radiation pattern cuts for the right- and left-hand circularly polarized wavefields plotted versus angle θ measured from the forward axis of the volute.

REFERENCES

- [1] A. T. Adams *et al.*, "The quadrifilar helix antenna," *IEEE Trans. Antenna Propag.*, vol. AP-22, pp. 173-178, Mar. 1974.
- [2] E. T. Kornhauser, "Radiation field of helical antennas with sinusoidal current," *J. Appl. Phys.*, vol. 22, pp. 887-891, July 1951.
- [3] R. W. P. King, *Fundamental Electromagnetic Theory*, 2nd ed. New York: Dover, 1963, ch. IV.
- [4] J. D. Kraus, "The helical antenna," *Proc. IRE*, vol. 37, pp. 263-272, Mar. 1949.
- [5] C. C. Kilgus, "Shaped-Conical Radiation Pattern Performance of the Backfire Quadrifilar Helix," *IEEE Trans. Antennas Propag.*, vol. AP-23, pp. 392-397, May 1975.
- [6] —, "Resonant quadrifilar helix," *IEEE Trans. Antennas Propag.*, vol. AP-17, pp. 349-351, May 1969.
- [7] —, "Multi-element, fractional turn helices," *IEEE Trans. Antennas Propag.*, vol. AP-16, pp. 499-500, July 1968.

Ultra-Wide-Band Transient Electromagnetic Scattering Laboratory

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Abstract—A unique free-field transient scattering facility is described which employs a dual-channel ultra-wide-band impulsive source. Following a brief historical review, the theory and operational characteristics of the laboratory are discussed. New technical implementations are then described which have yielded significant improvement in effective bandwidth and signal-to-noise ratio (SNR). High quality experimental validations are shown, and consideration is given to future enhancements.

I. INTRODUCTION

Electromagnetic scattering measurements are usually performed using continuous wave (CW) time-harmonic fields. When a coherent stepped-frequency mode is available, CW systems can be used to estimate transient scattering responses of linear, time-invariant (LTI) scatterers, where conversion from frequency-domain to time-domain is performed via a discrete inverse Fourier transform [1].

Scattering measurements made directly in the time-domain use an impulsive form of incident field. This approach was made possible by the advent of fast pulse sources and sampling oscilloscopes in the 1960's [2]. For non-LTI cases, when nonlinear and/or time-varying material effects are present, it is essential that direct time-domain measurements be performed. Even in the usual realm of LTI scattering, the time-domain approach can offer a faster and less expensive alternative to the coherent stepped-frequency method. Additionally, direct observation of transient responses can be advan-

tageous in testing structural alterations to reduce or enhance the response of scatterers.

The original Transient Electromagnetic Scattering Laboratory (TESL) was constructed at the Naval Postgraduate School in 1980 using a large outdoor ground plane [3]. In 1983 the TESL was redesigned using an indoor shielded anechoic chamber [4]. Another upgrade came in 1985, with the development of a new form of highly stable impulse source employing a GaAs FET power amplifier with effective bandwidth of 1-6 GHz [5]. A further 10 dB improvement in signal to noise ratio (SNR) was made in 1988, with the incorporation of a 20 GHz (HP 5412T) digital processing oscilloscope (DPO) [6].

The TESL has facilitated research into radar target identification using complex natural resonances [7], [8]. Because resonance extraction algorithms are highly sensitive to signal noise and clutter, there has been continuous motivation to improve the fidelity of TESL scattering measurements. In this paper, we describe the latest enhancement for increasing both the measurement bandwidth and the SNR of transient scattering signatures.

