

Time-Domain Electromagnetics and Its Applications

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Invited Paper

Abstract—The purpose of this paper is to introduce the reader to the elements of time-domain electromagnetics, which includes baseband-pulse technology and target-signature analysis. Baseband pulses are video or carrierless pulses of very short duration, whose spectral content is concentrated primarily from zero frequency through the microwave region of the spectrum. Work in baseband technology began more than ten years ago at the Sperry Research Center with emphasis primarily on its use as an analytical tool, initially to explore the properties of microwave networks [1], and subsequently to determine the intrinsic properties of materials [2]. The experimental phases of these studies were aided by the pioneering development by the Hewlett-Packard Company of their sampling oscilloscope [3]. These techniques were further extended to experimentally analyze and synthesize antenna radiating and receiving elements [4], [5]. Success in the antenna area led to the development of an indoor ground-plane scattering range to measure the impulse properties of targets or obstacles [6]. This type of range obviated the need for an expensive anechoic chamber since time-gating techniques permitted unwanted reflections from the walls and ceilings to be easily eliminated. A sampling oscilloscope and an instrumentation computer were used to process target-signature data as a function of illumination angle. At about the same time, analytical techniques were developed which permitted target-signature analysis to be carried out efficiently and accurately in the time domain [7].

Applications where threshold rather than target-signature data is sufficient were also investigated, and for these applications it became clear that because of cost, the sampling oscilloscope had to be replaced by a stable and fast-acting threshold device. To meet this requirement, two different types of tunnel-diode receivers were developed. The successful development of these receivers, together with the design of inexpensive microwave delay-line ranging techniques, led to the evolution of Baseband Radar (BAR) or free-space time-domain reflectometry. BAR devices have been designed and recently demonstrated for various applications, including auto precollision sensing, spaceship docking, airport surface-traffic control, auto braking, tanker-ship docking, harbor-collision avoidance, etc. These sensing applications cover ranges from 5 to 5000 ft [8].

In other applications, the development of high-speed subnanosecond logic has impacted the computer field, making higher speed computation possible. Further developments resulted in the construction of a subnanosecond, single coaxial cable scheme for multiplexing data between computer terminals [9]. More recently, baseband-pulse techniques have been applied to the problem of developing a short-range wireless communication link. Here, the low EM pollution and covertness of operation potentially provide the means for wireless transmission without licensing.

We review the research areas described above in more detail and refer the reader to references and a comprehensive bibliography where sources for detailed information can be found [10].

I. WAVEFORM GENERATION TECHNIQUES

WORK IN TIME-DOMAIN electromagnetics began in 1962 when attempts were made to verify the analytical solution of the transient behavior of the class of TEM-mode microwave networks whose impulse response, $h(t)$,

is a train of impulses; that is,

$$h(t) = \sum_{k=1}^{\infty} A_k \delta(t - \tau_k) \quad (1)$$

where A_k is a weighting constant, and τ is a time delay depending on the geometry of the network. Networks of this type include the familiar branch-line coupler, ring hybrid, propagation through certain stratified dielectric media, etc. [1]. Because of the particular form of (1), certain important observations can be made.

When the impulse response of a linear time-invariant network consists of a train of impulses of varied weighting, the response to an input $f(t)$ may be found conveniently by appropriately weighting and delaying replicas of $f(t)$. In particular, if the impulse train is uniformly spaced $\lambda_0/2c$ apart, the response to a step-modulated signal of the form

$$f(t) = \sin \omega_0 t \cdot U(t) \quad (2)$$

where

$$U(t) = \begin{cases} 1, & t > 0 \\ 0, & t < 0 \end{cases}$$

$$\omega_0 = 2\pi f_0 = 2\pi c/\lambda_0$$

can be found throughout any given half-period by a simple addition. This follows since the phase difference between weighted replicas of the input signal is either zero or 180° .

Unfortunately, we found that we could not purchase a generator having an output waveform described by (2). The problem was solved by generating the required waveform ourselves, synthetically. That is, as shown in Fig. 1, we first generated a step function using a tunnel-diode source (rise time < 100 ps) and fed this into a short-circuited stub junction of length L , having one-half the characteristic impedance of the main line. The result was a pulse of $2L/c$ s duration, where c is the speed of light in the medium, and 0.5 V amplitude. By feeding this signal to an identical short-circuited stub, the resulting waveform became ± 0.25 V in amplitude and resembled one square cycle of an RF carrier of period $4L/c$ s. Next, feeding this output to a series of open-circuited stubs of lengths $2L$, $4L$, $\dots$, NL , we produced 2 , 4 , $\dots$, N square cycles of microwave energy, provided the stub separation was more than NL ft. The measuring instrument (sampling oscilloscope) behaved as a low-pass filter, rounding off the sharp corners of the wave, and making the resulting waveform indistinguishable from the desired test waveform of (2). The disadvantage of this technique was that the output voltage V_0 , was reduced by the factor

$$V_0 = 2^{-k} \quad (3)$$

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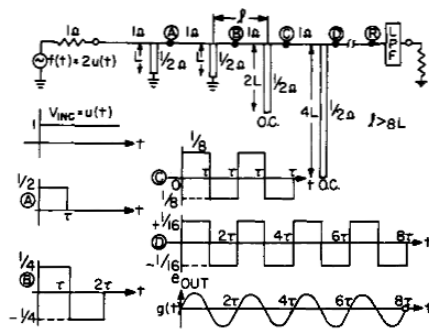


Fig. 1. Synthetic generation of wave energy.

TABLE I

Type	Step/Pulse	Best Available Rise-time at Amplitude		Notes
Mercury switch	Step	70 ps	100 V	Max. PRF = 200 Hz
Avalanche transistor	Pulse	150 ps	12 V	Device selection necessary
Tunnel diode	Step	25 ps 100 ps	0.25 1.0	Fastest transition time
Step recovery diode	Step	60 ps 225 ps 200 ps	20 V 50 V 200 V	commercially available specially ordered 4-stack
Hertzian	Impulse; also pulse modulated	100 ps 1 ns	1000 V 1000 V	Limited lifetime, spark gap
Avalanche diode	Impulse	400 ps	125 V	GHz rep. rate

where k is equal to the number of stubs employed. This limits to 8 or 16 the number of cycles that can be practically generated [11]. Later inventions by Proud [12] and Cronson [13] obviated the problems of loss by the use of line lengths separated by spark-gap devices or step-recovery diodes, respectively.

These experiments were important because they showed that virtually any frequency-coded pulse could be generated synthetically from a simple unit step function; a unipolar baseband pulse was simply the addition of two opposite-polarity step functions shifted in time. These simple observations stimulated the work that followed in the antenna and metrology areas of investigation.

Improvements in solid-state devices since these initial experiments have resulted in significant improvements in both the risetime and amplitude of generated step and pulse functions. Table I summarizes the state of the art in pulse-generation techniques [14]. In the sections that follow, the use of these waveform sources will be demonstrated.

II. TIME-DOMAIN METROLOGY

It is well known that a linear time-invariant network can be completely specified by its impulse response. However, it is only in the last decade, with the advent of the sampling oscilloscope and the baseband-pulse generator having a subnanosecond risetime, that the transient response of distributed networks and materials could be directly observed. These time-domain waveforms provided a fresh approach toward understanding the behavior of various networks and junctions. Since each network discontinuity contributed a separate return, the resultant display was generally amenable to simple physical interpretation. Early work in time-domain reflectometry verified this correspondence between transient network



Fig. 2. Time-domain measurement system.

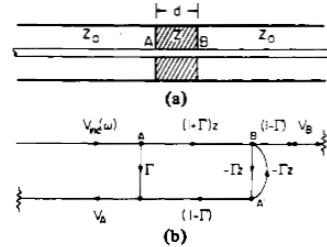


Fig. 3. (a) A coaxial line with annular disk of material to be measured and inserted. (b) Signal-flow graph for (a).

analysis and experiment [1]. Davis and Loeb [15] were among the first to demonstrate that time-domain waveforms could be used to obtain wideband frequency-domain information of two-port networks to 1 GHz. More recently, the convenience and accuracy of time-to-frequency translations has been greatly improved through the use of computer-controlled scanning procedures and fast Fourier transform techniques. A time-domain technique for measuring the scattering coefficients of microwave networks to 5 GHz was described by Nicolson in 1968 [16]. Later, an improvement of this system was reported [17] where the computer was used not only for averaging and transform computation, but also to monitor and dynamically correct for errors due to additive voltage drift and timing jitter.

Time-domain metrology has also proved a useful tool for determining the electrical properties of materials over large frequency ranges. In 1969 Fellner-Feldegg [18] sparked interest among physical chemists by reporting on a reflection method for determining the dielectric constant and relaxation time of various liquids. Independently, in 1969 Bagozzi [19] developed the theory for polar liquid dielectric coaxial lines, including the effects of skin-effect losses. Concurrently, Bagozzi *et al.* [20] applied the theory to the measurement of the dielectric relaxation time in Debye binary liquids using a time-domain transmission method. At the 1970 Conference on Precision Electromagnetic Measurements, Nicolson and Ross [21] presented a new method utilizing reflected and transmitted waves to obtain the complex permittivity and permeability of linear materials over a broad range of microwave frequencies. This work demonstrated the feasibility and practicality of time- to frequency-domain translation.

Further work in this area has been directed toward improving measurement accuracy, exploring various coaxial configurations, introducing more automation, increasing dynamic range, and extending both high- and low-frequency bounds [22]. In a recent paper, for example, Cronson *et al.* described a scheme for increasing the upper frequency of a materials measurement technique to 16 GHz [23]. The configuration used for the measurements is shown in Fig. 2. The pulse from the pulse generator typically has a risetime of less than 50 ps. It propagates down the coaxial delay line, through the feed-through sampling head of the oscilloscope, and is incident on the sample of material in the test-sample holder. The pulse is partially reflected and partially transmitted through the sample. The transmitted pulse reflects from a short-circuit, returns up the line, and passes again through the sample on its way to

the oscilloscope sampling head. It can be shown that these waveforms yield $S_{11}(\omega)$ and $S_{21}(\omega)$, where S_{21} and S_{11} are the scattering coefficients of the slab [2], [21].

The basic principle of the measurement can be defined with the aid of Fig. 3. Consider an annular disk of material with permeability $\mu = \mu_0 \mu_R$, permittivity $\epsilon = \epsilon_0 \epsilon_R$, and thickness d , installed in a coaxial air-filled line with characteristic impedance Z_0 as shown in Fig. 3(a). Within the region $0 \leq X \leq d$ the line will assume a new characteristic impedance $Z = \sqrt{\mu_R/\epsilon_R} Z_0$, where μ_R and ϵ_R may be complex. If d were infinite, then the reflection coefficient of a wave incident on the interface from the air-filled line would be given simply by

$$\Gamma = \frac{Z - Z_0}{Z + Z_0} = \frac{\sqrt{\mu_R/\epsilon_R} - 1}{\sqrt{\mu_R/\epsilon_R} + 1}. \quad (4)$$

For finite d , the transmission coefficient between faces A and B of the slab may be written

$$z = \exp[-j\omega\sqrt{\mu\epsilon}d] = \exp[-j(\omega/c)\sqrt{\mu_R\epsilon_R}d]. \quad (5)$$

The scattering coefficients S_{21} and S_{11} of the slab may be obtained from the flow graph, Fig. 3(b), namely,

$$S_{21}(\omega) = \frac{V_B}{V_{inc}} = \frac{(1 + \Gamma)(1 - \Gamma)z}{1 - \Gamma^2 z^2} = \frac{(1 - \Gamma^2)z}{1 - \Gamma^2 z^2} \quad (6)$$

$$S_{11}(\omega) = \frac{V_A}{V_{inc}} = \frac{(1 - z^2)\Gamma}{1 - \Gamma^2 z^2}. \quad (7)$$

Writing the sums and differences of the scattering coefficients as

$$\begin{aligned} V_1 &= S_{21} + S_{11} \\ V_2 &= S_{21} - S_{11} \end{aligned} \quad (8)$$

and if

$$X = \frac{1 - V_1 V_2}{V_1 - V_2}$$

then it can be shown that Γ may be obtained from the scattering coefficients, since,

$$\Gamma = X \pm \sqrt{X^2 - 1} \quad (9)$$

and the appropriate sign is chosen so that $|\Gamma| < 1$. Also,

$$z = \frac{V_1 - \Gamma}{1 - V_1 \Gamma}. \quad (10)$$

Now from (4) we define

$$\frac{\mu_R}{\epsilon_R} = \left(\frac{1 + \Gamma}{1 - \Gamma} \right)^2 = c_1 \quad (11)$$

and from (5) we define

$$\mu_R \epsilon_R = - \left\{ \frac{c}{\omega d} \ln \left[\frac{1}{z} \right] \right\}^2 = c_2. \quad (12)$$

