver antenna main-lobe gain is

$$G_{Rx} = \frac{4\pi A_{Rx}}{\lambda^2}.$$  \hspace{1cm} (3.4)

where $A_{Rx}$ is the receiver antenna area. Let us now find the signal power that bounces off the target and enters the receiver. The signal power reaching the target from the transmitter is defined as

$$S_{Tgt} = \frac{P_{Tx} G_{Tx}}{4\pi R_{T}^2 L_{Wall}},$$  \hspace{1cm} (3.5)
Table 3.1: Loss through material (measured at approximately 10 GHz, * taken at 7.01 GHz).

<table>
<thead>
<tr>
<th>Material</th>
<th>Size [cm x cm x cm]</th>
<th>Loss [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wallboard</td>
<td>121.8 x 196.9 x 1.17</td>
<td>0.53</td>
</tr>
<tr>
<td>Cloth Partition</td>
<td>140.7 x 153.1 x 5.93</td>
<td>6.07</td>
</tr>
<tr>
<td>Structure Wood</td>
<td>121.5 x 197.8 x 2.07</td>
<td>2.07</td>
</tr>
<tr>
<td>Plywood</td>
<td>121.9 x 197.51 x 1.52</td>
<td>2.76</td>
</tr>
<tr>
<td>Styrofoam</td>
<td>121.8 x 197.7 x 9.91</td>
<td>0</td>
</tr>
<tr>
<td>Brick (single)</td>
<td>8.7 x 19.8 x 5.83</td>
<td>4.48*</td>
</tr>
<tr>
<td>Concrete Block</td>
<td>19.45 x 39.7 x 19.5</td>
<td>13.62</td>
</tr>
</tbody>
</table>

where the wall attenuates the electromagnetic energy by $L_{Wall}$. The loss through various materials commonly used in buildings has been studied [12]. Table 3.1 lists the various loss values measured at approximately our carrier frequency. The slant range to the target is maintained at 20 km. Next, the signal bounces off the target and reaches the receiver with a signal power of

$$S_{Rx} = \frac{S_{Tgt}G_{Rx}\lambda^2\sigma}{(4\pi)^2R_t^2L_{Wall}} = \frac{P_{Tx}G_{Tx}G_{Rx}\lambda^2\sigma}{(4\pi)^3R_t^2R_r^2L_{Wall}^2}$$  \hspace{1cm} (3.6)$$

Since we are interested in the signal-to-interference plus noise ratio (SINR), we can now use (3.1), (3.8), and the thermal noise to find the SINR:

$$SINR = \frac{S_{Rx}}{L \cdot (kT_0F_nB + I_{DP})} T_{coh} F_s \cdot d$$ \hspace{1cm} (3.7)$$

$$SINR = \frac{P_{Tx}G_{Tx}G_{Rx}\lambda^2\sigma}{(4\pi)^3R_t^2R_r^2L_{Wall}^2L \cdot (kT_0F_nB + I_{DP})} T_{coh} F_s \cdot d.$$  \hspace{1cm} (3.8)$$

where the duty cycle ($d$) is between 0 and 1 and $T_{coh}$ is the coherent integration time. The bistatic receiver must use an appropriate $T_{coh}$ in order to achieve the Doppler resolution ($f_d$) needed to resolve the target of interest:

$$\frac{1}{T_{coh}} = f_d = \frac{2v}{\lambda}.$$  \hspace{1cm} (3.9)$$

where $v$ is the velocity of the target. The target is assumed to be moving at about 2.1 m/s (13 mph). Given a wavelength of 3 cm this velocity translates to a $T_{coh}$ of 7.3 ms.
Table 3.2: Parameters used in the baseline bistatic SNR analysis (see Fig. 3.2)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{Tx}$</td>
<td>570 W</td>
<td>$\tau$</td>
<td>40 $\mu$s</td>
<td>$F_p$</td>
<td>4 kHz</td>
</tr>
<tr>
<td>$P_{ave}$</td>
<td>91.2 W</td>
<td>$A_{Tx} = A_{Rx}$</td>
<td>21 cm $\times$ 21 cm</td>
<td>$\lambda$</td>
<td>3 cm</td>
</tr>
<tr>
<td>$G_{Tx} = G_{Rx}$</td>
<td>27.9 dB</td>
<td>$L$</td>
<td>5 dB</td>
<td>$B = F_S$</td>
<td>493 MHz</td>
</tr>
<tr>
<td>$R_T$</td>
<td>20 km</td>
<td>$\sigma$</td>
<td>1 m$^2$</td>
<td>$k$</td>
<td>1.38 $\times$ 10$^{-23}$ W-s/K</td>
</tr>
<tr>
<td>$F_n$</td>
<td>3 dB</td>
<td>$v$</td>
<td>2.1 m/s</td>
<td>$T_{coh}$</td>
<td>7.3 ms</td>
</tr>
</tbody>
</table>