II. LABORATORY DESCRIPTION

The free-field TESL incorporates a shielded anechoic enclosure, having inside dimensions of 6.2 m in length by 3.1 m square, as depicted in Fig. 1. Targets are supported on a virtually transparent styrofoam column placed about 2.5 m from the antennas. The metallic shielding in the chamber significantly reduces the effects of outside interference so that it is essentially negligible compared to either thermal noise provided by the receiver system (sampling head front-end) or antenna noise generated by ambient radiation of the absorbing material on the inside of the chamber. The back-wall is composed of 46 cm long pyramidal absorber, which has a reflection coefficient of -30 dB at 500 MHz (decreasing further with frequency throughout the passband of the range), while the source-wall uses 21 cm long pyramids. The side-walls, floor, and ceiling are covered by 21 cm longitudinal wedged material, which acts to channel energy toward the back-wall, with minimal reflections.

In the sampling process each measurement point is taken from a separate incident pulse in a pulse train. Currently available commercial sampling technology has not yet advanced to the point of making single-shot waveform captures with spectra exceeding about 6 GHz. The composite waveform sampling approach requires a highly repetitive incident pulse source and a coherently triggered sampling head. Sampling noise is introduced by any stochastic nonrepetitive behavior of the pulses within the pulse train (termed *amplitude jitter*) or by random timing errors in the triggering of the sampling head (known as *timing jitter*). The DPO system has a 35 ps rise-time tunnel-diode step-generator and four sampling heads combined in one assembly (HP 54121A four-channel test set). A very high level of timing coherence and repeatability are observed since the step-generator acts both as the fundamental incident field source and as the internal trigger source for the sampling heads.

A highly stable ultra-broad-band impulse-generator has been developed [9] by using the HP step-generator to drive two parallel GaAs FET bandpass power amplifier sections, each having rated maximum power output of 1 W. One section is composed of a cascaded low-power 15 dB preamplifier (HP 8349A) and a 28 dB power amplifier (Avantek APT-12066), which together have a nominal 3 dB bandwidth of 6-12 GHz. The other section, which amplifies the 1-6 GHz passband, is made up of a special coaxial delay-line followed by a 35 dB power amplifier (Avantek APT-6065). The output of each amplifier section feeds an individual 1-2

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Fig. 1. Transient Electromagnetic Scattering Laboratory (TESL) configuration.

GHz bandwidth TEM horn antenna (AEL H-1479). Signal recombination is accomplished by superposition of the incident fields at the target location. The detailed geometry of the amplifier-antenna configuration is shown in Fig. 2.

A special frequency domain deconvolution algorithm is employed to estimate the scattering response of the measured target to a user-specified incident time-domain plane wave field. There are two constraints on specifying this incident field: namely, it must have the same polarization as is transmitted and its bandwidth must fit within the passband of the TESL system. A caveat to the latter constraint is that the power spectral density (PSD) of the composite 1–12 GHz transmitted signal must not have any deep valleys across the frequency band. A severe problem of this type was initially discovered by making measurements without the coaxial delay-line in place. This delay-line acts to temporally align the upper and lower pass signals. Since there is some spectral overlap in the 5–7 GHz region between these two signals, the preamplifier time-delay introduces coherent phase additions and cancellations between common frequency components with possible notches appearing in the composite PSD around 6 GHz. An optimum delay-line length (accurate to subcentimeter tolerance) was found through a systematic “cut and try” approach [9].

III. ACQUISITION AND PROCESSING

Operation of the TESL can be understood by referring to the system diagram shown in Fig. 3. Transfer functions in the frequency-domain represent the physical interactions in the transmitting, scattering and receiving processes, which are all assumed to be LTI in nature. The system diagram has been simplified by combining signal paths in the dual-horn transmission system and by employing scalar signal notation to represent the single vector field polarizations being transmitted and received.

The input step pulse, with Fourier transform $X(f)$, drives the parallel amplifier and transmitting antenna configuration, whose composite transfer function is denoted by $H_T(f)$. The transmitted field is incident upon the scatterer and the absorber in the chamber, which have respective transfer functions of $H_S(f)$ and $H_C(f)$. Multiple scattering between the target and the chamber structure, including absorber and antennas, is symbolized by the two-way interaction arrow between $H_C(f)$ and $H_S(f)$. Signal energy that is directly coupled from the dual transmitting horns to the adjacent receiving antenna is accounted for by the antenna coupling transfer function, $H_A(f)$. The receiving antenna transfer function, $H_R(f)$, converts the received field into the signal being sampled by the DPO. Finally, *antenna noise*, due primarily to thermal emission by the chamber absorber, and *receiver noise*, which is innate to the

Fig. 2. Dual-channel impulse source and antennas.

