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Inverse synthetic aperture radar imaging using a coherent ultrawideband random noise radar system

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Abstract. The University of Nebraska-Lincoln has developed an ultrawideband random noise radar operating over the 1- to 2-GHz frequency range. The system uses the technique of heterodyne correlation, and is thus phase coherent. It has therefore been used in applications such as interferometry, polarimetry, and Doppler estimation. This assesses the performance of this radar as a range-Doppler imaging system, in particular, inverse synthetic aperture radar (ISAR). By performing turntable experiments, we examine a number of issues that arise both from the wideband nature of the radar system and from the randomness of the transmit signal. Although the ultrawideband nature of the signal does yield some ambiguity as to the cross-range resolution of the system, it is seen that one may use the usual equation for cross-range resolution in narrowband systems with reasonable confidence. In addition, the random nature of the transmit waveform gives rise to variations from look to look in the ISAR images. High correlations are obtained for voltage and power amplitudes, while the complex voltage and the phase angle show low correlations from look to look. Finally, we present polarimetric color images of a complex target that captures its polarimetric scattering characteristics, which may be useful in automatic target recognition. © 2001 Society of Photo-Optical Instrumentation Engineers. [DOI: 10.1117/1.1413212]

Subject terms: inverse synthetic aperture radar; radar imaging system; random noise radar; ultrawideband.

Paper 200220 received June 1, 2000; revised manuscript received May 3, 2001; accepted for publication May 8, 2001.

1 Introduction

Since the early 1950s,¹ research has been conducted on the feasibility of using a noise or noise-like signal as a radar transmit signal. These radars are relatively inexpensive to build and possess natural immunity to jamming and interference from other radar systems operating in the same theater of operations. They have been used for the measurement of range profiles,² Doppler estimation,³ detection of buried objects,⁴ interferometry,⁵ and inverse synthetic aperture radar (ISAR) and synthetic aperture radar (SAR) imaging,⁶⁻⁸ to name a few applications. This list is far from complete and does not reflect the extent of the literature on noise radars. Guosui, Hong, and Weimin⁹ provide a more complete listing of references. The ability to simultaneously estimate range and Doppler are the necessary ingredients for range-Doppler imaging.^{10,11} However, Doppler measurement presumes that the radar is phase coherent. Although a random noise signal is, by very definition, totally incoherent, it is possible to inject phase coherence in a system using the technique of heterodyne correlation. At the University of Nebraska-Lincoln, such a phase-coherent random noise radar, operating in the 1- to 2-GHz range, has been under investigation for a number of years, and whose potential for SAR or ISAR imaging is currently being explored. We present a series of turntable

experiments with this unique radar system that clearly underscores its ability to yield high resolution ISAR images.

2 Description of Radar System

A block diagram of the polarimetric random noise radar system is shown in Fig. 1. A brief description is presented herein. Narayanan et al.⁴ provide a detailed description.

The noise signal generated by OSC1 provides a wideband noise signal with a constant power spectral density in the 1- to 2-GHz frequency range, and a power output of 0 dBm. This output is split into two in-phase components in power divider PD1. One of these outputs is amplified in a 34-dB gain power amplifier AMP1. The output of the amplifier is connected to ANT1, which is a vertically polarized ridged horn antenna whose E/H plane beamwidths and gain at 1.5 GHz are 23 deg, 34 deg, and 17 dB, respectively.

The other output arm of power divider PD1 is connected to a combination of fixed and variable delay lines DL1 and DL2, respectively. These delay lines are used to provide the necessary time delay for the sampled transmit signal, so that it can be correlated with the received signal scattered from objects at different ranges. The variable delay line can be rapidly programmed to step through 128 delays in steps of 0.159 ns.

To perform coherent processing of the noise signals, a unique frequency translation scheme is employed. The primary component of this technique is a 160-MHz phase-

Fig. 1 Block diagram of ultrawideband polarimetric random noise radar system.

locked oscillator OSC2. This is connected via power divider PD2 to the IF input terminal of MXR1. This coherent noise signal is split by power divider PD3 into two channels: the copolarized and the cross-polarized channels. Also, the second output of the power divider PD2 is again split into two 160-MHz signals in power divider PD4 and fed into two I/Q detectors.