Table 3.2 has the various parameters used in the bistatic analysis. The target is located at the origin of our model and at a slant range of 20 km from the transmitter. Figure 3.2 has the results of our first scenario where the receiver is located at a range of 200 m from the target ($R_r = 200$ m). The side-lobe gain $G_{RxSL}$ must be under -70 dB for the receiver to have a positive SNR. Three different wall constructions are considered: 1) plywood and wall board, 2) brick, plywood, and wall board, and 2) concrete block, plywood, and wall board. Figure 3.2 shows that with a $G_{RxSL}$ less than -70 dB the signal reflected from the target will have a positive SNR when it is located behind a sheet of plywood and wall board. When brick is added to the wall the SNR drops to about 0 dB with the $G_{RxSL} = -80$ dB. Adding concrete block to the first wall would require a $G_{RxSL}$ to be less than -90 dB (extrapolated). The range $R_r$ was reduced to 50 m. Figure 3.3 has the new results which indicates that the wall that includes brick will now have a positive SNR with a $G_{RxSL}$ of -70 dB. Similarly, the $G_{RxSL}$ requirement for the wall with concrete block has dropped to about -80 dB to produce a positive SNR value.

### 3.2 Indirect Path Analysis

Figure 3.4 shows the signal from the transmitter reaching the target via an indirect path. Rather than the target being located behind a wall, the target is located behind a larger structure, and it is only within the line of sight of the bistatic receiver. We will use the
Figure 3.2: Bistatic receiver through-wall analysis ($R_r = 200$ m).

Figure 3.3: Bistatic receiver through-wall analysis ($R_r = 50$ m).
Figure 3.4: Bistatic receiver indirect path schematic.

The electromagnetic theory of plane waves to obtain a reflection coefficient that will be used to attenuate the signal power reaching the target. The electromagnetic plane wave reflecting from the two walls obeys Snell’s law of reflection [13, 14]

$$\theta_I = \theta_R$$  \hspace{1cm} (3.10)

and Snell’s law of refraction,

$$\frac{\sin \theta_I}{\sin \theta_T} = \sqrt{\frac{\mu_2 \varepsilon_2}{\mu_1 \varepsilon_1}} = \sqrt{\frac{\mu_o \varepsilon_o \varepsilon_r}{\mu_o \varepsilon_o}} = \sqrt{\varepsilon_r}$$ \hspace{1cm} (3.11)

where the wave interface is between air (1) and wall (2) and $\varepsilon_r$ is the dielectric constant of the wall.

The wave impedances in each medium are defined as

$$\text{air} \rightarrow \eta_1 = \sqrt{\frac{\mu_o}{\varepsilon_o}} = 120\pi$$ \hspace{1cm} (3.12)

$$\text{wall} \rightarrow \eta_2 = \sqrt{\frac{\mu_2}{\varepsilon_2}} = \sqrt{\frac{\mu_o \mu_r}{\varepsilon_o \varepsilon_r}} = \eta_1 \sqrt{\frac{\mu_r}{\varepsilon_r}} = 120\pi \sqrt{\frac{1}{\varepsilon_r}}.$$ \hspace{1cm} (3.13)

The reflection coefficient as a function of polarization of the electromagnetic wave has been derived [13, 14]. The plane of incidence is defined as perpendicular to the interface.
Parallel polarization is defined as the electric vector being located on the plane of incidence. The reflection coefficient with parallel polarization is defined as

$$\Gamma_\parallel = \frac{E_r}{E_i} = \frac{\eta_2 \cos \theta_T - \eta_1 \cos \theta_I}{\eta_2 \cos \theta_T + \eta_1 \cos \theta_I}$$

(3.14)

Perpendicular polarization has the electric field vector located perpendicular to the incident plane, and the reflection coefficient is defined as

$$\Gamma_\perp = \frac{E_r}{E_i} = \frac{\eta_2 \cos \theta_I - \eta_1 \cos \theta_T}{\eta_2 \cos \theta_I + \eta_1 \cos \theta_T}$$

(3.15)

The loss of signal power due to the reflections from the wall bounce is defined as,

$$L_{WB} = (\Gamma^2)^{N_r}$$

(3.16)

where $\Gamma$ is defined by either (3.14) or (3.15) because of our small incident angle (8.3°) and $N_r$ has the total number of reflections from the two walls. The signal power reaching the target from the transmitter is defined as

$$S_{Tgt} = \frac{P_{Tx} G_{Tx} L_{WB}}{4\pi R_T^2}.$$  
(3.17)