Fig. 3. Transfer function system representation.

DPO system, are lumped together at the DPO input as an equivalent noise source $N(f)$.

TESL measurements are used to estimate the transient field scattering response at the receiving antenna location, $e_s(t)$, due to a user-specified incident plane wave signal, $e_i(t)$, impinging upon the target. These signals represent the linearly polarized E -field vector components associated with the respective receiving and transmitting horn antennas. Denoting the Fourier transform, $E_i(f) = \mathcal{F}\{e_i(t)\}$, the desired response is given by

$$e_s(t) = \mathcal{F}^{-1}\{E_i(f) \cdot H_S^T(f)\} \quad (1)$$

where the T superscript indicates that the desired target is positioned as the scatterer in the chamber. A calibration sphere scattering measurement, having transfer function $H_S^C(f)$, will also be required.

The user-selected incident field signal, $e_i(t)$, is not actually transmitted. Using the deconvolution algorithm, $e_s(t)$ can be synthesized for any incident signal whose spectrum, $E_i(f)$, is bandlimited within the measurement passband of 1–12 GHz.

In performing the deconvolution signal processing, target scattering, as embodied in $H_S(f)$, is extracted from the combined effects of the other frequency-dependent elements of the system (e.g., amplifiers, antennas, chamber clutter, etc.). This is a multi-step procedure that requires the acquisition of three time-sampled and

ensemble-averaged waveforms: 1) target, with selected scatterer in place; 2) background, with an empty chamber, and; 3) calibration, using a metallic sphere. Usually 1024 time-sampled data points are acquired in a 20 ns time-window, each digitized to 12-bit accuracy. Shorter time-windows will be used in the waveform displays to be presented to allow expanded resolution of details. A measured waveform is typically assembled from the ensemble-average of 2048 individually sampled waveforms. Assuming uncorrelated noise, a 33 dB increase in signal-to-noise ratio (SNR) results from this process.

Example received voltage waveforms for the background and broadside backscattering from a 10 cm long brass wire, with radius 1.18 mm, are overlayed in the 14 nsec time-window of Fig. 4. Subtraction of the background from the target waveform is displayed in Fig. 5. A small subtraction remnant, occurring both before and after the transient scattering return, is quantized by a 12 bit sampling resolution when using the most sensitive DPO vertical scale factor of 2 mV/division. The background waveform is also subtracted from the calibration measurement. These background subtractions eliminate the signal components due to direct clutter and antenna coupling. As can be seen from Fig. 3, these transfer function segments are common to all three measurements. Background subtracted scattering and calibration signal transforms are given by

$$Y_S(f) = X(f)H_T(f)\{H_S^T(f) + H_{SC}^T\}H_R(f) + N_S(f) \quad (2a)$$

$$Y_C(f) = X(f)H_T(f)\{H_S^C(f) + H_{SC}^C\}H_R(f) + N_C(f) \quad (2b)$$

where $N_S(f)$ and $N_C(f)$ are the respective noise differences of target minus background and calibration minus background. The multipath clutter contributions in H_{SC}^T and H_{SC}^C are not canceled. In the time-domain, these multipath effects are delayed due to causality from the main target return and are also quite small, as is evidenced by the measured result in Fig. 5.

