Compact Range Radar Cross-Section Measurements Using a Noise Radar

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***Abstract*— This paper discusses the measurement of radar cross section (RCS) with a very low-cost system that transmits band-limited random noise over the band from 1.0 to 4.0 GHz. The received signal is correlated with a delayed version of the transmitted signal. A variable delay line is used to obtain the response at various delay times. This yields the radar target impulse response as a function of delay. This can be transformed to yield both the amplitude and phase of the scattering matrix in the frequency domain.**

***Index Terms*—Electromagnetic measurements, radar cross sections.**

I. Introduction

In 1959, Horton [1] proposed a distance-measuring radar-transmitting modulated noise such that the distance was obtained from the correlation function. Four years later, Grant *et al.* [2] suggested that range ambiguities in radar systems may be reduced by transmitting wide-band noise. A number of noise radar systems have been used for target impulse response measurements. In the system developed by Cooper [3], the correlation was obtained from a digital correlator after converting from microwave to video frequencies. Forrest and Meeson [4] used a system where the noise is down converted before the delay line in order to minimize the delay-line losses. A discussion of a fixed delay-line system for inverse synthetic aperture radar (ISAR) imaging is given in Walton *et al.* [5]. A more complex noise radar was implemented by Narayanan [6] using a variable delay line with an intermediate mixing frequency to measure both the in-phase and quadrature-phase components.

In the radar discussed in this paper, wide-band noise is generated and radiated from a wide-band antenna. Some of the transmit signal is delayed and correlated with the received signal. The radar target impulse response is directly proportional to the correlation versus delay time response of the target. The correlation is implemented using a microwave mixer followed by a low-pass filter. A variable delay line is implemented using a set of microwave switches and incrementally increasing lengths of coaxial cable. The result is a very low-cost implementation of a coherent ultrawide-band radar.

We will discuss design and construction of the radar system as well as operation of the system in the Ohio State University compact range [7]. Calibration techniques unique to this system will be described and examples of the complex measured scattering matrix of spheres will be given.

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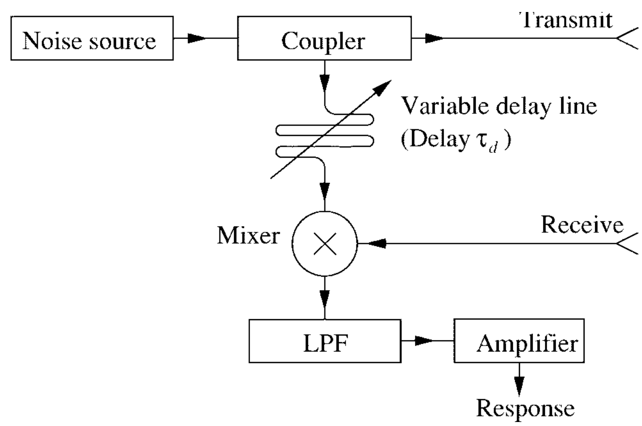

Fig. 1. Schematic representation of the noise radar.

II. The Radar System

In this section, we provide a brief description of the noise radar system. Consider the system shown schematically in Fig. 1. The noise source is realized by a 50-Ω load connected to the input of a high-gain amplifier. The coupler then provides a reference signal that is propagated through a coaxial cable delay line to produce the delayed signal. A mixer and low-pass filter provides the correlation between the received and delayed signals. The power level is such that the delay line signal drives the local oscillator port of the mixer at the required power level.

An important part of this system is the use of a variable delay line to measure the target response as a function of delay. The delay increments are achieved using four computer controlled mechanical ten-way switches. The ten output ports of two of the switches are connected with ten coaxial cables. Each cable is approximately 25 mm longer than the previous one. Switching the signal through the different lines thus allows the length to be varied in ten 25-mm increments. The ports on the other set of switches are connected with cables each of which is approximately 250 mm longer than the previous one. Connecting the two sets of switches in series allows incrementing the delay cable length in 100 increments of approximately 25 mm each. Due to the propagation speed in the cable, the delay-time increment is 0.12 ns. The actual delay times for each of the 100 increments were measured and tabulated.