We now discuss the signal processing of the copolarized channel. The cross-polarized channel operation is identical, so it will not be repeated. One of the outputs of PD3 is amplified by AMP4 and the output signal is used as the local oscillator (LO) input to a biasable mixer MXR2, whose RF input is obtained from the copolarized channel of the receive antenna ANT2 and a low noise amplifier AMP2. The receive antenna ANT2 consists of a broadband constant 7.5-dB gain log-periodic antenna with E/H plane beamwidths of 65 and 100 deg, respectively. Amplifier AMP2 is used to improve the noise figure at the receiver input. In general, the RF input signal to the mixer MXR2 consists of transmitted noise at 1 to 2 GHz scattered and reflected from various objects/interfaces. However, since the LO signal has a unique delay associated with it, only the signal scattered from the appropriate range bin will mix with the LO to yield an IF signal at frequencies centered around 160 MHz. Signals scattered or reflected from other range bins will not provide an approximately constant frequency of 160 MHz, since the return signal and the LO signal decorrelate quickly. The output of the mixer MXR2 is connected to a narrowband bandpass filter FL1 of center frequency 160 MHz and bandwidth 5 MHz, ensuring that only 160 MHz plus the Doppler information encoded in the noise gets through.

The output of filter FL1 at 160 MHz is split into two outputs in power divider PD5. One of these outputs is amplified and detected in a 70-dB dynamic range 160-MHz logarithmic amplifier AMP6. The other output of the power divider PD5 is connected to one of the inputs of an I/Q detector IQD1, whose reference input is one of the outputs from PD4. Both of the signals are approximately at 160 MHz; thus the I/Q detector provides the in-phase (I) and quadrature (Q) components of the phase difference between the two signals. Since frequency translation preserves phase differences, the I and Q outputs can be related to the polarimetric copolarized scattering characteristics. The system therefore produces the following outputs at various ranges as set by the delay lines: copolarized amplitude, copolarized phase angle, cross-polarized amplitude, and cross-polarized phase angle. Thus, the system outputs can be related to the polarimetric scattering characteristics of the target.

Finally, we form an output of the radar system for further processing in the following manner. Let A be the output of the logarithmic amplifier AMP6 (AMP7) but converted to a linear scale. Let I and Q be the outputs of the I/Q detectors adjusted for I/Q imbalance. The receiver output r is expressed as

$$r = A \exp[j \arctan(Q/I)], \quad (1)$$

for further analysis. Using the output of the logarithmic amplifier allows us to faithfully capture wide dynamic range amplitude variations.

3 Theory of Stochastic ISAR Imaging

We consider bandlimited white noise in the frequency range 1 to 2 GHz as our interrogating waveform. As our receiver is coherent, i.e., it is a quadrature receiver, we may view this waveform as complex. It can be modeled as a stochastic voltage waveform $\mathcal{S}[t]$ that is zero mean, ergodic, and possessing a power spectral density function

$$S(f) = \begin{cases} S_0, & f_0 - \beta/2 \leq f \leq f_0 + \beta/2 \\ 0, & \text{otherwise} \end{cases}, \quad (2)$$

where β is the bandwidth (1 GHz), and f_0 is the mean frequency (1.5 GHz). The associated correlation function is given by

$$\begin{aligned} B(\tau) &= \int \exp(2\pi j f \tau) S(f) df \\ &= E_0 \exp(2\pi j f_0 \tau) \text{sinc}(\beta \tau), \end{aligned} \quad (3)$$

where $E_0 = \beta S_0$ is the total power in the signal. This gives the system a down-range resolution of