Then,

$$\begin{aligned} \mu_R &= \sqrt{c_1 c_2} \\ \epsilon_R &= \sqrt{\frac{c_2}{c_1}}. \end{aligned} \quad (13)$$

Complex permeability and permittivity are thus obtained from measurement of the transmission and reflection scattering coefficients of a slab of the material. The delay-line lengths are

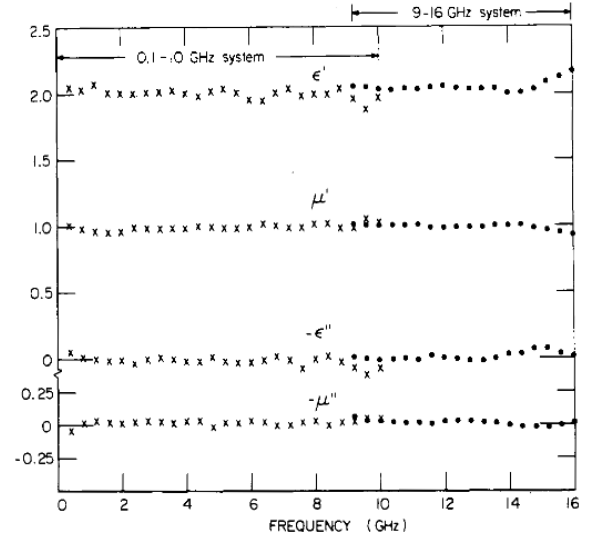


Fig. 4. Relative permittivity ϵ^* and permeability μ^* for 5.0-mm-thick Teflon.

chosen sufficiently long so that at the sampling head the incident, the reflected, and the doubly-transmitted transient-voltage waveforms do not overlap. Consideration of the lengths shown in Fig. 3 shows that the reflected and transmitted waveforms each have a "time window" over which they may be measured without contamination by other reflections of $2L/c$ s. Typical results of this technique for the measurement of Teflon are shown in Fig. 4. The use of time-domain techniques to further evaluate the properties of microwave components was presented by Cronson and Mitchell in December 1973 [24]. Additional work may be found in referenced U.S. Government reports [10].

III. ANTENNAS

An understanding of the time-domain behavior of antennas is important because processes in nature generally do not follow sinusoidal patterns but are impulsive in their time dependence. Also, in modern radar applications very broadband signals (i.e., signals having an effective short-pulse duration) are being used for target identification. The effect of short-pulse signals in antenna systems becomes observable wherever the propagation delay of the signal on the structure is comparable to the rise time or duration of the pulse. When this situation occurs, only certain segments are involved in the radiation process at any one time. For example, consider the radiation field of a dipole of total length $2h$, diameter $2a$, and driven by a generator of internal impedance Z_g . If $h \gg a$, so that the creeping waves over the circumference of the wire can be neglected, then it can be shown that in the boresight direction the impulse response of such a structure is a train of impulses of varying weight (see (1)) and of spacing h/c s, where c is the speed of light [25]. Furthermore, if the length of the antenna is very long, then over any given time interval the impulse response can be represented as a single impulse, which means that any transmitted waveform of duration less than the impulse spacing can be reproduced without distortion. Once the radiation response of an element is known, the response as a receiving antenna can also be found from the Rayleigh-Carson relationship; in essence the current induced in the antenna is

equal to the integral of the time history of the radiation field [25].

The time-domain approach to antenna theory provides a significant degree of insight that is not obvious from the conventional frequency-domain approach. For example, the time-domain approach suggests new criteria for the design and synthesis of wide-band-radiating elements [26]. Conventionally, to minimize distortion one requires that the system function (i.e., the Fourier transform of the impulse response) of an antenna have a flat amplitude spectrum and a linear phase function over a given band, with minimum specifications outside the band. The difficulty here is that the in-band data can be used to predict the gross response of the system to a step- or pulse-modulated input, but cannot be used to accurately predict the settling time or time-domain residues, respectively. The fine structure of the response depends on the out-of-band characteristics.

Specifications in the time domain can be written to conveniently describe and constrain the impulse response of an antenna. In particular, the duration of the impulse response can be constrained to precisely τ seconds. Devices of this type can be described as having a time-limited impulse response (TLIR). In general, systems having a TLIR have interesting mathematical properties. For example, their Laplace transform is an entire function having no poles in the entire complex p plane. Practically, the duration of the transient for a step-modulated excitation of the form

$$f(t) = \sin 2\pi f_0 t \cdot U(t) \quad (14)$$

cannot exceed the duration of the impulse response. For a pulse-modulated input of the form

$$p(t) = \sin 2\pi f_0 t \{U(t) - U(t - nT)\} \quad (15)$$

where $T = 1/f_0$ and n is an integer, the total dispersion is limited to twice the duration of the impulse response. There are no time-domain residues introduced by the antenna. If the duration of the impulse response $h(t)$ is less than $T/2$, then signal distortion is limited to less than half of an RF cycle at the resonant frequency f_0 .

The shape of $h(t)$ determines the distortion introduced during the duration of the transient, and the magnitude of the steady-state sinusoidal response for $t > \tau$ seconds. As an example, consider a causal system having a TLIR of duration τ given by

$$h(t) = \delta(t) + \delta(t - \tau) \quad (16)$$

where $\delta(t)$ is a delta function. The response to a step-modulated input described by (14), where $f_0 = \frac{1}{2}\tau$, is given by

$$r(t) = \begin{cases} \sin \frac{\pi}{\tau} t, & 0 \leq t \leq \tau \\ 0, & t > \tau. \end{cases} \quad (17)$$

Steady state is reached after τ seconds. If the sign of the second impulse is reversed, then the response is given by

$$r(t) = \begin{cases} \sin \frac{\pi}{\tau} t, & 0 < t < \tau \\ 2 \sin \frac{\pi t}{\tau}, & t > \tau. \end{cases} \quad (18)$$

The system "settles" to its steady-state response after τ s, but now $r(t)$ is sinusoidal for $t > \tau$.

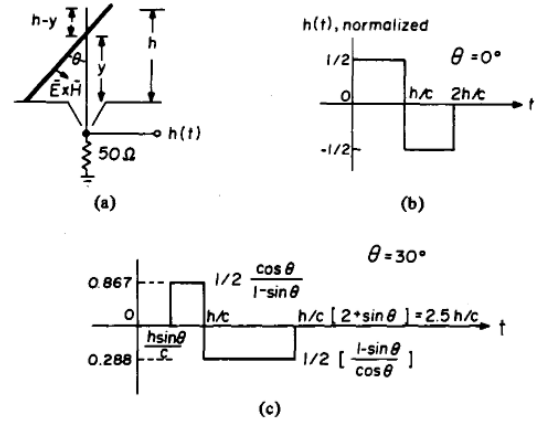


Fig. 5. Impulse response of a matched dipole. (a) Configuration. (b) Impulse response, $\theta = 0^\circ$. (c) Impulse response, $\theta = 30^\circ$.

As a second and more pertinent example, consider the thin vertical monopole mounted above a ground plane shown in Fig. 5. Here the driving-point surge impedance of the monopole has been approximately matched at the base by using a long tapered conical TEM-mode section beneath the ground plane. A detailed description of this element appears in the literature [4]. The impulse response in the receive mode for such an antenna is shown at boresight and for an angle of incidence of 30° in Fig. 5(b) and (c), respectively. The response of this antenna to a step-modulated signal described by (14), where $f_0 = \frac{1}{2}\tau$, can be found by convolution. The experimental result obtained with the waveform generator described earlier is shown in Fig. 6 and indicates that the buildup time of the response is less than one RF cycle, as predicted.

In addition to the matched dipole we have also built a so-called wave-trapping type (WTT), characterized by a short wire, length h , and a thin metal top hat. There is also a matched-monopole type (MMT), where h is relatively long, and the wave is matched at the tip by a metallic film of submicrometer thickness deposited on a thin glass plate [5]. Both antennas have a TLIR. The WTT has also been used very effectively as a wide-band probe in a cavity. Other members of the class of antennas having essentially a TLIR are used in the many baseband radar applications described in Section V.

IV. TIME-DOMAIN SIGNATURE ANALYSIS

With the advent of subnanosecond technology, it became clear that both electromagnetic-signature measurements, by virtue of hardware advancements, and electromagnetic-signature computations, by virtue of digital computer advances, were possible directly in the time domain. The measurements or calculations are carried out by illuminating the target with a smoothed-impulse waveform. An example of a smoothed impulse is a Gaussian regularization of an impulse given by

$$e(t) = \frac{a_n}{\sqrt{\pi}} \exp -a_n^2 t^2$$

which converges to a delta functional as a_n goes to infinity. The resulting smoothed-impulse response contains all the information about the electromagnetic scattering properties of the target over the frequency band that is defined by the spectrum of the incident smoothed impulse. The utility of using

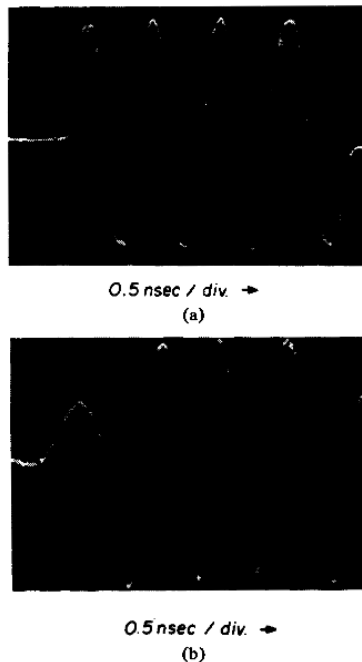


Fig. 6. Pulse-modulated response, actual, $\theta = 0^\circ$. (a) Transmitted pulse. (b) Received pulse.

the impulse response as a characteristic signature was noted in 1965 [27].

The development and perfection of the technology necessary to obtain the transient solution to electromagnetic-scattering problems directly in the time domain has coincided with a growing interest in wide-band radar design for higher resolution and target-classification problems. There has also been great interest in reducing the radar cross section of targets, such as aircraft, over a wide frequency band by placing radar absorbing materials in the vicinity of target scattering centers. The electromagnetic-pulse (EMP) problem associated with nuclear explosions, which has been the object of much attention recently, is also an area which is essentially a transient problem.

Having assessed the conception, advantages, and applications of the time-domain approach to obtaining radar target signatures, the remainder of this section addresses the fundamental concepts of time-domain signature analysis. The configuration and measurement characteristics of the time-domain scattering range are described first, followed by a generalized approach to deriving the space-time integro-differential equations which describe the scattered field at any point in space. Specific space-time integro-differential equation solutions are then developed for several representative target geometries. Next, the impulse-response-augmentation technique, a means of obtaining an estimate of the impulse response from the smoothed-impulse response, is discussed. Finally, inverse scattering, in which the radiated and received waveforms are employed to determine target geometry, is described.

A. Time-Domain Scattering-Range Technology

The conventional approach to determining target signatures is to illuminate the target with a single-frequency waveform and measure the amplitude and phase of the return [28]. While this approach provides narrow antenna beamwidths and high

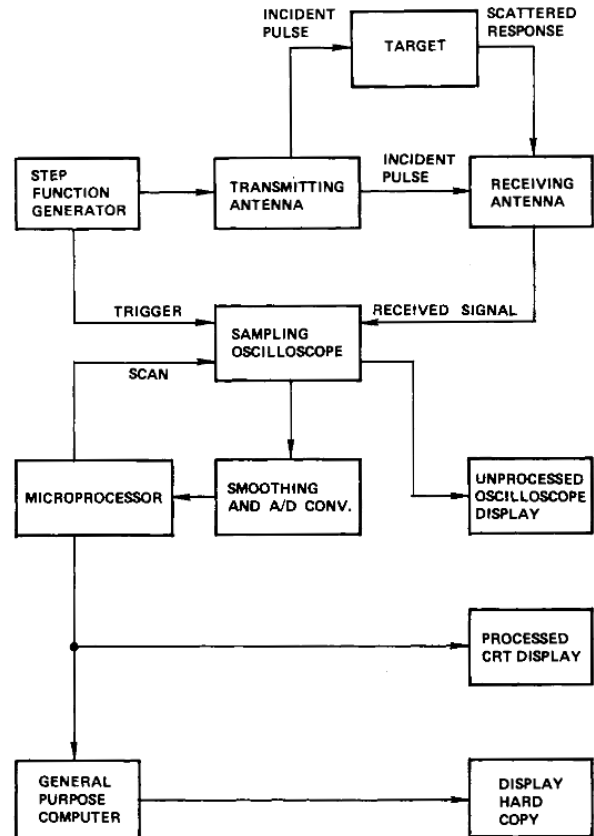


Fig. 7. Functional block diagram of video time-domain scattering range.

signal-to-noise ratios (SNR), extraneous returns from objects in the vicinity of the target can only be eliminated by placing the scattering range in an expensive anechoic chamber. To obtain the broadband or transient responses, the measurements must be repeated over the entire band of frequencies contained in the spectrum of the transient excitation.