The noise signal may also be characterized by its spectrum $\tilde{S}(\omega)$. The response of the target, including the transmit and receive system paths, may be described by the transfer function $\tilde{H}_x(\omega)$. The transfer function of an ideal delay line is $e^{-j\omega\tau_d}$ where τ_d is the delay time. The delay line is, therefore, characterized by $\tilde{H}_d(\omega, \tau_d)\, e^{-j\omega\tau_d}$ to account for the nonideal behavior (loss and dispersion versus frequency) of the line. The inputs to the mixer may then be written in the time domain by taking the inverse Fourier transform of the noise spectrum multiplied by the appropriate transfer function. The output of the mixer is, therefore, given by the product of two integrals. One of these integrals may be removed by the approximation that the output of the low-pass filter contains only the dc component. (In practice, it will cover a narrow band of frequencies which must be considered in order to compute the signal-to-noise ratio (SNR) [8].) The resulting dc signal is proportional to

$$R(\tau_d) = \int_{-\infty}^{\infty} |\tilde{S}(\omega)|^2 \tilde{H}_x(\omega) \tilde{H}_d^*(\omega, \tau_d)\, e^{j\omega\tau_d}\, d\omega. \quad (1)$$

If $\tilde{H}_d^*(\omega, \tau_d)$ is independent of τ_d, $R(\tau_d)$ may be interpreted as the inverse Fourier transform of $|\tilde{S}(\omega)|^2 \tilde{H}_x(\omega) \tilde{H}_d^*(\omega)$. Thus, the measured response as a function of delay may be interpreted as the impulse response of the target as measured with the radar system. From (1), we can see that if we take a Fourier transform of the measured response, $R(\tau_d)$, then we obtain the product of the target scattering coefficient $(\tilde{H}_x(\omega))$ and the radar transfer function $(|\tilde{S}(\omega)|^2 \tilde{H}_d^*(\omega))$.

There are, however, two practical effects that need to be considered in obtaining the spectrum in this way. First, the output of the mixer will contain some dc offset. In general this will depend on the power input to the mixer. The offset due to the delayed signal varies with the delay setting due to the loss in the cable. Therefore, a reference measurement must be subtracted from the target measurement. (The received power will also change the dc bias, but this effect is much smaller than that resulting from the delayed signal.) Fig. 2 shows the measured response for a 457-mm-diameter sphere as well as the no-target response. The slope of the data is a result of the dc offset varying with delay length. The difference between these curves is the sphere response, which is transformed to the frequency domain.

The second effect is due to the fact that $\tilde{H}_d^*(\omega, \tau_d)$ *is* a function of τ_d. We can measure this effect by characterizing the delay line using a network analyzer. Although this is a function of frequency, we can average the result for each delay increment and normalize the raw data by these terms. This will compensate for the average value of attenuation as the delay-line length is increased. (Note that we will be unable to compensate for the specific frequency rolloff effects present in the longer delay segments.)

III. Calibration

Calibration is the process of converting the experimental measurement data to a measure of the radar target RCS. First, we subtract the no-target measured impulse response data from the data where the target is present. In our case, this will remove both the dc offset (as discussed in Section II) and clutter (scattering from objects other than the target) [7]. Next, we compute the Fourier transform of this difference. Thus

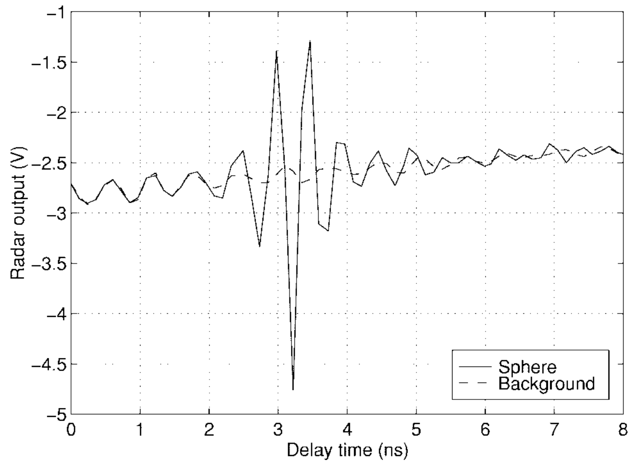

Fig. 2. Measured response of 457-mm-diameter sphere as a function of the delay time. The background profile obtained in the absence of the sphere is also shown.