$$\Delta x = c/2\beta = 15 \text{ cm}. \quad (4)$$

For a target rotating with axis of rotation transverse to the radar line of sight (LOS), we can define a natural target-centric coordinate system. For simplicity, we assume that the axis of rotation is perpendicular to the LOS. Let the LOS define the x axis, the axis of rotation define the z axis, and choose the y axis to be perpendicular to these two directions, as seen from Fig. 2. Assume that the target is rotating at a constant angular velocity Ω and let R_0 be the distance between the radar antenna and the origin of our coordinate system. Although a target in the ultra-wideband paradigm may be dispersive, for simplicity, we assume the far-field weak-scatterer model with nondispersive scatterers, and we denote $\rho(x, y, z)$ as the target density function. This assumption does yield a reasonable scattering model, however, for a more detail scattering model this should be relaxed. Under this assumption the received waveform is then given by

$$r(t) = \int \int \int \rho(x, y, z) \mathcal{S}[t - \tau(x, y, t)] dx dy dz, \quad (5)$$

where $\tau(x, y, t) = 2(R_0 - x')/c$ with $x' = x \cos(\Omega t) + y \sin(\Omega t)$. Note that τ is the two-way delay from the radar to a target subscatterer. The z -direction variation does not

Fig. 2 The geometry of inverse synthetic aperture radar.

play any appreciable role in ISAR imaging,¹² and thus is not considered in what follows. We therefore define the two-dimensional target density function $\rho(x, y)$ as

$$\rho(x, y) = \int \rho(x, y, z) dz.$$

Following the range-Doppler approach¹¹ to ISAR imaging, we make the approximation

$$\tau(x, y, t) \approx \frac{2(R_0 - x)}{c} + \frac{2\Omega y}{c} t = \tau(x) + \alpha(y)t. \quad (6)$$

This approximation transforms Eq. (5) into

$$r(t) \approx \int \int \rho(x, y) \mathcal{S}[(1 - \alpha(y))t - \tau(x)] dx dy. \quad (7)$$

On reception, $r(t)$ is mixed with a delayed replica of $\mathcal{S}[t]$, i.e. $\mathcal{S}^*[t - \tau_n]$, and sampled in time to yield

$$\begin{aligned} R(t_m, \tau_n) &= \int \int \rho(x, y) \mathcal{S}[(1 - \alpha(y))t_m - \tau(x)] \\ &\quad \times \mathcal{S}^*[t_m - \tau_n] dx dy \quad 1 \leq m \leq M, 1 \leq n \leq N. \end{aligned} \quad (8)$$

The desired measured data is the expectation of $R(t_m, \tau_n)$ given by

$$\mathbf{E}\{R(t_m, \tau_n)\} = \int \int \rho(x, y) B[\tau_n - \tau(x) - \alpha(y)t_m] dx dy. \quad (9)$$

The function $B[\tau_n - \tau(x) - \alpha(y)t_m]$ in Eq. (9) is the time domain point-spread function (also called the correlation function). In the previous section, the process of frequency translating the delayed replica to the upper sideband, mixing this with the return signal, and then passing this through a bandpass filter and I/Q detector approximates the previous expectation operation. As we see in what follows, this is because only the same frequency components of the replica and return signal correlate.

If the integration time T (the span of the time samples t_m) is small enough, then the rotation of the target will produce approximately constant Doppler frequencies for subscatterers in a particular range bin. We take the Fourier transform of $B[\tau_n - \tau(x) - \alpha(y)t_m]$ with respect to the variable t_m to obtain the frequency domain point-spread function $\hat{B}[\tau_n, y_m; \tau(x), \alpha(y)]$. As shown in Bell and Narayanan,¹³ this function is localized around the point $\tau_n = \tau(x)$, $y_m = \alpha(y)f_0$. The resolution in the τ direction, i.e., the down-range resolution, is proportional to $1/\beta$, and the resolution in the y direction, i.e., the cross-range resolution, is proportional to $1/T$ (when α is small). Figure 3 shows the absolute value of the frequency domain point-spread function for the zero Doppler (stationary target) and the nonzero Doppler (moving target) cases. Using the fact that $\tau(x)$ and $\alpha(y)$ are proportional to the x and y coordinates, respectively, we arrive at our desired image.

Fig. 3 The frequency domain point-spread function: (a) surface plot with no Doppler present, (b) contour plot with no Doppler present, (c) surface plot with mean Doppler of 100 Hz, and (d) contour plot with mean Doppler of 100 Hz.