The cost of an anechoic chamber and the difficulty in making coherent measurements over a broad frequency band, coupled with advances in subnanosecond-pulse technology, led to the development in 1967 of a scattering range that measures the transient response directly in the time domain by illuminating the target with a smoothed-impulse waveform [6]. The resulting smoothed-impulse response is useful for several reasons [29]–[35]. First, it contains all the information about the electromagnetic scattering properties of the target over the frequency band defined by the incident smoothed impulse. Since the impulse response is closely related to the actual target geometry, it is particularly useful for target classification. Second, the radar cross section (or, equivalently, the frequency response) can be obtained from the impulse response by a Fourier transform. Third, the response of the target due to any waveform can be obtained from the impulse response by a simple convolution procedure. For the purpose of target classification, the impulse response is a particularly useful characterization since it is closely related to the actual target geometry. Finally, the impulse response can provide a better understanding of electromagnetic scattering phenomena. This

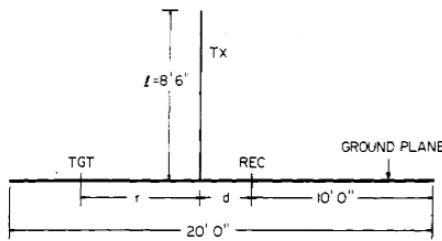


Fig. 8. Geometrical configuration of video time-domain scattering range ($d = 18'$, $r = 36'$).

section describes the basic operating principle of a typical time-domain scattering range [32].

A functional block diagram of the time-domain scattering range is shown in Fig. 7. The system signal source is a high-voltage switch which generates a 300-V step function with a rise time less than 100 ps. The signal is radiated from a vertical monopole antenna protruding through a circular ground plane 20 ft in diameter. The ground plane provides isolation from equipment and cabling that is necessary for its operation. Since the ground plane acts as an electromagnetic mirror, its application is limited to the measurement of targets with mirror symmetry. The radiated wave is then scattered by a target and received on a flush-mounted coaxial horn antenna, which essentially smoothes and differentiates the signal, and thus provides the smoothed-impulse response of the target. The received waveform is sampled by a sampling oscilloscope that has been triggered by the initial pulse and whose sampling-gate deflection is under the control of a microprocessor. Unprocessed data are displayed on the oscilloscope CRT while the sampled-and-held waveform is passed through a low-pass filter, digitized, read into the microprocessor, and stored on magnetic tape automatically. This system has been designed to correct for long-term timing drift and/or amplifier drift. In addition, the waveforms are stored in such a way that they are ready for the subsequent operations of averaging (to remove short-term noise) and baseline processing. The effects of a time-varying baseline are subtracted from measured waveforms to improve system accuracy. Finally, the microprocessor is connected to a general-purpose computer for more sophisticated and/or more rapid signal processing.

The salient characteristics of the range are the speed and simplicity with which transient responses (or, equivalently, broadband frequency-domain data) can be obtained. These advantages accrue because the time-domain scattering range yields an "uncontaminated" interval of time between the arrival of the direct wave and the arrival of unwanted reflections. The sketch in Fig. 8 shows the relative location of the elements on the ground plane, and the photographs in Fig. 9 show the range response as it appears at the oscilloscope (no data processing has been used at this point). The transmitted signal travels outward from the base of the wire antenna and is received at R at time $t_0 = d/c$ (where c is the speed of light). This time is marked by the pulse at the left end of the trace in Fig. 9. The outgoing wave reaches the target at $t = r/c$, is reflected, and arrives at the receiver at $t_1 = (2r + d)/c = t_0 + (2r/c)$. The targets are usually located anywhere from 2 to 5 ft from the transmitting antenna; therefore, target returns lie in the region marked by the doublets at t_1 and t_2 in the lower photograph. The response at the right edge of the trace which occurs after t_3 marks the arrival of the pulse reflected by the table edge and the effects of the pulse radiated from the tip of

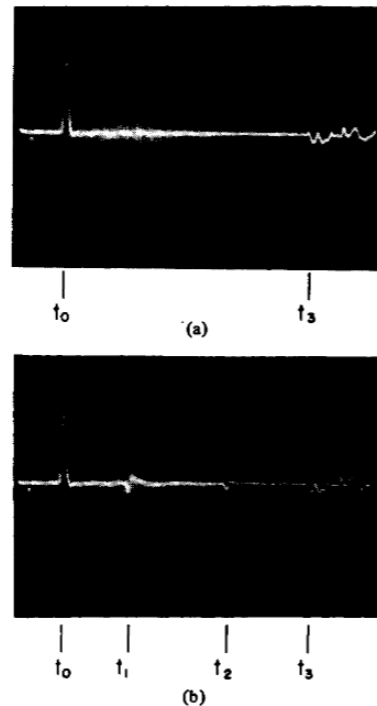


Fig. 9. Response of video time-domain scattering range showing incident pulse and time window (horizontal scale: 2 ns/div; vertical scale: 200 mV/div.). (a) Incident pulse with clear range. (b) Incident pulse with markers at 2 ft (t_1) and 5 ft (t_2) from the transmitting antenna.

the transmitting antenna. The entire region between the direct transmission and the table-edge response forms a convenient time "window" to view the target response and allows one to "gate out" (in time) unwanted reflections. Thus undistorted transient target responses can be viewed without resorting to anechoic chambers. In addition, a single time-domain measurement obviates the requirement for measurement of the amplitude and phase response at many frequencies.

The accuracy of the measurement system can be estimated by measuring the standard deviation of the measured voltage. For the case where the peak of the incident pulse is approximately 500 mV and the 10-mV scale on the sampling oscilloscope is used, the standard deviation of the measured response has been measured as 0.06 mV or approximately 0.6 percent of the peak value of the measured return. In addition, the measured responses can be further processed on a general purpose computer by means of a convolution procedure to obtain the response due to a Gaussian-shaped incident pulse (or any other incident waveform whose spectrum is contained within that of the measured incident pulse) rather than the approximate smoothed impulse used in the actual measurements. Some measured results are displayed along with calculated responses later in this section.

The scattering range has an effective pulsewidth of approximately 200 lbf/in² at the 3-dB points. The measurement system can be used to determine the target response over a frequency band that extends from 100 MHz to approximately 4.5 GHz. This facility has been used to measure, directly, the smoothed impulse response of targets and to verify the results obtained by calculations. It has also been used in its own right

to evaluate the response of multi-scattering center targets, to identify the scattering centers, to measure their locations, and to measure their amplitude. The target response due to an arbitrary radar waveform can be obtained from these baseband measurements by a simple calculation provided that the necessary information is present in the measured result. That is possible in many cases by use of proper scaling and by application of existing knowledge concerning how scattering-center location and amplitude vary with the frequency of the illuminating waveform.

In summary the time-domain scattering range provides a valuable tool for diagnostic study, for verification of computational techniques, and for predicting the responses of real targets when illuminated by practical radars' waveforms.

B. Space-Time Integral-Equation Technique

1) *General*: Before the advent of high-speed digital computers, most electromagnetic scattering problems were solved in the frequency domain. The classical boundary-value-solution technique was applied to the problems where the target surface coincided with the coordinate surfaces of a separable coordinate system and was, therefore, limited to only a few target classes. Approximate techniques such as geometric optics, physical optics, and the Rayleigh approximation were applied to other problems to obtain estimates of the scattered response. However, with the development of high-speed digital computers, frequency-domain integral-equation techniques were developed [36], [37] which allowed solution of the scattering problem for simple bodies of arbitrary shape from near zero frequency well into the resonance region. These integral-equation solutions have provided many interesting and useful results; however, certain difficulties exist near the resonances of the internal-scattering problem [38]. Next, it was observed that an integro-differential equation and its solution could also be obtained directly in the time domain [7], [39]–[41]. The alternate approach of solving Maxwell's equations directly in the time domain by use of finite-difference techniques has also been developed and applied to a number of problems [42]–[46]. The integro-differential-equation approach requires only the solution on the surface of the scatterer, and thus provides advantages in both computer storage and running time over the finite-difference approach for scattering problems involving perfect conductors. In this paper, discussion will be limited to the integro-differential equation approach which will be called the Space-Time Integral-Equation Technique. For convenience, the integro-differential equations will be referred to as integral equations even though they do contain partial derivatives. This direct time-domain technique had advantages where the solution of transient electromagnetic-scattering problems was desired, in addition to eliminating the internal resonant problem observed in the frequency-domain solutions.

The geometry of the problem involved in deriving the integral equation for the current density on the surface of a perfectly conducting scatterer is shown in Fig. 10. There is an electromagnetic wave incident on a target which sets up currents on the surface of the target in such a way that the boundary condition on the surface is satisfied. These currents in turn produce a scattered field. Once the expression for the currents has been obtained, the problem is solved, since the field at any point in space can be computed from them.

The technique for obtaining an expression for the surface currents [29] is to start with the expression for the total field

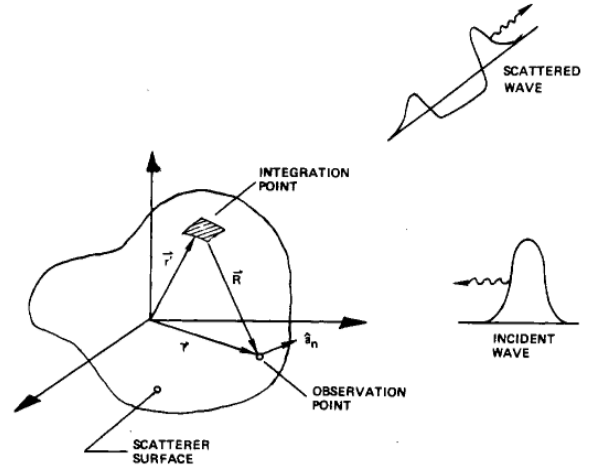


Fig. 10. Geometry of problem.

at an arbitrary point in space due to the incident field and the surface currents. This arbitrary point in space is then moved to a point on the surface of the scatterer using a limiting procedure. Next, the boundary conditions on the tangential components of the fields at the surface are applied, resulting in a space-time integral equation which can be solved numerically. Once the numerical representation of the surface current has been found, the field at any point in space, in particular the far field, can be calculated.

For the case of perfectly conducting scatterers being discussed here, either the H -field boundary condition or the E -field boundary condition is sufficient to derive the integral equation. If the H -field boundary condition is used, the H -field or Maue (1948) integral-equation [47] results. This is a second-kind equation, and no space derivatives on the surface are required. From a numerical solution standpoint this equation is well suited for closed conducting bodies, but gives problems on open thin surfaces or wires where the thickness goes to zero. If the E -field boundary condition is used, the E -field integral equation results. This is a first-kind equation, and space derivatives on the surface are required. From a numerical standpoint, it is well suited for targets with zero body thickness, and can handle thin wires and open thin surfaces.

2) *Closed Conducting Surfaces*: The solution of scattering from closed conducting surfaces is accomplished by numerical solution of the H -field integral equation, given as [29]

$$\vec{J}(\vec{r}, t) = 2\hat{a}_n \times \vec{H}^i(\vec{r}, t) + \frac{1}{2\pi} \int_S \hat{a}_n \times \left\{ \left[\frac{1}{R^2} + \frac{1}{Rc} \frac{\partial}{\partial \tau} \right] \vec{J}(\vec{r}', \tau) \times \hat{a}_R \right\}_{\tau=t-R/c} dS' \quad (19)$$

where

- $\vec{r}$ position vector to the observation point
- $\vec{r}'$ position vector to the integration point
- $R = |\vec{r} - \vec{r}'|$
- $\hat{a}_R = \frac{\vec{r} - \vec{r}'}{R}$
- t time in seconds
- c the speed of light

$\hat{a}_n$ unit normal at $\vec{r}$
 S surface of the target
 $\vec{H}^i$ incident magnetic field
 $\vec{J}$ surface current density.