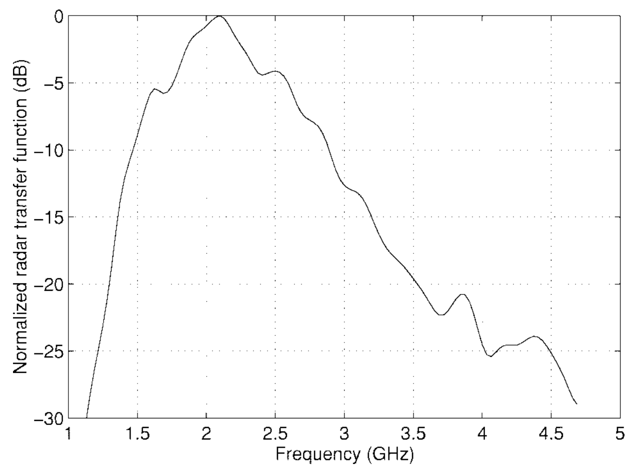

Fig. 3. The radar transfer function as a function of frequency.

$$\tilde{H}_m(\omega) = \mathcal{F}[h_{mt}(\tau_d) - h_{nt}(\tau_d)] \quad (2)$$

where $\mathcal{F}[h(t)]$ is the Fourier transform and $h_{mt}(\tau_d)$ and $h_{nt}(\tau_d)$ are the measured response data with and without the target. We can break up $\tilde{H}_m(\omega)$ as

$$\tilde{H}_m(\omega) = \tilde{H}_r(\omega)\tilde{H}_t(\omega) \quad (3)$$

where $\tilde{H}_r(\omega)$ is the frequency-domain transfer function of the radar and $\tilde{H}_t(\omega)$ is the frequency-domain radar-scattering coefficient of the target. $\tilde{H}_r(\omega)$ may be found from measurements on a known target. Let $\tilde{H}_{mc}(\omega)$ be the frequency-domain response obtained from (2) for a target with known scattering coefficient $\tilde{H}_c(\omega)$. Then, from (3)

$$\tilde{H}_r(\omega) = \frac{\tilde{H}_{mc}(\omega)}{\tilde{H}_c(\omega)}. \quad (4)$$

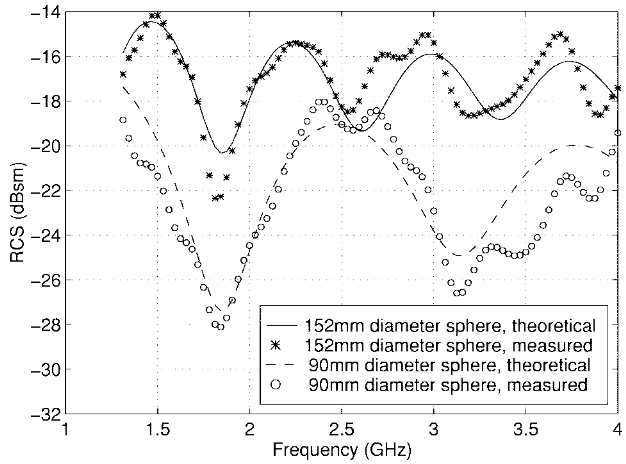

Fig. 4. RCS amplitude for two target spheres.

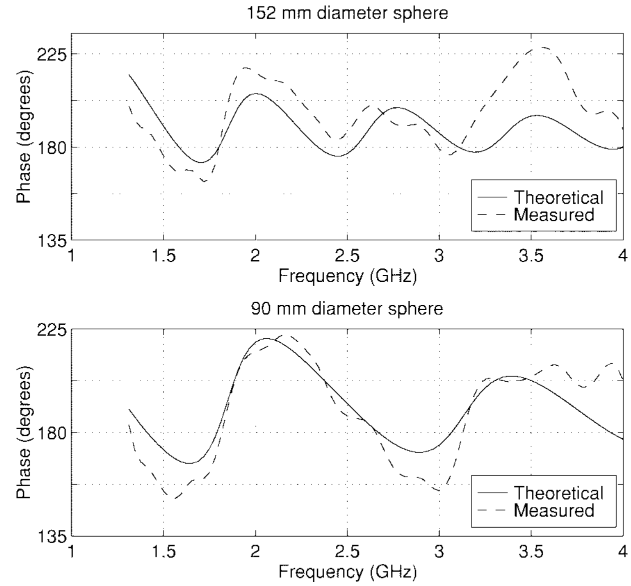

Fig. 5. Phase of the radar-scattering coefficient (referred to the front of the sphere) for two target spheres.

Fig. 3 shows the radar transfer function obtained by measurements on a 457-mm-diameter sphere (normalized to the maximum value). The magnitude of the radar transfer function is decreasing rapidly as the frequency is increased beyond 2.5 GHz. This is due to the increased delay-line loss at higher frequencies and the decrease in the transmit amplifier gain with frequency. The measured time-domain response is measured in delay increments of 0.12 ns, which implies an aliasing frequency of just above 4 GHz. In order to do a Fourier transform without aliasing the data, the target was moved and measured at three positions 6.4 mm apart. These sets were then interleaved and transformed using a discrete Fourier transform with the tabulated values of delay time.