The next step is deconvolution, whereby the antenna transfer function effects which appear in (2) are eliminated and the target response is estimated for a specified incident field waveform. Our deconvolution process [3], [5], [9], employs an extension of the "optimal compensation deconvolution" procedure developed by Riad [10]. The target scattering spectral estimator is given by

$$E_i(f)\hat{H}_S^T(f) = \frac{Y_S(f)Y_C^*(f)}{|Y_C(f)|^2 + L}\{E_i(f)H_S^C(f)\} \quad (3)$$

with the estimated time-domain target signature, $\hat{e}_s(t)$, obtained by an inverse FFT of this function. The smoothing parameter L , is selected so that the effective bandwidth of the deconvolved response in (3) just falls within that of the measurement system.

Computation of the calibration sphere scattered field, as produced by the specified incident waveform, is denoted by $E_i(f) \cdot H_S^C(f)$ in (3). This computation is performed using a stepped-frequency Mie series, whereby the complex phasor scattered field at the location of the receiving antenna is evaluated. To increase accuracy, far-field approximations are *not* employed and the small bistatic angle between the transmitting and receiving antennas is incorporated.

The selected incident field waveform is composed of a fast, positive-polarity Gaussian function and a slower, negative-polarity Gaussian, having the composite double-Gaussian (DG) form

$$e_i(t) = A_1 e^{-\alpha_1(t-t_0)^2} - A_2 e^{-\alpha_2(t-t_0)^2} \quad (4)$$

This waveform allows control of both lower and upper frequency rolloffs so that only a very small percentage of the energy in $E_i(f)$

Fig. 4. Measured antenna voltages for thin-wire target (broadside incidence) and background with empty chamber. Showing 716 point time-samples using 2048 ensemble averages. Brass thin-wire length = 10 cm, radius = 1.18 mm.

Fig. 5. Target signal with background subtracted for the 10 cm thin-wire of Fig. 4.

appears outside of the 1–12 GHz measurement band. Amplitudes and decay constants in the DG waveform are also chosen to provide a peak amplitude of 10 V/m and 0 dc, with $E_i(0) = 0$. The DG waveshape is illustrated in the figure inserts of the validations in the following section.

IV. VALIDATIONS AND CONCLUSIONS

Transient backscattering waveforms for the 10 cm long thin-wire, due to the DG incident pulse, are displayed in Figs. 6 and 7 for respective aspect angles of 90° (broadside) and 30° from end-on. Results of two alternative numerical computations are also shown in each of these figures. One of these is based on a thin-wire time-domain integral equation (TDIE) formulation, which time-marches both spatially discretized current and charge, [11]. The other computation is produced by the inverse FFT of a stepped frequency computation based on Hallén's frequency-domain integral equation (FDIE), [12]. A short 3 ns time-window is used to show detailed comparisons of the significant portions of the waveforms. The agreement in Fig. 6 between 90° case measurements and computations (particularly FDIE) is quite good, even far into the late-time.

The comparisons in the 30° aspect case of Fig. 7 are quite good on average, although not as good as for broadside. First of all, the induced backscattering is much smaller for the 30° canted wire so

Fig. 6. Comparison of measured deconvolved DG transient 90° backscattering with TDIE and FDIE computations for a 10 cm long thin-wire with radius 1.18 mm.

Fig. 8. Comparison of measured deconvolved DG transient backscattering and Mie series based computation for a 4.05 cm radius metal sphere.

Fig. 7. Comparison of measured deconvolved DG transient 30° backscattering with TDIE and FDIE computations for the 10 cm long thin-wire with radius 1.18 mm.

Fig. 9. Comparison of measured deconvolved DG transient backscattering and Mie series based computation for a 6.14 cm radius metal sphere.

that fixed precision computational or signal processing errors appear relatively larger. In addition, there are errors in both computations for the very early backscattering response of the canted wire. This small return is primarily due to scattering from the first illuminated end of the wire. Although different thin-wire approximations are used for the TDIE and FDIE computations, neither are accurate in their consideration of induced charges on the ends of the wire. The initially induced traveling wave current flows away from the illuminated end and produces a large scattered field about the forward scattered direction, which is unobserved here. The large impulsive "doublet" in the backscattering is produced after a round trip time delay by reflected current from the far end of the wire. A second round trip of the induced traveling wave current, accompanied by dispersive radiation damping, results in the second, and smaller, backscattered impulse. Agreement between measurement and computation appears to degrade beyond the third impulse. The reason for this is unclear, although part of the difference may be due to end-effect errors in the computations caused by the thin-wire approximations used.