The first term in the right-hand side of (19) may be considered the source term and represents the direct influence of the incident field on the current at the observation point $(\vec{r}, t)$. Moreover, this term, when applied to the illuminated side of the scatterer, yields the familiar physical-optics approximation for the surface current. The integral term on the right-hand side of (19) represents the influence of currents at other surface points on the current at $(\vec{r}, t)$. The crucial observation here is that the influence of other currents on the current at $(\vec{r}, t)$ is delayed by R/c , which makes the "marching on in time" numerical solution of (19) feasible.

Once the current density has been obtained, the field at any point in space can be computed. In most applications of interest, however, the quantity of interest is the far-scattered field, given by

$$\vec{H}^s(\vec{r}, t) = \frac{1}{4\pi rc} \int_S \left\{ \frac{\partial \vec{J}(\vec{r}', \tau)}{\partial \tau} \right\}_{\tau=t-R/c} \times \hat{a}_r dS' \quad (20)$$

where

$\vec{H}^s$ far scattered field
 r distance of observer from target
 $\hat{a}_r$ unit vector pointing from target to observer.

The technique for solution of the integral is to first write (19) in terms of its tangential components

$$J_u(\vec{r}, t) = J_u^i(\vec{r}, t) + I_u(\vec{r}, t)$$

$$J_v(\vec{r}, t) = J_v^i(\vec{r}, t) + I_v(\vec{r}, t)$$

where

$J_u^i = (2\hat{a}_n \times \vec{H}^i)_u$
 $J_v^i = (2\hat{a}_n \times \vec{H}^i)_v$
 I_u u component of the second term on the right-hand side of (19)
 I_v v component of the second term on the right-hand side of (19).

Next, the target surface is divided into patches, and the current is computed at the center, $\vec{r}_i$, of each patch

$$J_u(\vec{r}_i, t) = J_u^i(\vec{r}_i, t) + \rho_{ui} J_u(\vec{r}_i, t) + \sum_{k \neq i} f_{uik} \Delta S_k \quad (21)$$

where

$\rho_{ui} = \frac{1}{2} \sqrt{\frac{\Delta S_i}{4\pi}} (k_{ui} - k_{vi})$ is the contribution due to patch at r_i ,
 k_{ui}, k_{vi} the principle curvatures at $\vec{r}_i$,
 ΔS_i the area of the patch at $\vec{r}_i$,
 f_{uik} the contribution to $J_u(\vec{r}_i, t)$ due to currents on other patches at earlier points in time.

Rearrangement of terms in (21) yields the recurrence relation in time for the surface current

$$J_u(r_i, t) = \frac{J_u^i(\vec{r}_i, t) + \sum_{k \neq i} f_{uik} \Delta S_k}{1 - \rho_{ui}} \quad (22)$$

The ideal excitation would be an electromagnetic impulse.

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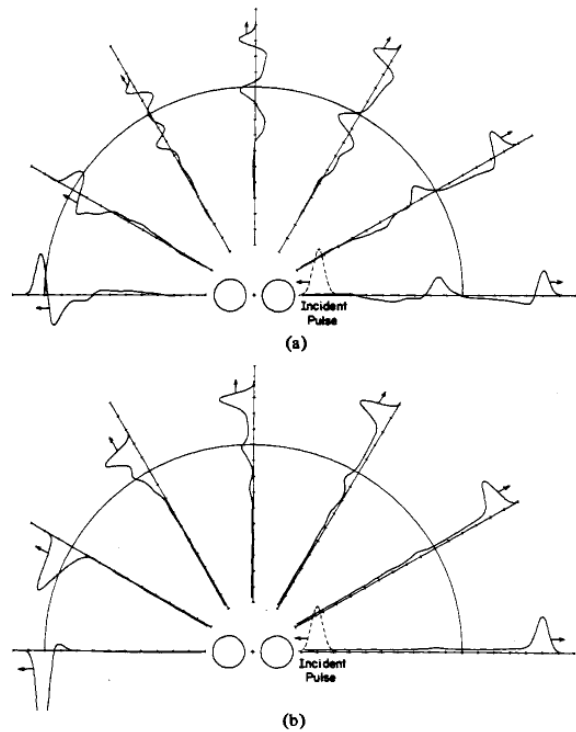


Fig. 11. Smoothed impulse response of twin cylinders with radius a (end-on incidence). (a) TE case. (b) TM case.

However, it is not amenable to numerical solution. What has been used is a smoothed impulse or a regularization of an impulse. The Gaussian form

$$e(t) = \frac{a_n}{\sqrt{\pi}} \exp(-a_n^2 t^2) \quad (23)$$

has been chosen because it has rapidly decaying frequency- and time-domain tails and is well suited for numerical solution. The response computed using this excitation is the smoothed-impulse response of the target.

For solution, the target geometry must be represented numerically. This is carried out by dividing the surface into patches which give a good geometry representation, are approximately curvilinear square, and have approximately equal areas. The minimum spacing between patch centers should be less than $\frac{1}{8}$ the width of the incident smoothed impulse.

The time variable must also be divided into increments for numerical solution. In the work to date, time has been divided into equal increments. The basic time increment should be less than $\frac{1}{8}$ the width of the incident smoothed impulse. Moreover, the basic time increment should be chosen to be less than the minimum spacing between patch centers for stability reasons.

Equation (22) is solved for the current density on a digital computer by simply marching on in time. Once the current density has been obtained, the far-scattered field is computed by performing the integration in (20) numerically. Procedures have been developed, implemented, and tested for computing the smoothed-impulse response of multiple cylindrical scatterers, smooth convex bodies, and bodies with edges.

As an example, the case of twin cylinders with radius a and center-to-center spacing of $2a$ is considered [48] in Fig. 11.

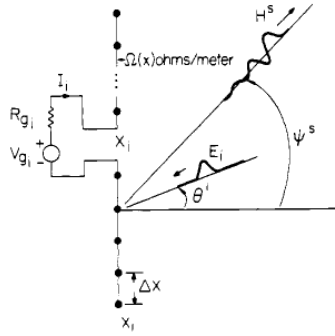


Fig. 12. Geometry of thin wire antenna as a scatterer, receiver, or radiator.

The TE-smoothed impulse response of twin cylinders with end-on incidence is shown in Fig. 11(a). In this space snapshot and those that follow, all the space dimensions are to scale except the distance to the scatterer. The basic space unit a (the radius of the cylinder in this case), and both the space-variation of the scattered field and the size of the scattering body, are drawn to the same scale. The large semi-circle is the distance reference of the scattered field and represents the locus of points in space that the peak of the incident pulse would have reached if it had been reflected from the origin. The specular return in the backscatter direction exhibits itself as an initial positive pulse. The initial part of the return is identical with that obtained from a single circular cylinder. This is expected since the incident wave does not reach the second cylinder until later in time. Although the second cylinder is "hidden" from the view of the observer in the backscatter direction, the second positive pulse that this observer sees can be attributed to a "reflection" of the incident wave from the front side of the back cylinder. The third positive pulse can be attributed to a wave traveling around the rear of the second cylinder.

The TM-smoothed impulse response of twin cylinders with end-on incidence is shown in Fig. 11(b). The initial part of the return in the backscatter direction is exactly the same as that obtained for a single cylinder. As expected, there is very little return in the backscatter direction that can be attributed to the shadowed cylinder.

3) *Thin Wires and Open Thin Surfaces*: As mentioned earlier, the E -field integral equation is more appropriate from a numerical standpoint for solving problems involving wires and open thin surfaces. The general space-time integral equation for a conducting body is given by [49]

$$0 = \epsilon \frac{\partial \vec{E}_{\tan}^i(\vec{r}, t)}{\partial t} + \left\{ \left[\nabla(\nabla \cdot) - \frac{1}{c^2} \frac{\partial^2}{\partial t^2} \right] \cdot \left[\frac{1}{4\pi} \int_S \left\{ \frac{\vec{J}(\vec{r}', \tau)}{R} \right\}_{\tau=t-R/c} dS' \right] \right\}_{\tan} \quad (24)$$

where

- ϵ permittivity of space
- ∇ del operator
- $\tan$ subscript denoting the component of the vector which is tangent to the surface at $\vec{r}$.

This is an integral equation of the first kind since the unknown, $\vec{J}$, appears only inside the integral on the right-hand side.

For the case of thin wires, (24) can be reduced by using the approximation that the current has no azimuthal component

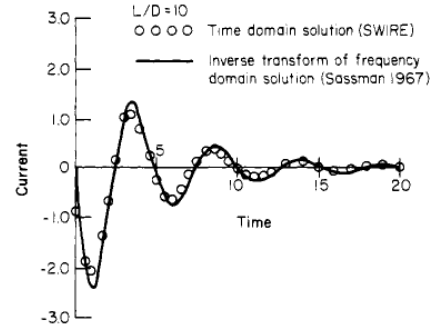


Fig. 13. Comparison of step response of wire scatterer computed by time-domain and frequency-domain techniques.

or variation. Using this technique, time domain solutions have been obtained for straight wires [40], [41], [49], [50], for wires with loading [51], for bent wires and wire structures [52]–[56], and for wires with nonlinear loads [57]–[59].

For the case where a thin straight wire is located on the x -axis as shown in Fig. 12, (24) reduces to

$$\left[\frac{\partial^2}{\partial x^2} - \frac{1}{\alpha} \sqrt{\frac{\epsilon}{\mu}} \left(\Omega(x) + \frac{R_g(x)}{\Delta x} \right) \frac{\partial}{\partial t} - \frac{\partial^2}{\partial t^2} \right] I(x, t) = -\frac{1}{\alpha} \left[\sqrt{\frac{\epsilon}{\mu}} \left(\frac{\partial E_x^i(x, t)}{\partial t} + \frac{1}{\Delta x} \frac{\partial V_g(x, t)}{\partial t} \right) \right] + \frac{1}{4\pi} \square^2 \int_{\text{Non Self-Segments}} \left\{ \frac{I(x', \tau)}{R} \right\}_{\tau=t-R} dx' \quad (25)$$

where

- I wire current
- V_g generator voltage
- R_g generator resistance
- Ω distributed resistance along the wire
- μ permeability of space
- Δx length of the segment that the wire is divided into
- $\alpha = \frac{1}{2\pi} \ln [(\Delta x + \sqrt{\Delta x^2 + a^2})/a]$
- a the wire radius
- $\square^2 = \left(\frac{\partial^2}{\partial x^2} - \frac{\partial^2}{\partial t^2} \right)$ the wave operator
- t time now redefined in units of light meters

Note that the left-hand side of this equation contains the one-dimensional wave operator with a loss term, and the right-hand side contains known source terms or currents at earlier points in time. Equation (24) is solved numerically by dividing the wire into segments and computing the currents at the end of each segment on a digital computer by marching on in time. This solution [51] is carried out in a manner similar to that used by Courant *et al.* [60] in their numerical solution of the one-dimensional-wave equation.

Fig. 13 displays the smoothed step response of a wire scatterer with length-to-diameter ratio of 10 and no loading. The comparison between the time-domain solution [49], [51], [61] and the alternate frequency-domain solution [62] shows good agreement.

4) *Flat Plates*: For the case of flat plates, (24) can be simplified by noting that the surface current flows in only one plane. The geometry of a rectangular flat plate is displayed

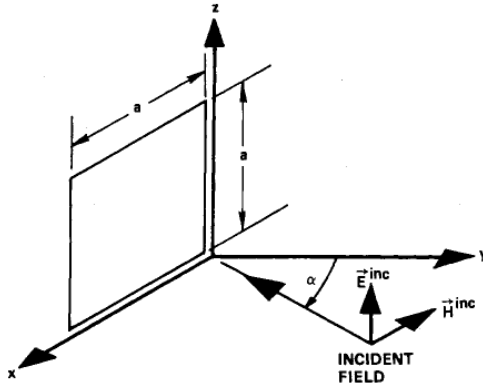


Fig. 14. Geometry of scattering from a rectangular flat plate.

in Fig. 14, where the plate is located in the $y = 0$ plane. For this geometry, the components of the surface current are obtained from (24) as

$$\begin{aligned} \square^2 J_X + \frac{\partial^2}{\partial X \partial Z} J_Z &= \frac{1}{\gamma} \left[-\epsilon \frac{\partial E_X^i}{\partial t} - (\square^2 + \boxtimes) \frac{1}{4\pi} \right. \\ &\quad \left. \cdot \int_{\text{Non-Self}} \left\{ \frac{J_X}{R} \right\}_{\tau=t-R} dS' \right] \\ \square^2 J_Z + \frac{\partial^2}{\partial X \partial Z} J_X &= \frac{1}{\gamma} \left[-\epsilon \frac{\partial E_Z^i}{\partial t} - (\square^2 + \boxtimes) \frac{1}{4\pi} \right. \\ &\quad \left. \cdot \int_{\text{Non-Self}} \left\{ \frac{J_Z}{R} \right\}_{\tau=t-R} dS' \right] \quad (26) \end{aligned}$$

where

- $\boxtimes$ the cross derivative terms in the $\nabla(\nabla \cdot)$ operator
- $\gamma = 2/(\sqrt{\Delta s/\pi} - (\delta/2)^2 - \delta/2)$
- Δs patch area
- δ plate thickness.