Finally, the calibrated target scattering coefficients may be found from [7]

$$\tilde{H}_t(\omega) = \frac{\tilde{H}_c(\omega)}{\tilde{H}_{mc}(\omega)} \tilde{H}_m(\omega) = \frac{\tilde{H}_m(\omega)}{\tilde{H}_r(\omega)}. \tag{5}$$

IV. Measured Results

The noise radar system was set up in the Ohio State University compact RCS measurement range [7] and three different sized spheres were placed on a Styrofoam support in the test zone of the range. The sphere diameters were 457 mm (the calibration sphere), 152, and 90 mm. A fixed length delay line was used to set the initial delay value to a value near the front of the zone to be tested.

Data were collected by stepping the variable delay line in the previously discussed 100 steps of delay. This covered the time-domain extent of the spheres. Three step scans were taken on each sphere, where the sphere was moved 6.4 mm (corresponding to a time delay of 0.04 ns for two-way propagation) between each of the three scans. These three scans were interleaved before data processing to remove any aliasing effects as described in Section II. The data for the 152- and 90-mm-diameter spheres were calibrated by taking the 45-mm-diameter sphere as a reference as discussed in Section III.

The calibrated results are plotted in Figs. 4 and 5. The figures show, respectively, the amplitude (or radar cross section, in dBsm) and the phase (in degrees, referenced to the front of the sphere) of the scattering coefficient.

The calibrated amplitude is typically within 1 dB of the theoretical value except for the smaller sphere at the high-frequency region. The phase is within 20° with the same exception. This disagreement between experiment and theory is probably partly due to compact range stray signals and/or target and target support interaction effects. These stray signals and interactions will effect the data from both the test targets and the calibration targets [9], [10]. Also, as can be seen in Fig. 3 the radar transfer function is 20 dB down from its maximum at 3.5 GHz and it may be expected that the results will not be as good at higher frequencies.

V. Conclusion

This paper discusses the design and operation of a very low cost ultrawide-band instrumentation radar. The system uses a noise radar with a variable delay line to achieve ultrawide-band operation.

Measurements were made in the Ohio State University compact RCS range. We presented calibrated scattering matrix amplitude and phase results for 152- and 90-mm spheres over the band from 1.3 to 4.0 GHz.

We showed that the system was able to measure calibrated amplitude to ±2 dB and phase to ±20° over the operational band.

References

[1] B. M. Horton, "Noise-modulated distance measuring systems," *Proc. IRE*, vol. 47, no. 5, pp. 821–828, May 1959.

[2] M. P. Grant, G. R. Cooper, and A. K. Kamal, "A class of noise systems," *Proc. IEEE*, vol. 51, no. 7, pp. 1060–1061, July 1963.

[3] G. R. Cooper, and C. D. McGillem, "Random signal radar," Final Rep. TR-EE67-11, Purdue Univ. School Elect. Eng., Lafayette, IN, June 1967.

[4] J. R. Forrest and J. P. Meeson, "Solid-state microwave noise radar," in *RADAR-77 IEEE Conf.*, London, U.K., Oct. 1977, pp. 531–534.

[5] E. K. Walton, V. Fillimon, and S. Gunawan, "ISAR imaging using UWB noise radar," in *Proc. AMTA Symp.*, Seattle, WA, Sept. 30–Oct. 3, 1996, pp. 167–171.

[6] R. M. Narayanan, Y. Xu, P. D. Hoffmeyer, and J. O. Curtis, "Design and performance of a polarimetric random noise radar for detection of shallow buried targets," in *Proc. SPIE Detection Technol. Mines Minelike Targets*, Orlando, FL, Apr. 1995, vol. 2496, pp. 20–30.

[7] E. K. Walton and J. D. Young, "The Ohio State University compact radar cross-section measurement range," *IEEE Trans. Antennas Propagat.*, vol. AP-32, pp. 1218–1223, Nov. 1984.

[8] I. P. Theron, E. K. Walton, S. Gunawan, and L. Cai, "Signal-to-noise ratio calculations and measurements for the OSU noise radar," Rep. 732168-1, Ohio State Univ., ElectroSci. Lab., Columbus, OH, Nov. 1996.

[9] E. K. Walton and A. Moghaddar, "Imaging of a compact range using autoregressive spectral estimation," *IEEE AES Syst. Mag.*, vol. 6, no. 7, pp. 15–20, 1991.

[10] E. K. Walton, "The measurement of radar cross section," in *Radar Target Imaging*, Boerner/Überall, Eds. Berlin, Germany: Springer-Verlag, 1994, ch. 5.

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