A second set of fidelity tests for the TESL involves measured transient backscattering from small aluminum spheres, having radii of 4.05 and 6.14 cm. Respective time-domain comparisons with computations for DG illuminated metal spheres are shown in Figs. 8 and 9. A short 3 ns time-window is again used. The time-domain

computation is produced by an inverse FFT of the stepped-frequency phasor scattered field provided by a Mie series formulation. Both the main specular reflection of the DG incident field and the first creeping wave return are apparent in the measured and computed results for both of these spheres.

There is a continuing quest to improve both the SNR and the frequency bandwidth of TESL measurements. In the realm of increased bandwidth, the present dual amplifier system is employing essentially the entire available spectrum of the low-power DPO step-generator which acts as the basic source for the transmitter. This low-level pulse source is the current bottleneck to increased bandwidth, since the DPO operates to 20 GHz and a third amplifier-antenna section could be added to accommodate the 12 to 20 GHz band. *Highly stable* fundamental pulse sources having bandwidths much exceeding 12 GHz do not appear to be commercially available, although their development will eventually come about as technology progresses. A simple expedient to extend the effective bandwidth of any scale model range is to employ a succession of scaled replicas of the targets.

Increased SNR for the TESL is currently being pursued by replacing each of the power amplifiers with a newer version having double the peak output power and higher gain. This is not envisioned as a major improvement for most targets since the ensemble averaging procedure presently provides very high SNR, as is evidenced by the results shown. Higher transmitted power will be

important for the case of targets having low innate backscattering, such as very small objects or larger targets employing broadband low-observable characteristics. Additional improvements in processed signal fidelity will be investigated by tests involving alternate deconvolution procedures [13].

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REFERENCES

- [1] J. D. Young, "Radar imaging from ramp response signatures," *IEEE Trans. Antennas Propagat.*, vol. AP-24, pp. 276-282, May 1976.
- [2] J. D. DeLorenzo, "A range for measuring the impulse response of scattering objects," in *1967 NEREM Rec.*, vol. 9, Nov. 1967, pp. 80-81.
- [3] C. W. Hammond, "The development of a bistatic electromagnetic scattering laboratory," M.S. thesis, Elect. Eng. Dept., Naval Postgraduate School, Dec. 1980.
- [4] M. A. Morgan, "Time-domain scattering measurements," *IEEE Antennas Propagat., Newsletter*, vol. 6, Aug. 1984.
- [5] M. A. Morgan and B. W. McDaniel, "Transient electromagnetic scattering: Data acquisition and signal processing," *IEEE Trans. Instrum. Meas.*, vol. 37, pp. 263-267, June 1988.
- [6] S. Sompaee, "Computer algorithms for measurement control and signal processing of transient scattering algorithms," M.S. thesis, Elect. Comput. Eng. Dept., Naval Postgraduate School, Dec. 1989.
- [7] M. A. Morgan, "Scatterer discrimination based upon natural resonance annihilation," *J. Electromagn. Waves Appl.*, vol. 2, no. 5/6, pp. 481-502, 1988.
- [8] P. D. Larison, "Evaluation of system identification algorithms for aspect-independent radar target identification," M.S. thesis, Elec. Comput. Eng. Dept., Naval Postgraduate School, Dec. 1989.
- [9] N. J. Walsh, "Bandwidth and signal-to-noise ratio enhancement of the NPS transient electromagnetic scattering laboratory," M.S. thesis, Elec. Comput. Eng. Dept., Naval Postgraduate School, Dec. 1989.
- [10] S. M. Riad, "Impulse response evaluation using frequency-domain optimal compensation deconvolution," presented at 23rd Midwest Symp. Circuits Syst., Aug. 1981, Univ. Toledo, Toledo, OH.
- [11] E. P. Sayre and R. F. Harrington, "Time-domain radiation and scattering by thin-wires," *Appl. Sci. Res.*, vol. 26, pp. 413-444, 1972.
- [12] R. S. Elliott, *Antenna Theory and Design*. Englewood Cliffs, NJ: Prentice-Hall, 1981.
- [13] F. I. Tseng and T. K. Sarkar, "Deconvolution of the impulse response of a conducting sphere by the conjugate gradient method," *IEEE Trans. Antennas Propagat.*, vol. AP-35, pp. 105-109, Jan. 1987.