The left-hand side of (26) contains a wave operator operating on the major component and a mixed-space derivative operator operating on the other current component. It is this mixed-derivative operator that provides the coupling between the two components of surface current in (26). On the right-hand side are terms due to the incident field and due to currents at other points on the surface, but at earlier points in time. For numerical solution, the plate is divided into curvilinear square patches, and the current is computed at the center of all interior patches by marching on in time using the numerical representation of (26). The current on the edge and corner patches is obtained by applying the boundary conditions. For the cases studied to date, the component of the current perpendicular to the edge is set equal to zero on the edge patches. The component of the current parallel to the edge is obtained by noting that it must vary as one over the square root of the distance from the edge. Finally, both components of the current at the corner patches are set equal to zero.

As an example of the results which have been obtained using this technique, the smoothed-impulse response for a square flat plate is shown in Fig. 15. The calculated results for normal incidence are compared with results measured on the time-domain scattering range and shown to be in good agreement in

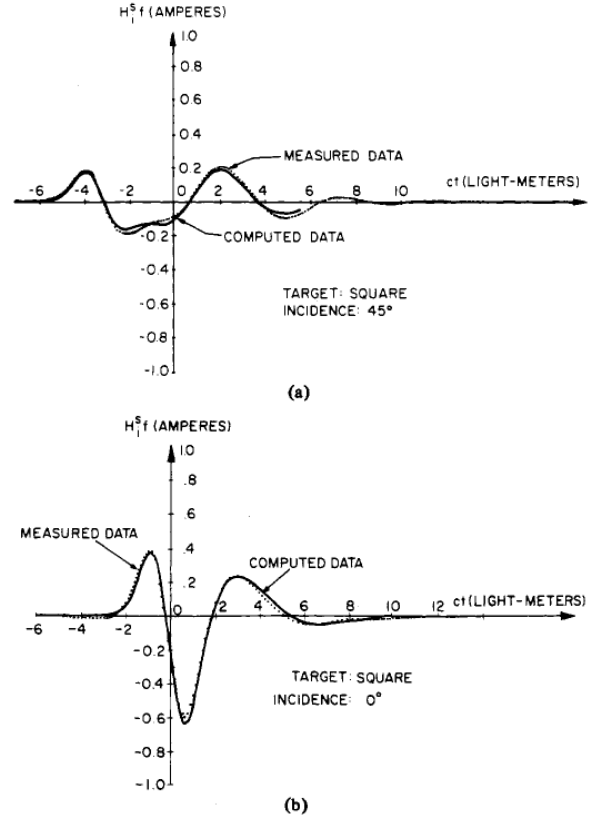


Fig. 15. Smoothed-impulse response of a square flat plate.

Fig. 15(a). Note that the initial response approximates a smoothed doublet as expected. This is followed by a second positive pulse which is due to a wave traveling across the plate face and returning to the observer. The calculated and measured results for a 45° angle of incidence also agree well, as can be seen in Fig. 15(b). This technique has also been extended to the case of nonplanar surfaces [33], [63], [64].

5) *Bodies with Wire Antennas and Struts*: The problem of determining the scattering properties of targets with wire antennas or struts is important in aircraft and satellite surveillance. Since this problem contains both closed surfaces and thin wires, a hybrid integral-equation approach is applied which uses the E -field integral equation over the wire portions and the H -field integral equation over the surface portions. The integral equations are [49], [65], for the wire current

$$\begin{aligned} \square^2 I(l, t) &= -\frac{1}{\alpha} \left[\sqrt{\frac{\epsilon}{\mu}} \frac{\partial E_R^i(l, t)}{\partial t} + \sqrt{\frac{\epsilon}{\mu}} \frac{\partial E_{SI}(l, t)}{\partial t} + \frac{1}{4\pi} \square^2 \right. \\ &\quad \left. \cdot \int_{\text{Non-Self-Wire}} \left\{ \frac{I(l', \tau)}{R} \right\}_{\tau=t-R} dl' \right] \quad (27) \end{aligned}$$

where

- l dimension along wire length
- E_R^i l th component of the incident E field
- E_{SI} l th component of the E field due to the surface currents
- t time in light meters

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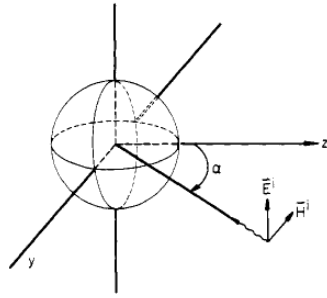


Fig. 16. SSS satellite model (length-to-diameter ratio of each antenna is 50).

and for the surface current

$$\vec{J}(\vec{r}, t) = 2\hat{a}_n \times \vec{H}^i(\vec{r}, t) + 2\hat{a}_n \times \vec{H}_w(\vec{r}, t) + \frac{1}{2\pi} \int_S \hat{a}_n \times \left\{ \left[\frac{1}{R^2} + \frac{1}{R} \frac{\partial}{\partial \tau} \right] \vec{J}(\vec{r}', \tau) \times \hat{a}_R \right\}_{\tau=t-R} dS' \quad (28)$$

where

$\vec{H}_w$ $\vec{H}$ field due to the wire currents
 t time in light meters.

The numerical solution of the coupled space-time integral equations, (27) and (28), is carried out by first dividing the surface into patches and the wires into segments. Next, the current at the center of each patch on the surface and at the end of each segment on the wire is computed by solving the numerical representation of the coupled equations (27) and (28) in a manner similar to that used for wires alone and surfaces alone. The current at both the wire free-end and at the wire surface-end are obtained by applying the boundary conditions for those two points.

As an example of an application of this technique, a model of the small scientific satellite (SSS), shown in Fig. 16, is considered. This target consists of a sphere located at the origin with four antennas attached, each a sphere-diameter long, coincident with the x and y axis. Both the E -plane- and H -plane-smoothed impulse response of this target is displayed in Fig. 17 for the case of an incident wave traveling in the negative z direction. It is interesting to relate the response in the backscatter direction to the target geometry. The specular return from the nose of the sphere appears first and is of the same value as was obtained for the case of scattering by a sphere alone. Next, there is a second peak that can be attributed mainly to the specular return from the wire antennas. This is then followed by a damped oscillation with a spacing between zero crossings that is approximately 2.5 sphere diameters. This is less than the distance between the opposite free tips of the wire (3 sphere diameters), but more than the length of an image representation of each antenna (2 sphere diameters). Moreover, the decay of these oscillations is much more rapid than would occur for the wires alone in free space with the same length-to-diameter ratio. This demonstrates a significant interaction between the antennas and the sphere surface.

6) *Bodies with Fins Attached:* The problem of determining the scattering properties of targets with fins attached is important in missile detection and classification. This problem has been addressed directly in the time domain [66] by using a hybrid integral-equation approach. This formulation used an E -field integral equation over the fin portions and an H -

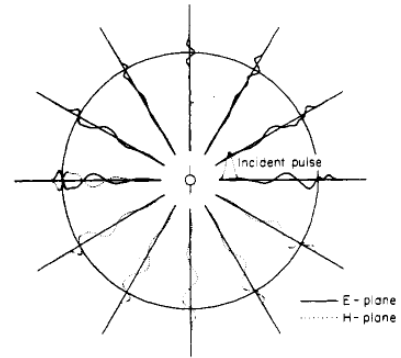


Fig. 17. Smoothed-impulse response of SSS satellite model.

field integral equation over the surface portions. The integral equations are, for the surface current

$$\vec{J}_S(\vec{r}, t) = 2\hat{a}_n \times \vec{H}^i(\vec{r}, t) + \frac{1}{2\pi} \int_S \hat{a}_n \times \left\{ \left[\frac{1}{R^2} + \frac{1}{R} \frac{\partial}{\partial \tau} \right] \vec{J}_S(\vec{r}', \tau) \times \hat{a}_R \right\}_{\tau=t-R} dS' + \frac{1}{2\pi} \int_F \hat{a}_n \times \left\{ \left[\frac{1}{R^2} + \frac{1}{R} \frac{\partial}{\partial \tau} \right] \vec{J}_F(\vec{r}', \tau) \times \hat{a}_R \right\}_{\tau=t-R} dS' \quad (29)$$

where

$\vec{J}_S$ surface current
 $\vec{J}_F$ fin current.

And for the fin current

$$\square^2 \vec{J}_F(\vec{r}, t) + \boxtimes \vec{J}_F(\vec{r}, t) = \frac{1}{\gamma} \left\{ -e \frac{\partial \vec{E}^i}{\partial t} - \square^2 [\vec{A}_{FNS}(\vec{r}, t) + \vec{A}_S(\vec{r}, t)] - \boxtimes [\vec{A}_{FNS}(\vec{r}, t) + \vec{A}_S(\vec{r}, t)] \right\}_{\tan} \quad (30)$$

where

$$\vec{A}_S = \frac{1}{4\pi} \int_S \left\{ \frac{\vec{J}_S(\vec{r}', \tau)}{R} \right\}_{\tau=t-R} dS'$$

$$\vec{A}_{FNS} = \frac{1}{4\pi} \int_{\text{Non-Self-Fin}} \left\{ \frac{\vec{J}_F(\vec{r}', \tau)}{R} \right\}_{\tau=t-R} dS'$$

$$\gamma = \frac{2}{\left[\sqrt{\frac{\Delta S}{\pi} - \left(\frac{\delta}{2} \right)^2} - \frac{\delta}{2} \right]}$$

ΔS fin patch area
 δ fin thickness.

The numerical solution of the coupled space-time integral equation given in (29) and (30) is carried out by marching on in time using a digital computer after first dividing the surface and fin into patches. The surface current is computed at the center of each surface patch using a numerical representation of (29) in a manner similar to that used for the solution by

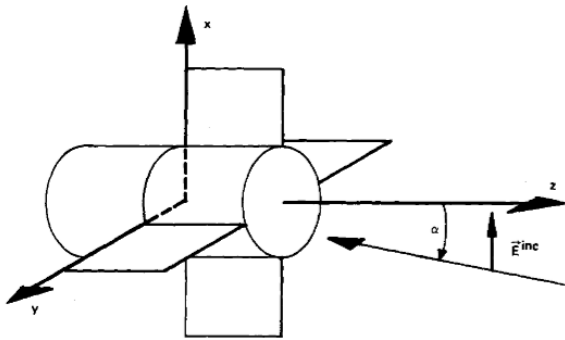


Fig. 18. Geometry of cylinder with fin attached.

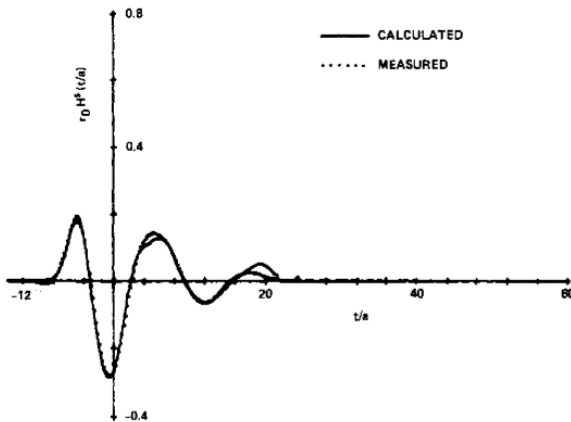


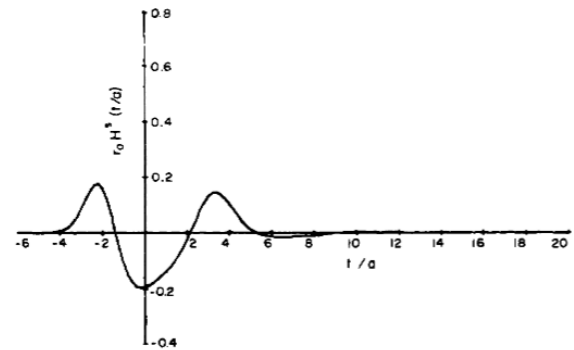
Fig. 19. Smoothed-impulse response of cylinder with fin attached for axial incidence.

surfaces alone. The fin current is computed at the center of each interior fin patch using a numerical representation of (30) in a manner similar to that used for a flat plate. The currents on the free-space edge patches and on the surface join patches of the fin are established by application of the appropriate boundary conditions.