To obtain the desired image, one must first estimate the mean value $\mathbf{E}\{R(t_m, \tau_n)\}$. However, in ISAR applications, we frequently have only a few, sometimes only one, realizations of $R(t_m, \tau_n)$ to work with. To evaluate what one realization, or one look, of $R(t_m, \tau_n)$ looks like, we use a Fourier polynomial with random coefficients to approximate $\mathcal{S}[t]$. More specifically, given any number $\epsilon > 0$, there exist values T , N , constants $\{f_i\}_{i=-N}^N$, and zero mean uncorrelated random variables $\{\mathcal{B}_i\}_{i=-N}^N$ (all depending on ϵ), such that

$$\mathbf{E}\left\{\left|\mathcal{S}[t] - \sum_{i=-N}^N \mathcal{B}_i \exp(2\pi j f_i t)\right|^2\right\} < \epsilon, \quad (10)$$

for $-T \leq t \leq T$.¹⁴ The correlation function, $\tilde{B}(\tau)$, of the approximating process,

$$\tilde{s}(t) = \sum_{i=-N}^N \mathcal{B}_i \exp(2\pi j f_i t), \quad (11)$$

is given by

$$\tilde{B}(\tau) = \sum_{i=-N}^N \mathbf{E}\{|\mathcal{B}_i|^2\} \exp(2\pi j f_i \tau). \quad (12)$$

If one lets $s(t)$ be a realization of $\mathcal{S}[t]$, b_i be a realization of $\mathcal{B}_i$, and assumes a nondispersive point scatterer $\rho(x_0, y_0) = \rho_0$, then we can compute a realization of $R(t_m, \tau_n)$ as

Fig. 4 Geometry of the experimental setup.

$$\begin{aligned} & \rho_0 s\{[1 - \alpha(y)]t_m - \tau(x)\} s^*(t_m - \tau_n) \\ &= \rho_0 \left[\sum_{i=-N}^N |b_i|^2 \exp(2\pi j f_i [\tau_n - \tau(x)]) \right. \\ & \quad \times \exp[2\pi j f_i \alpha(y) t_m] + \sum_{i \neq k} b_i b_k^* \exp\{2\pi j [f_i \tau_n \\ & \quad \left. - f_k \tau(x)]\} \exp(2\pi j \{[1 - \alpha(y)]f_i - f_k\} t_m) \right]. \quad (13) \end{aligned}$$

Correlating at $\tau_n = \tau(x_0)$ yields

$$\begin{aligned} & \rho_0 s\{[1 - \alpha(y)]t_m - \tau(x)\} s^*(t_m - \tau_n) \\ &= \rho_0 \left[\sum_{i=-N}^N |b_i|^2 \exp[2\pi j f_i \alpha(y) t_m] \right. \\ & \quad + \sum_{i \neq k} b_i b_k^* \exp[2\pi j (f_i - f_k) \tau_n] \\ & \quad \left. \times \exp(2\pi j \{[1 - \alpha(y)]f_i - f_k\} t_m) \right]. \quad (14) \end{aligned}$$

This equation gives us a model of what $r(t)$, as given by Eq. (1), may look like. As alluded to earlier, the process of downconverting to an intermediate frequency level, and then mixing down and passing through a quadrature detector, will result in the two sums in Eq. (14) being restricted to those frequencies such that

$$|[1 - \alpha(y)]f_i - f_k| < \eta, \quad (15)$$

where η is the resultant passband of FL1 (FL2) and the quadrature detector. The first sum is our (single look) estimate of the time domain image, while the double sum will contribute to image degradation. An ensemble average will serve to improve our image quality; however, as mentioned before, few samples are usually available for this kind of statistical estimation. Because of this, we take the point of view that one realization represents our mean image. As shown in the next section, we do get surprisingly good images that support our point of view.

Table 1 Description of targets used.

Target	Description
A	Cylinders of radii 15 and 33 cm.
B	Cube of dimensions 60×60×60 cm.
C	Corner reflector of side 35 cm.
D	Airplane shown in Fig. 5.