Sidelobe Performance in Quadratic Phase Conformal Arrays

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Abstract—Sidelobe performance in an array with a quadratic phase distribution is related to the array curvature and conditions are derived that insure well-behaved sidelobes for quadratic phase conformal arrays. A similar condition is derived for parabolic shaped arrays and results

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are given to illustrate the validity of the derived requirement for well-behaved sidelobes for both parabolic and circular arrays.

I. INTRODUCTION

It is well known that the behavior of sidelobes in a conformal array is dependent upon the array geometry. For certain canonical array geometries such as linear, circular, spherical, and cylindrical arrays, the far-field patterns can be formulated in analytically tractable expressions [1]. However, for other nonlinear array geometries, the far-field patterns are not conveniently calculated in a tractable form and the sidelobe performance for these arrays must be completed numerically [2]. In this communication we show that sidelobe performance in a quadratic phase conformal (QPCA) array (an array with a quadratic phase distribution across the aperture) is related to the array curvature and derive conditions that will insure well-behaved sidelobes for quadratic phase conformal arrays. Also, the sidelobe performance of a linear array with nonuniform spacing is related to the curvature of an equivalent conformal array.

The primary reason for analyzing QPCA's is that the analysis provides immediate insights into the relationship between array curvature and sidelobe performance. Using the expression for phase as a function of arclength derived in (2) below, an explicit expression for a QPCA geometry can be derived. This expression is somewhat complicated (magnitude of the tangent vectors lie on an arc of $\sin^{-1}$) and, as such, conformal arrays have not been designed with quadratic phase geometry.

However, the results derived from QPCA's apply to more general array configurations. By expanding phase as a function of arclength in a Taylor series, all array geometries are seen to exhibit near quadratic phase characteristics for sufficiently limited aperture size (i.e., when the third-order series terms become insignificant). In particular, the parabolic array is shown to have quadratic phase when phase is expressed in Cartesian coordinates rather than as a function of arclength and the circular array has near quadratic phase for limited aperture size. In both the circular and parabolic cases, the results derived from QPCA's are shown to successfully predict sidelobe degradation as a function of array curvature.

Consider a conformal array defined on the curve $(x, y(x))$ in the two-dimensional plane where y is a continuously differentiable function of x . For simplicity we will assume that $y(0) = 0$ and that the antenna elements are omnidirectional. Assume that a plane wave is impinging on the array at an angle θ with respect to the positive y -axis. The phase of the plane wave at $(x, y(x))$ relative to the phase at $(0, 0)$ is given by

$$p_\theta(x, y) = \frac{2\pi}{\lambda} (x \sin \theta + y \cos \theta)$$

where λ is the wavelength of the propagation signal. In order to analyze sidelobe degradation, we will assume that the plane wave signal impinges at $\theta = 0^\circ$. The phase of the electrical pattern at $(x, y(x))$ for an incident wave impinging at an angle θ given that we have a signal wave impinging at 0° is given as the difference of the respective phases (i.e., the beam is formed to maximize the pattern responses at 0°).

$$p(x, y) = p_\theta - p_0 = \frac{2\pi}{\lambda} (x \sin \theta + y(\cos \theta - 1)). \quad (1)$$

The curve y can be expressed independent of any coordinate frame by defining y in terms of its tangent as a function of the arclength parameter.

$$\alpha(s) = \tan^{-1}(y')$$