The signal bounces off the target and reaches the receiver with a signal power of

$$S_{Rx} = \frac{S_{Tgt} G_{Rx} \lambda^2 \sigma}{(4\pi)^2 R_T^2} = \frac{P_{Tx} G_{Tx} G_{Rx} \lambda^2 \sigma}{(4\pi)^3 R_T^2 R_r^2} L_{WB}$$  
(3.18)

Since we are again interested in the signal-to-interference plus noise ratio (SINR), we can now use (3.1), (3.18), and the thermal noise to find SINR:

$$SINR = \frac{S_{Rx}}{L \cdot (kT_0 F_n B + I_{DP})} T_{coh} F_s \cdot d.$$  
(3.19)

$$SINR = \frac{P_{Tx} G_{Tx} G_{Rx} \lambda^2 \sigma L_{WB}}{(4\pi)^3 R_T^2 R_r^2 L \cdot (kT_0 F_n B + I_{DP})} T_{coh} F_s \cdot d.$$  
(3.20)

where $T_{coh}$ is the coherent integration time.

The parameters listed in Table 3.2 were also used for this portion of the analysis. The target is assumed to be located down a 10 m wide corridor. The length of the corridor, combined with the angle listed in Table 3.3, was changed to produce two different number of reflections: single reflection and 3 reflections. Figure 3.5 has the SINR as a function
Table 3.3: Parameters used in the baseline bistatic indirect path SINR analysis

<table>
<thead>
<tr>
<th>Material</th>
<th>Dielectric Constant</th>
<th>$\theta_I$</th>
<th>$|\Gamma|$</th>
<th>$|\Gamma_\perp|$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plywood</td>
<td>2.45</td>
<td>8.3°</td>
<td>0.22</td>
<td>0.22</td>
</tr>
<tr>
<td>Brick</td>
<td>4.48</td>
<td>8.3°</td>
<td>0.35</td>
<td>0.36</td>
</tr>
<tr>
<td>Concrete Block</td>
<td>2.19</td>
<td>8.3°</td>
<td>0.19</td>
<td>0.20</td>
</tr>
</tbody>
</table>

of dielectric constant involving a single wall bounce. Figure 3.5 shows that a positive SNR is observed when the signal bounces off the three different wall materials considered, plywood, concrete block, and brick, with the receiver located between 14 m and 438 m.

The corridor length was increased in length to produce three wall bounces. Figure 3.6 shows the new results. The new results show that with a 14-m receiver-to-target range, the SNR will be negative for most dielectric constants in the range of plywood and concrete block. If the walls are made of brick, the receiver can be located at a range of 100 m and have a positive SNR.
Figure 3.5: Indirect path – single bounce.

Figure 3.6: Indirect path – three bounce.
Conclusions

There is a growing need for persistent SAR surveillance of very large areas. Accomplishing this goal with multiple cooperating close-range observers may present operational advantages over the use of a single high-cost orbiting platform. This thesis analyzed the feasibility of deploying multiple SAR systems to observe contiguous regions of a larger area. A single monostatic SAR system was modeled first in order to provide a baseline for comparison. Two cooperating systems using LFM and coded waveforms were also modeled. Preliminary results indicate that two cooperating platforms using identical LFM waveforms will require the multi-path interference to be minimized by reducing the antenna main beam overlap. This requires the two airborne transmitters to be synchronized in position (i.e., located along the same side of the area of interest and having the same directional velocity vector). The antenna size also has to be increased from a 10.6 cm x 21 cm to a 21 cm x 21 cm to reduce the antenna’s cross range beamwidth. Alternatively, orthogonally frequency modulated waveforms could be employed to eliminate the effects of mutual interference between two platform transceivers. Mutual interference may be eliminated by using orthogonal frequency modulated waveforms [16, 17].

The bistatic receiver through-wall study indicated two variables are required to ensure a positive through-wall SNR value. The direct path interference must be minimized by reducing the receiver antenna side lobe. The receiver also needs to be located within 50 m of the target. The wall attenuates the X-band signal power considerably. The indirect path study also had a similar conclusion. Having the receiver close to the target increases the received signal power which produces higher SNR values. Also, each wall bounce reduces the signal power reaching the target and then the receiver.
Future research will increase the understanding of and ability to exploit multiple radar systems operating in close proximity to each other. Additional studies should closely address system engineering, specifically involving algorithm development for exploiting the bistatic data. Next, integration with range-Doppler imaging should be investigated. Finally, the use of nulling should be considered to lower the clutter, both between the two airborne transmitters and also between the transmitters and the bistatic receiver.
Bibliography