4 Experimental Setup

To examine the effectiveness of an ultrawideband random noise radar configured in ISAR mode, we conducted a series of turntable experiments. To minimize background clutter, we collected data in an anechoic chamber, as shown in Fig. 4. Test targets were placed on a turntable of 1.2 m diam. The top of the turntable was constructed of acrylic (dielectric constant of 3.3). This was placed on a wooden frame that housed an electric motor and motor control unit. The wooden frame was then covered with a microwave absorbing panel (Millimeter Wave Technology, Inc. MAC-6112) with relative absorption of about -19 dB over 1 to 2 GHz.

We used several test targets in our experiments. These are listed in Table 1. Target A consisted of two metallic cylinders of diameters 15 and 33 cm, respectively. Target B was a 60×60×60-cm cube with its center placed on the turntable's axis of rotation, constructed of cardboard, and covered with aluminum foil. This target possesses both co-polarized (the flat faces) and cross-polarized (the edges) scattering characteristics. Target C was a 35-cm corner reflector with its apex placed slightly off of the turntable's axis of rotation. The final target, Target D, was a model plane, as shown in Fig. 5. Its fuselage was constructed of an aluminum conduit, and its wings were made of cardboard covered with aluminum foil. All of the targets were placed directly on the turntable (as shown in Fig. 4) except for the plane, which was elevated on cardboard pylons to create a grazing angle of 14 deg.

We desired to simulate the collection of data in all range bins simultaneously, i.e., to simulate collecting data with a

Fig. 5 Geometry of the model airplane (Target D).

Fig. 6 The cross-range resolution experimental setup and results: (a) target arrangement for the cross-range resolution experiment, and the ISAR images with narrowband predicted cross-range resolution of (b) 15 cm and (c) 30 cm (normalized returns).

bank of correlation receivers described in Sec. 2. To do so, the following data collection procedure was used.

1. Rotate target at a known angular velocity.
2. Set delay line to first range bin.
3. Collect 360 deg of data at this range bin starting from a reference point.
4. Advance delay line to next range bin.
5. Starting at the same reference point in the rotation of the target as in step 3, collect 360 deg of data.
6. Repeat through all of the relevant range bins.

The noise signal that we utilize (1 to 2 GHz) decorrelates in about 1 ns. Thus, it was considered adequate to execute step 4 at 0.477 ns (3 delay steps of DL2) to satisfy the

Nyquist sampling criterion. To a certain degree, this collection method simulated a single realization (look) of the return $R(t_m, \tau_n)$.

5 Experimental Results

5.1 Cross-Range Resolution

If our system were a narrowband system, then the cross-range resolution, Δy , would be given approximately by¹¹

$$\Delta y = \frac{\lambda}{2\Delta\theta}, \quad (16)$$

where λ is the wavelength of the carrier frequency, or mean wavelength, and $\Delta\theta$ is the rotation angle over which the

Fig. 7 Single look images of the (a) side and (b) corner of the cube (using power in mW).

measurement is made ($\Delta\theta$ is usually a small angle). However, since the fractional bandwidth of our system is 2/3, the narrowband approximation may or may not be valid. Furthermore, over such large fractional bandwidths, the scattering from the target is likely to be dispersive in nature, and our nondispersive approximation may not be strictly valid. As seen in Fig. 3 and discussed in Bell and Narayanan,¹³ the main lobe of the point-spread function is localized around the location of a point scatterer. To examine whether one can use Eq. (16) effectively, we conducted an experiment involving the two cylinders (Target A) described in the previous section. The target arrangement is shown in Fig. 6(a). Using the mean wavelength of 20 cm, Eq. (16) predicts a cross-range resolution of 15 cm (giving a square resolution window) for $\Delta\theta \approx 20$ deg, and a cross-range resolution of 30 cm for $\Delta\theta \approx 40$ deg. We point out that our L-band radar is not well suited for these experiments, owing to its large wavelength, since to achieve cross-range resolution on the same scale as down-range resolution, we need to rotate through large angles. This will

Table 2 The statistics of the correlation coefficients.