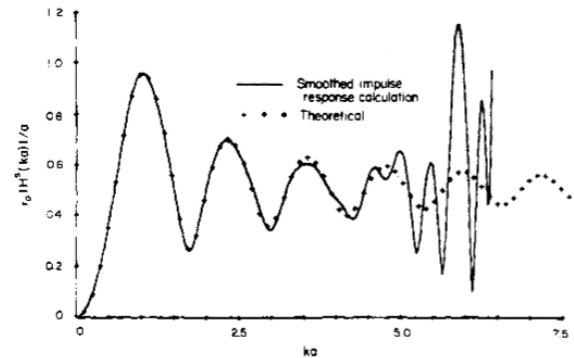
As an example of an application of this technique, consider the finned cylinder target shown in Fig. 18. This target consists of a right cylinder of radius a with length-to-diameter ratio of two-to-one with its axis coinciding with the z -axis and centered at the origin. Four fins (one cylinder diameter on a side) are arranged symmetrically about the cylinder body coinciding with the $x=0$ and $y=0$ planes. A vertically polarized smoothed impulse whose base width is equal to the cylinder length is used to illuminate this target. The smoothed impulse response that has been calculated for this target is compared in Fig. 19 for an axial angle of incidence with the measured result. The agreement between the calculation and the measurement is good. The initial portion of the response is due to the return from the face of the cylinder and the leading vertical fin edges. This is followed by the return from the far edge of the vertical fins, a creeping wave return, and oscillation due to the interaction between the fins and the cylinder body.

C. Impulse Response Augmentation Technique

What has been described up to now in this section is the space-time integral equation, its numerical solution, and



(a)



(b)

Fig. 20. Response of sphere with radius a . (a) Smoothed-impulse response. (b) Frequency response.

results obtained using a smoothed-impulse excitation. This numerical solution of the integral equation yields the smoothed impulse response of target sizes up to several pulse widths. If transformed into frequency domain terms, this technique yields results from zero frequency up well into the resonance region of the target. The impulse-response augmentation technique, first suggested in 1968 [7] and later developed in 1973 [32], [67], couples the calculated smoothed-impulse response with knowledge about the singular portion of the impulse response, such as location, type, and size of singularity, and knowledge about the asymptotic variation in frequency. This information is used to obtain an estimate of the impulse response (or, equivalently, an estimate of the frequency response) over the entire spectrum. The philosophy motivating this approach is to make use of all information available about the scattering problem in order to obtain an estimate of the "total" response. This may be viewed as a hybrid approach to the EM-scattering problem which combines results obtained by numerical techniques with results obtained by asymptotic techniques. Using this technique, results have been obtained for smooth convex bodies (sphere, sphere-capped cylinders, and prolate spheroid), for bodies with edges (right circular cylinders), and for open thin surfaces (square flat plates) [32], [33], [66]–[69]. To date, the technique has dealt only with the far field, although it could also be applied to the near field or the surface currents.

As an example of the results obtained with this technique, the case of a sphere is considered. Fig. 20(a) displays the far-field smoothed-impulse response of a sphere with radius a in the backscatter direction. The width of the incident pulse is

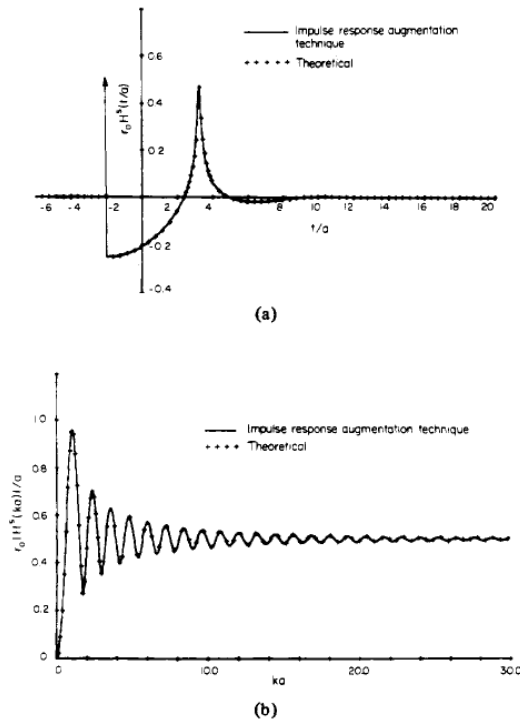


Fig. 21. Response of sphere with radius a . (a) Impulse response. (b) Frequency response.

twice the sphere diameter. Fig. 20(b) displays a comparison between the frequency response obtained by transforming the calculated smoothed-impulse response with the familiar Mie-series solution. The two results agree well up to a ka of 4.5 where $k = 2\pi/\lambda$ is the wavenumber. Fig. 21 compares the impulse response and frequency response obtained by use of the impulse-response-augmentation technique with the theoretical results and shows very good agreement. It is interesting to relate the features of the impulse response in Fig. 21(a) to the geometry of the target. The initial return is an impulse of amplitude 0.5 that is due to the specular reflection from the nose of the sphere. This is followed by a negative step of amplitude 0.25 that is due to the return from the vicinity of the specular point and would be predicted by a physical-optics approximation. It is interesting to note that the slope of the waveform following the negative step is a function of the difference in the principal curvatures at the specular point [66]–[68]. Finally, the second positive pulse at $t/a \approx 3.1$ is due to the creeping wave that travels around the rear of the sphere and returns to the observer.

D. Inverse Scattering

The previous discussion of time-domain signature analysis has been directed towards the direct-scattering problem, i.e., given the incident waveform and the target geometry, find the scattered field. This section addresses the inverse-scattering problem in which it is desired to find the target geometry when only the incident waveform and the scattered waveform are given. The solution of this problem has direct application to the problem of classification and identification of radar targets. This discussion addresses the time-domain solution of this problem.

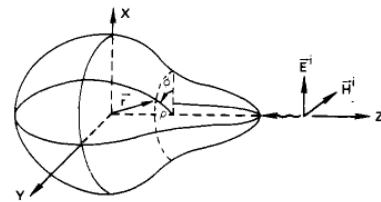


Fig. 22. Geometry of rotationally symmetric scattering problem.

The specific case considered is that of a rotationally symmetric scattering problem which consists of a rotationally symmetric target illuminated by a wave from the axial direction as shown in Fig. 22. This target geometry can be represented simply by its contour function ρ as defined in Fig. 22. The space-time integral equation for the solution of this inverse problem is given by [66]

$$\rho(z, t) = \left[2r_0 H_R^s(\vec{r}, t) - \frac{1}{2\pi} \int_S \{ \vec{J}_{CR}(\vec{r}', \tau) \times \hat{a}_r \}_{\tau=t-R} dS' \right]^{1/2} \quad (31)$$

where

- ρ contour function of the target
- r_0 distance of observer from the target
- H_R^s far scattered field ramp response of target
- H_R^i incident ramp H field
- $\vec{J}_R$ total current
- $\vec{J}_{CR} = \vec{J}_R - 2\hat{a}_n \times \vec{H}_R^i$

This expression represents the exact relationship between the scatterer response H_R^s and the target geometry, ρ . In particular, the target contour function ρ is given in terms of the scattered far field H_R^s , and in terms of "correction currents," $\vec{J}_{CR}$ that appear at earlier computed surface points at earlier points in time. This space-time integral equation has been solved by iteration of the marching on in time approach [32], [69]–[72].

An example of the application of this technique is shown in Fig. 23 for the case of a sphere-cap flat-end cylinder when viewed from the sphere-cap end. Note that the first iteration gives a good estimate of the spherical end cap and sides of the cylinder; but it took 13 iterations to give a reasonable estimate of the far end of the target. Although there is some rounding of the far edges, this technique can provide a powerful tool in obtaining information concerning the geometry of a target in the shadow region and in this case, shows good agreement with the actual contour in Fig. 23(e).

V. BASEBAND RADAR

The previous sections of this paper have discussed applications of time-domain electromagnetics involving both the amplitude and time variation of the smoothed-impulse response. In the early seventies it became apparent that subnanosecond-pulse technology could be employed in a variety of proximal sensing applications [8]. Since the presence or absence of a return is sufficient data for such a BASEband Radar (BAR) system, the impulse-response sensing system can be designed to determine threshold (location) information only, leading to significant reductions in cost and complexity over the electronics required for target signature analysis. This can be accomplished by em-

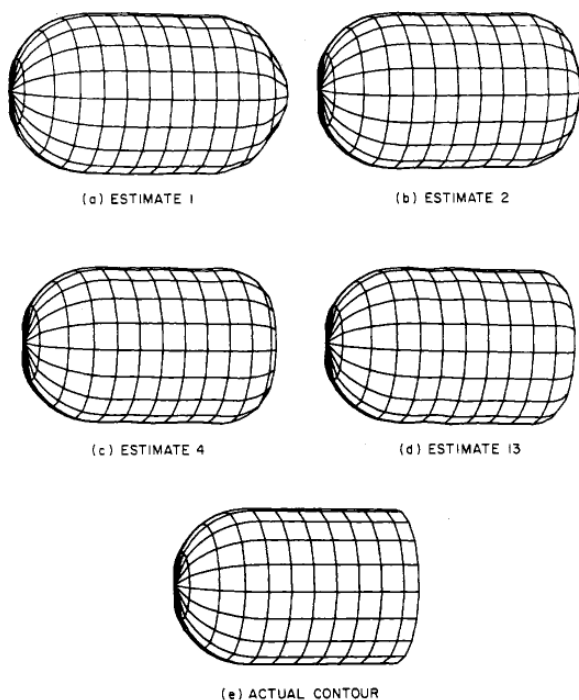


Fig. 23. Contours of sphere-capped flat-end cylinder obtained by time-domain solution of inverse scattering problem.

playing a receiver which, over a narrow time window, senses when any part of the target return exceeds a given threshold level and measuring the time delay between this return and the "main bang". Two different types of receivers, discussed below, were designed to meet this requirement [73].

A. Baseband Receivers

1) *The Single-Hit Receiver:* The threshold device is the key element in the single-hit receiver. There are only two devices that have been found capable of triggering on very-fast low-energy pulses—the avalanche transistor and the tunnel diode. Of the two, the tunnel diode is preferred because its V - I characteristics are well defined and it is almost an order of magnitude more energy-sensitive than the avalanche transistor. The disadvantages of the tunnel diode are as follows.

- 1) Its V - I characteristics are temperature sensitive.
- 2) It is a two-terminal device whose input and output must be taken across the same terminals.
- 3) It is bistable, and once the state is changed it must be reset before another operation can be performed.

By using analog and digital-feedback techniques we found it possible to obviate the common disadvantages of the tunnel diode and take full advantage of its higher sensitivity and lower cost. A diagram of a typical single-hit superregenerative receiver front-end is shown in Fig. 24.

To explain the operation of the single-hit receiver, consider that the constant-current emitter-follower source i feeding the tunnel diode is linearly increasing from zero as a function of time. When it reaches level 3 (see Fig. 24) the tunnel diode switches to its higher state 3a. A step transition in voltage, ΔV , is transmitted through a buffer amplitude to a pulse stretcher and a first-avalanche-transistor detector. The pulse

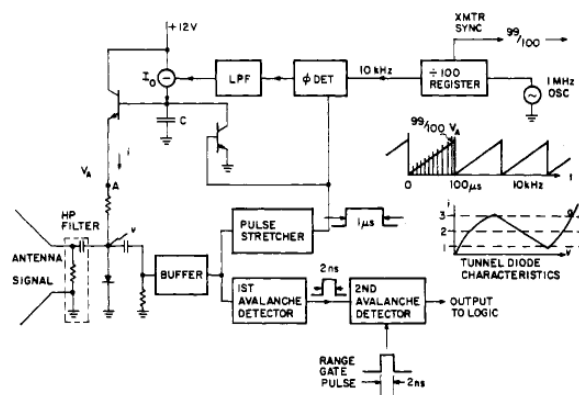


Fig. 24. Block diagram of single-hit receiver.

stretcher (monostable) multivibrator transforms the fast subnanosecond leading edge to a $1\text{-}\mu\text{s}$ pulse.

The output of the pulse stretcher performs two functions. One signal discharges the capacitor that initially produced the linearly increasing current that caused the tunnel diode to change state in the first place. After discharge, the current through the tunnel diode begins to rise again, resulting in a sawtooth-like waveform at node A as shown in Fig. 24. The period of the sawtooth waveform is set, for example, at 10 kHz by choosing C and I_0 appropriately.

A portion of the output from the pulse stretcher is also fed to an input port of a phase comparator. There the trailing edge of the stretched pulse is compared to a 10-kHz square-wave reference signal derived from the output of a register which divides down a sinusoidal 1-MHz clock source by a factor of 100.

The output from the phase detector is filtered and fed back to control the constant-current source I_0 ; controlling I_0 determines the slope of the sawtooth waveform shown in Fig. 24 and ultimately stabilizes the tunnel diode against temperature changes in its V - I characteristics. For example, if the transition current of the tunnel diode tends to increase with increasing temperature, then the period of the sawtooth will tend to increase. An error voltage will then develop at the output of the phase detector that will increase the output of the constant-current source, tending to increase the slope of the sawtooth and restore the period. The integrity of the period is maintained by the loop gain, which is about 100. Thus, variations due to temperature are reduced by a factor of 100.

Since the period of the sawtooth is held reasonably constant by analog feedback, we may use the various stages of the $\div 100$ register to dissect the $100\text{-}\mu\text{s}$ period accurately. Indeed, the register neatly divides the period into 100 equal duration states. If the transmitter is triggered to fire on the 99th state, the signal that returns from a target (e.g., only nanoseconds later) will find the tunnel diode in its most sensitive region; the tunnel diode free-runs $1\text{ }\mu\text{s}$ later. Because of the time constant of the present-temperature stabilization loop, the transmitter cannot be activated more than every other pulse, or at a maximum 5 kHz rate.

Finally, the output of the first avalanche detector is also fed to a high-speed AND network which is in the form of another avalanche transistor. There the reconstituted target signal is compared to the output of the range-gate generator which is synchronized to the main bang. If both the range gate and

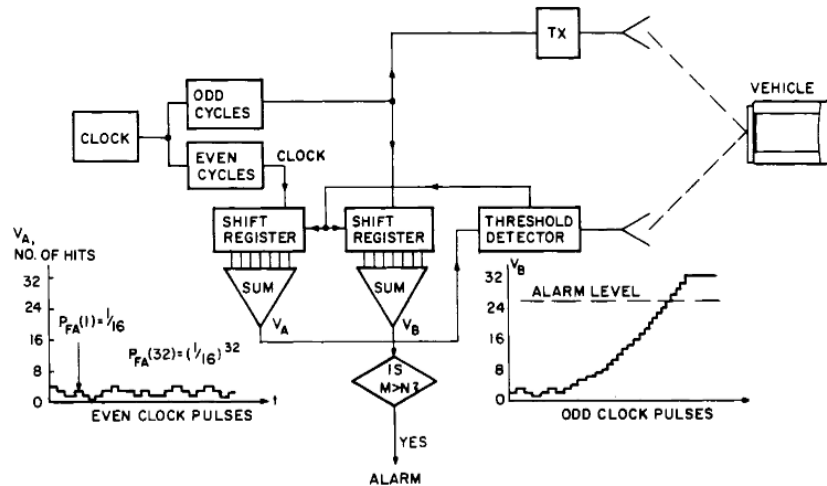


Fig. 25. Constant false-alarm rate (CFAR) receiver.

target signal are time coincident, then an output pulse is fed to the logic.

2) *The Constant False-Alarm-Rate Receiver [74]*: By employing processing techniques, the sensitivity of the tunnel-diode receiver can be increased significantly at the expense of circuit complexity. The scheme used to accomplish the increase in sensitivity employs a constant-false-alarm-rate (CFAR) processor. As the threshold level of the single-hit receiver is reduced, the sensitivity is obviously increased, but at the expense of the false-alarm rate. If the hits obtained from noise alone are fed to a register that assigns a +1 each time a noise spike exceeds the threshold and a zero when it does not, then the outputs from, say, 32 stages of this register can conveniently be summed. By appropriately feeding the register, a sliding average of detector hits in any group of 32 consecutive periods can be obtained. The threshold can be arbitrarily lowered until, on the average, three or four hits due to noise alone are contained in the register: the transmitter is disabled during this time. A dc voltage derived from the sum port can be used via a feedback loop to automatically maintain, on the average, an arbitrary number of pulses in the register due to noise. This loop sets the constant bias current through the tunnel-diode threshold detector in the absence of signal (i.e., on alternate pulse-repetition periods when the transmitter is disabled). When the transmitter is enabled, the threshold for establishing the presence of a target is set in a separate register-and-sum network. For example, a target is said to be present when 25 or more of the 32 stages of the second register are filled. A block diagram of the CFAR receiver is shown in Fig. 25. These two receiver designs have been used in the various applications described in the next section.

B. Application to Low-Cost Radar (Free-Space Time-Domain Reflectometry)

The perfection of a waveform source, efficient radiating and receiving antenna elements, and inexpensive threshold-sensing devices leads naturally to the development of a short-range free-space time-domain reflectometry or BAR system [8]. There are many useful applications for devices of this type, especially if the cost can be kept well below \$1000 for a transceiver pair. The range, naturally, is limited by such a system because very little energy is transmitted and detection

is primarily a function of energy on target. A unique feature of a BAR system is its high-resolution capability and virtually interference-free operation because of range gating and the processing action of the CFAR receiver.

The primary application areas at this time appear to be in:

- 1) the automobile-electronic areas, where a BAR sensor has been successfully demonstrated for both precollision sensing and braking;
- 2) airport ground-traffic control, where intersection control at high-density terminals and independent altitude warning is desired, especially under Category 3 landing conditions;
- 3) ship docking for both spaceship and large oil tanker use.

1) *Automobile Electronic Sensing*: At the Fourth European Microwave Symposium we reported our progress in automobile precollision sensing (e.g., the BARBI system) where the range was 5 ft or less [75]. During the interim period, work on a BAR braking sensor was conducted. In an audited test by the U.S. Government in Boulder, CO, in May 1975 and later in June 1977, the use of the BAR sensor in a binaural configuration was demonstrated for sensing objects at a range of about 150 ft. Beam narrowing is accomplished by signal processing using the transmitter-receiver configuration shown in Fig. 26 in conjunction with a microwave tapped delay line [76], [77]. By preselecting the ports at which the signals from both receive channels coalesce, it is not difficult to show that the effective beamwidth θ can be narrowed in accordance with the relationship

$$\theta = 2 \sin^{-1} \frac{\Delta}{2a} \quad (32)$$

where Δ is the difference between the tap locations at which the pulses first coalesce, and a is the separation between the transmitter and receiving units. Using this technique it is possible to achieve an effective beamwidth of several degrees. It was demonstrated at these tests that the resolution of the system is fine enough to distinguish between a human being, a bicyclist, a motorcyclist, and an automobile only feet apart.

2) *Airport Ground Control Sensors*: There are several situations in an airport where a need exists for a short-range high-

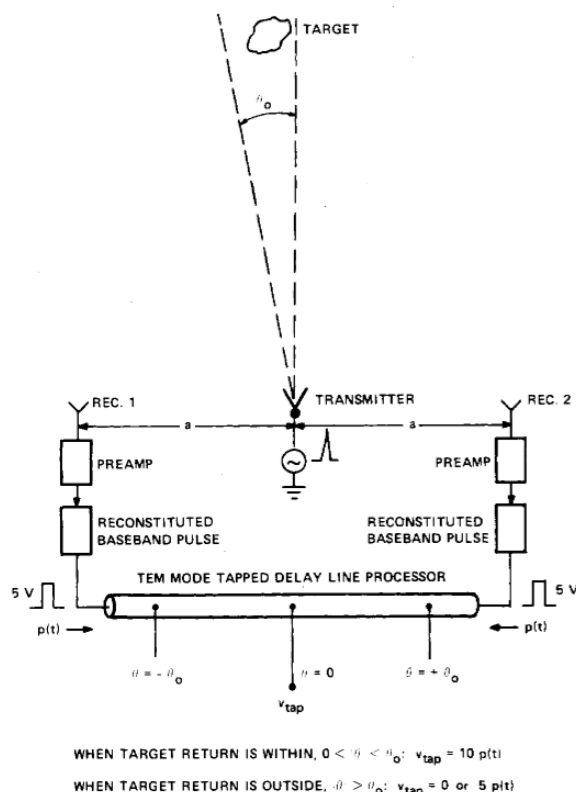


Fig. 26. Beam shaping by time-domain coincidence techniques.

resolution sensor; among these are:

- 1) runway clear—determining whether the complete runway is clear of traffic; if not, determining where the obstacle is;
- 2) taxiway intersection control—determine presence, direction of travel and speed of aircraft approaching an intersection;
- 3) taxiway guidance—for very-low-visibility conditions, give an indication of aircraft position relative to a taxiway centerline;
- 4) ground-based altimeter—at the approach end of a runway, provide an indication of aircraft height from the ground when it is at altitudes up to 1000 ft.

Tests have been made on all of these applications, and one such test is illustrated in Fig. 27. The transmitter and receiver units were set up adjacent to an active taxiway with two receivers flanking the transmitter. As the aircraft taxied past the equipment, the difference in times of acquisition by the two receivers was used to gate the clock to a counter, whose output in turn fed a small calculator programmed to display velocity directly. Tests on a variety of aircraft from small private planes to large jets gave a range of velocity readings from 10 to 30 mi/h, which correlated well with speed measurements made using a stop watch. A similar system has been used to measure automobile velocity from the side of a road (instead of looking along it as for Doppler methods), giving results at 20 mi/h which agreed within ± 1 mi/h with a pair of mechanical switch pads on the roadway. For

runway/clear applications, tests have shown that complete coverage requires a several-hundred-foot spacing between successive units alongside the runway.

3) *Ship-Docking Sensor*: The capability to display directly in feet the range to a large-broadside target has been used to demonstrate use of a baseband radar for tanker docking at distances out to 1600 ft. It is significant that a return is obtained only when some metallic part of the target is normal to the line of the radar, and hence although the system has broad angular coverage, it only detects the tanker when it is normal to the beam-pointing angle: the ultimate range of the present receiver against the side of a large tanker is about 5000 ft. With a 1-s integration interval, range uncertainty during tests was about 1-ft rms out to 300 ft and 3-ft rms to 1000 ft. For a slowly docking ship, range rates can be measured to about 1 ft/min by digital smoothing and differentiation.

C. Guided-Wave Reflectometry

The previous subsections have dealt with the free-space radiation of baseband-like pulses in a radar mode. There are also applications of baseband techniques in guided-wave modes where there is little external radiation, and the system performs strictly as a reflectometer. Examples of these applications include liquid-level sensing and a new collision-avoidance system for rapid-transit vehicles.

1) *Sensing Liquid Level*: There are many applications where the accurate sensing of liquid level is of importance. Some of these applications include the sensing of oil or liquid natural gas levels in tanker ships and the measurement of fuel supply in aircraft. Present techniques for accomplishing liquid-level measurement use primarily float-type devices and suffer from a lack of accuracy and reliability. Especially in those areas where highly viscous or caustic fluids are involved, floats seem to fail after brief use. Ultrasonic as well as EM sensors have also been employed. Two EM types which have been employed with limited success (i.e., depending on the corrupting effects of the fluid) have been the straight (air dielectric) coaxially supported cable [78] and the segmented coaxial capacitor [79]. In both devices, the presence of the fluid changes the capacity per unit length of the line and the accompanying resonant frequency and/or the reflection coefficient at the air-fluid interface, and becomes a measure of the liquid level. Eventually, the space between the inner and outer conductor of the line or coaxial capacitor becomes contaminated.

The time-domain reflectometer method described here has been experimentally verified and is similar to the coaxial-cable approach inasmuch as the air-liquid interface results in a mismatch and a measurable change in the reflection coefficient. The major difference is that there is no need for an outer conductor: with the air of a single-wire Goubau-line system one is able to measure the liquid level in a tank, and, with the aid of a processor, can also identify the liquid contained in the tank [80]. The scheme makes use of a baseband pulse source (e.g., 200 ps in duration) feeding a single Teflon-coated wire which extends to the bottom of perhaps a 100-ft tank; for a 100-ft tank the diameter of the wire should be greater than $\frac{1}{8}$ in. Processing equipment can be minimal if liquid level is the only concern. Somewhat more sophistication is required for the quantitative analysis of the liquid in the tank.