Cube		
Image	Mean	Standard Deviation
complex voltage	0.5173	0.0499
voltage amplitude	0.7301	0.0347
power in dBm	0.3982	0.0253
phase angle	0.1273	0.0228
Corner Reflector		
Image	Mean	Standard Deviation
complex voltage	0.5968	0.0529
voltage amplitude	0.7781	0.0421
power in dBm	0.3686	0.0400
phase angle	0.1172	0.0263

result in image blurring due to the range walk phenomenon. Despite this difficulty, the ISAR images for these two values of $\Delta\theta$ shown in Figs. 6(b) and 6(c) demonstrate that we do achieve a cross-range resolution on the order of that predicted by Eq. (16). Indeed, we can see that the smaller cylinder is distinguishable from the larger cylinder when $\Delta\theta \approx 40$ deg, but the two are much harder to distinguish when $\Delta\theta \approx 20$ deg, as predicted by Eq. (16). In what follows, we use this relationship to place cross-range scales on all our displayed images.

5.2 Variation from Look to Look

Each image that one obtains with a random signal radar is a single realization, or look, of a random field. Single look images of the corner and side of the cube are shown in Fig. 7. As seen from Eq. (14), not only is there variation from ambient noise both in the environment and the hardware, but there is also variation from the cross terms that result from the fact that the noise signal does not completely decorrelate. As mentioned before, in many ISAR applications, the target under interrogation is uncooperative, thereby making it very difficult to collect a number of images that one would use to average out these variations. These facts motivate us to examine the variation from look to look. We used the cube (Target B) and the corner reflector (Target C) to conduct an experiment to investigate this.

The experiment was conducted by performing six data sets (trials) on the same target. Each of the 360-deg polarimetric data sets yielded a number of complex images. We selected 17 images from each data set for the analysis. Each image was viewed as being composed of four bands: HH, HV, VH, and VV. To quantify the variation from look to look, we correlated corresponding images in different trials, and examined the correlation coefficients. That is, using Image 1 as an example, we calculated

$\text{Cor}(\text{Trial } m: \text{Image 1}, \text{Trial } n: \text{Image 1}),$

$$m \neq n, m, n = 1, \dots, 6,$$

Fig. 8 Histograms of the image correlation coefficients. Histograms for (a) the complex voltage (coefficients in modulus), (b) voltage amplitude, (c) power in dBm, and (d) phase angle.

yielding $5 + 4 + 3 + 2 + 1 = 15$ correlation coefficients for Image 1. Since we used 17 images, this gives a total of 255 correlation coefficients.

Four correlation coefficients were calculated for each pair using the following parameters:

1. the complex voltages,
2. the amplitude of the voltages,
3. the power in units of dBm, and
4. the complex phase angles $\{\exp[j \arctan(Q/I)]\}$.

For cases 1. and 4. we used the modulus of the complex correlation coefficient for analysis. The means and standard deviations of the distribution of correlation coefficients are shown in Table 2. The histograms of the correlation coefficients for the cube are shown in Fig. 8. The statistics reveal that the individual images correlate well using either of the voltage amplitudes. For the power (in dBm), the statistics reveal a much lower level of correlation do to the fact that

the logarithm function stretches and gives finer detail to the image. The complex phase angles do not correlate well. This along with the relatively high correlation of the amplitude images gives a lower average value for the modulus of the complex correlation coefficients. The high correlation of the voltage amplitudes and the low correlation of the phase angles can be attributed to both the nature of wide-band random noise signal, and the nonidealities of the hardware components. Addressing the latter concern first, the voltage amplitudes exhibit high correlation, since the logarithmic amplifiers (AMP6 and AMP7) have a wide dynamic range allowing them to capture amplitude variations faithfully. We believe the low correlation between the phase angles may be attributable to the phase uncertainties associated with the wideband probing signal. Note from Eq. (14) that since the coherent return information is encoded over a large bandwidth, the Doppler phase shift for rotating targets is spread over a large range of values, making it somewhat ambiguous. Furthermore, we note that the inher-

Fig. 9 Typical histograms of an image: (a) power in dBm, (b) voltage amplitude, and (c) phase angle.

ent frequency cross terms again make phase detection ambiguous. However, if our assumption that our target is non-dispersive is valid, these terms do not degrade the voltage amplitudes quite as much.

To examine the images further, we plotted the histograms of the images themselves, both in amplitude and in phase. Figure 9 shows representative histograms. The amplitude histograms tend to be distributed in what appears to be a log-normal distribution, while the power (in dBm) histogram appears Gaussian, as expected. The ambiguity in phase angles tends to suggest that the phase angle histograms will be uniformly distributed. This is somewhat the case, however we do see a bias toward 0 and $\pm\pi$. The authors are unsure as to the cause of this bias. It may be due to instabilities in the I/Q detectors or it may be inherent in the image model. This is currently being investigated.

5.3 Single Look Images of the Plane

To examine how the system images more complex targets, we examine the images of the airplane (Target D). As mentioned before, the model was placed so as to induce a graz-

ing angle of 14 deg. The images shown are the amplitude of the image normalized to the maximum and minimum of the four bands (VV, VH, HV, and HH). Figure 10 shows the VV polarization, the mean of the VH and HV polarizations, and the HH polarization. The pixelwise transform x^γ , with $\gamma=1.2$, was applied to enhance the features of the airplane. Despite the fact that these are single look images, the wings and fuselage of the airplane can be readily recognized. It is also possible to capture the polarimetric characteristics of the target by forming a color red/green/blue (RGB) image of the target, which can yield useful clues about the nature of strong scattering centers on the target for automatic target recognition (ATR) applications. Figure 11 is such an RGB color composite polarimetric image: red corresponds to the VV return, green to the mean of the VH and HV returns, and blue to the HH returns. We see the green returns along the edges of the fuselage where the wings meet it, just where one might expect a dihedral-type reflection. The outline of the airplane can be reasonably well identified in the color image.

Fig. 10 Polarimetric images of the airplane. (a) VV, (b) VH/HV, and (c) HH.

6 Conclusions

We have shown that a random noise radar can successfully obtain reasonably good ISAR images. Continued developments in delay line technology, digital radio frequency memory, and signal processing will make it possible to use random noise radars as range-Doppler imaging systems in future military and civilian applications. Further investigations are planned with modifications to the radar system.

The need to rotate through a large angle ($\Delta\theta \approx 40^\circ$) to obtain a square resolution window will always degrade image quality. This can be remedied via postprocessing to compensate for range walk; however, an obvious implementation that would greatly reduce the problem is to change the frequency band and operate at higher frequencies. For example, it would be inexpensive to modify the radar system in Fig. 1 by placing a 9-GHz oscillator, a mixer, and bandpass filter, just before AMP1 (with other appropriate modifications), thus making the transmit signal in X band (10 to 11 GHz). On reception, one would mix down again to the 1- to 2-GHz level before passing the

received signal through AMP2 and AMP3. This would reduce $\Delta\theta$ by a factor of 10 to achieve the same cross-range resolution.

Other methods of image formation are being investigated. We explore the most common method used for range-compressed data: the FFT. However, other algorithms, although more complex, may result in better single look images. For example, the filtered back-projection algorithm could be used, or by processing the range compressed data into the frequency domain, one could implement polar reformatting. However, the latter may not result in better images, as the preprocessing step will ultimately use an FFT algorithm and hence be closely related to the method already used.

The assumption that the target is nondispersive may not be a valid assumption for ultrawideband systems. Relaxing this assumption and modifying our point-scatter model (see Borden¹²) may yield better images, and possibly motivate new imaging formation algorithms. Again, the resulting al-

Fig. 11 RGB color composite polarimetric image of the airplane.

gorithm would be more complex, but the trade-off in image quality may be worth it.

Finally, a deeper investigation into the theoretical aspects of radar imaging using stochastic waveforms¹³ may reveal a better understanding of the scattering mechanisms, and hence, motivate other image formation techniques.

Acknowledgments

This work was supported by ONR contract # N00014-97-1-1061. The authors would like to thank Dr. William Miceli for his helpful suggestions, and the members of the Environmental Remote Sensing Laboratory at the University of Nebraska-Lincoln, for many helpful conversations.

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