The scheme can best be explained with the aid of Fig. 28. A rectangular pulse $p(t)$ is fed past a parallel line coupler and to a coaxial horn and a dielectric coated line. The leakage signal fed to the processor via the coupler establishes the $t = 0$ reference signal. The signal reflecting off the edge of the horn

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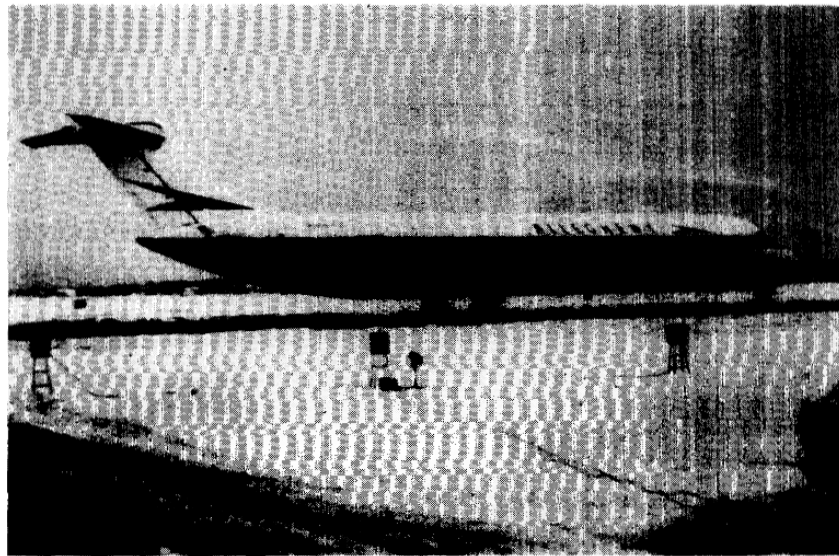


Fig. 27. Airport intersection-control set.

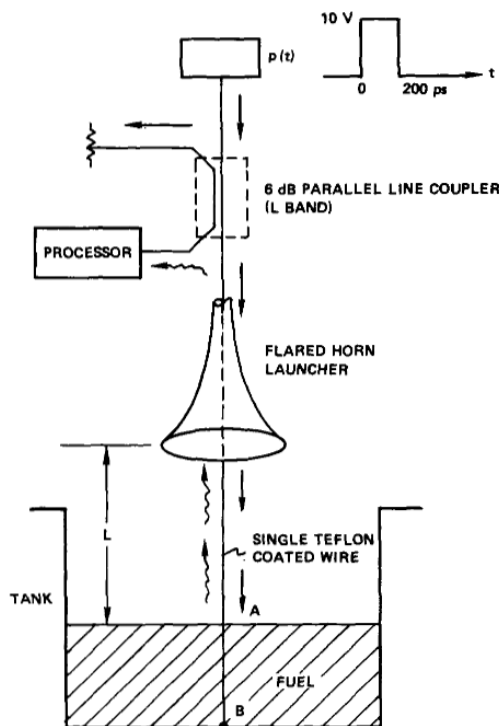


Fig. 28.

produces an unwanted signal in the processor a known distance after $p(t)$ and is gated out. The signals of importance are:

- 1) the amplitude and time delay of the signal reflecting off the surface at A which uniquely determines the height of the liquid in the tank and the properties of the liquid;
- 2) the signal reflecting off the bottom of the tank at B for calibration purposes especially if the tank also contains a second liquid (e.g., water).

This technique results in a liquid-level accuracy in the order of an inch or better for the first 10 ft into the liquid. Early experiments conducted indoors used oil and water (e.g., the oil had a dielectric constant of 5) in a large laboratory sink. Other experiments conducted in an outdoor water tower further verified the level-determining theory. Additional work is currently proceeding to optimize the design of the horn, coupler, and processor. In cases where only the liquid level is of concern, the processor design is considerably simplified. It was determined that the energy in the incident pulse is at least three orders of magnitude below the danger level for the explosion of aviation fuel.

2) *Collision Avoidance System Using Baseband Reflectometry (CASBAR)*: A guided-wave approach for avoiding collisions between rapid-transit vehicles is particularly attractive when the vehicles are committed to a track or guideway. This approach permits one to see around curves as well as eliminating anomalous responses from overhead structures and tunnels. In addition, there is virtually no radiation in a guided wave system.

One such proposed scheme, known as CASBAR, employing baseband techniques is shown in Fig. 29. A surface-wave guiding structure similar to that used in the liquid-level sensor lies parallel to the track. Each vehicle carries a self-contained radar system with logic circuits connected to the train control. The operational sequence starts with a short pulse launched on the main guide via a surface wave directional coupler. The wave travels along the guide until it is partially reflected by one or more passive reflectors adjacent to the guide mounted on the forward vehicle. This reflected signal is coupled into

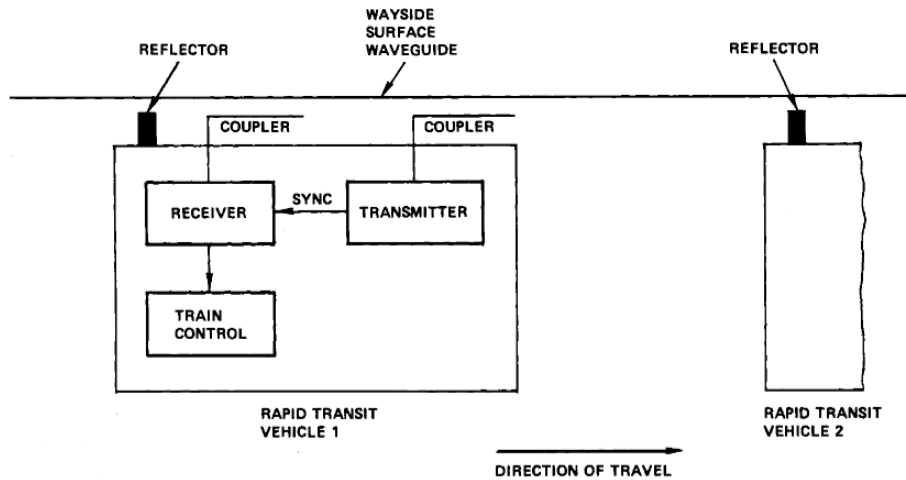


Fig. 29. CASBAR scheme.

the tunnel diode receiver by another directional coupler. (For redundancy, another system may be installed on the other side of the guideway.) A target may be detected at a range of several-hundred feet or more; range rate is also computed by the receiver processor. A breadboard model of this system has recently been completed and tested in cooperation with the Boeing Aerospace Company and the Urban Mass Transit Administration personnel. The details will be reported to the technical Community in the near future when the analysis of the results is completed.

D. Other Applications

In a more recent application, directed at propagating auto-traffic information (e.g., density and velocity of the stream) in a large urban environment, baseband techniques are being applied to a free space communications link. Here the low-cost, high-reliability, low EM pollution, apparent lack of licensing requirements, and insensitivity to incidental jamming make it an attractive candidate as a communications link at least over short ranges. The accomplishment of synchronization which establishes a range gate and hence interference-free operation is established via a free space link. This is realized by capturing the 19-kHz subpilot carrier of a local FM stereo multiplex station both at the transmitter and receiver and synchronizing the main bank and range gate, respectively, to the zero crossing of the carrier signal. Many large urban areas in the United States operate stereo multiplex stations 24 hours each day. Experiments have shown that over-modulation by the stations (i.e., in the left + right and the left - right channels) is well controlled by the majority of the stations transmitting classical music; the over-modulation produced by rock music stations perturbs the 19-kHz zero crossings by almost 1 μ s as compared to 200 ps for stations broadcasting classical music. By FCC standards, in the United States the frequency of the subpilot carrier is maintained plus or minus two cycles. This signal is derived at the output of the ratio detector or discriminator of an inexpensive FM receiver placed at both ends of the link and used to trigger the baseband transmitter and range-gate circuitry.

Work in this area is continuing to determine, experimentally, the effects of multipath fading.

Finally, the development of subnanosecond-pulse technology has impacted the field of computers. The use of these techniques has increased substantially the speed at which computations are made; logic speeds can now be measured in GHz, as recently reported in the literature [81], [82]. In addition, subnanosecond-pulse techniques have successfully been employed to multiplex data. In one application, N parallel lines of digital data are multiplexed into a single serial line which is sent to a remote terminal where it is reconstituted into the original N -channel message [9]. As semiconductor technology further improves, we can expect this area to assume increasing importance.

VI. SUMMARY

The advances in the state-of-the-art in subnanosecond-hard-ware and software technology have brought to the engineer much more powerful tools for the solution of time-domain electromagnetic problems. The development of subnanosecond-pulse techniques, which began with waveform generation, has now progressed beyond the fields of metrology and computer application to radar, target-signature analysis, and communications theory. The time-domain scattering range represents a valuable aid for verification of calculations, for determination of target responses, and for better understanding of the transient scattering phenomena itself. The space-time integral equation technique provides a powerful tool for calculation of the transient scattered response of targets and provides valuable insight into the solution of target-classification problems. The BAR approach to sensing shows that a fresh look at the radar problem, at least for short-range application, can result in a high-resolution and inexpensive sensor which uses virtually no microwave components. Previous radar approaches to the short-range problem have involved scaled-down versions of the conventional and more complex long-range systems; the BAR unit more closely resembles a free-space time-domain reflectometer. Because a BAR is so broadband, there are no highly resonant circuits present, and hence there are only loose mechanical mounting tolerances to meet. We expect the next major progress in this field to be in the communications area. Secure and low EM polluting systems having ranges in excess of thousands of feet, and data rates of several kilobits seem available.

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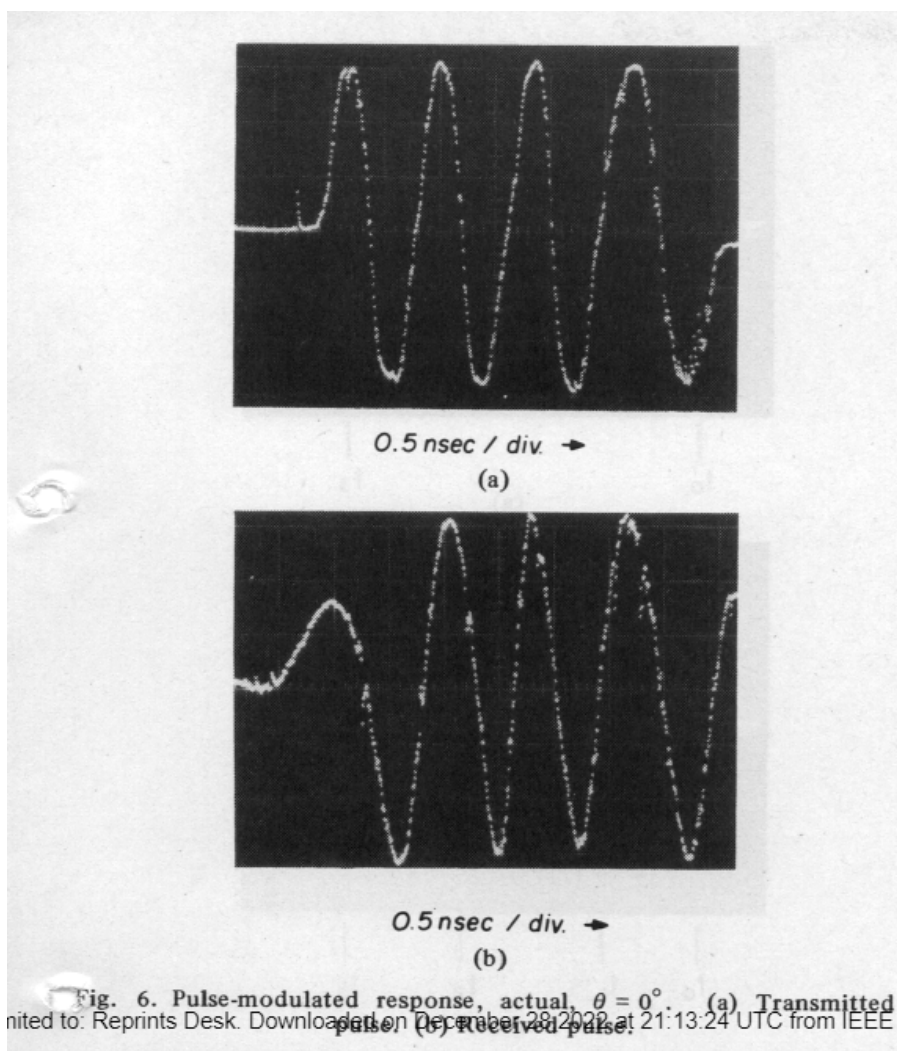
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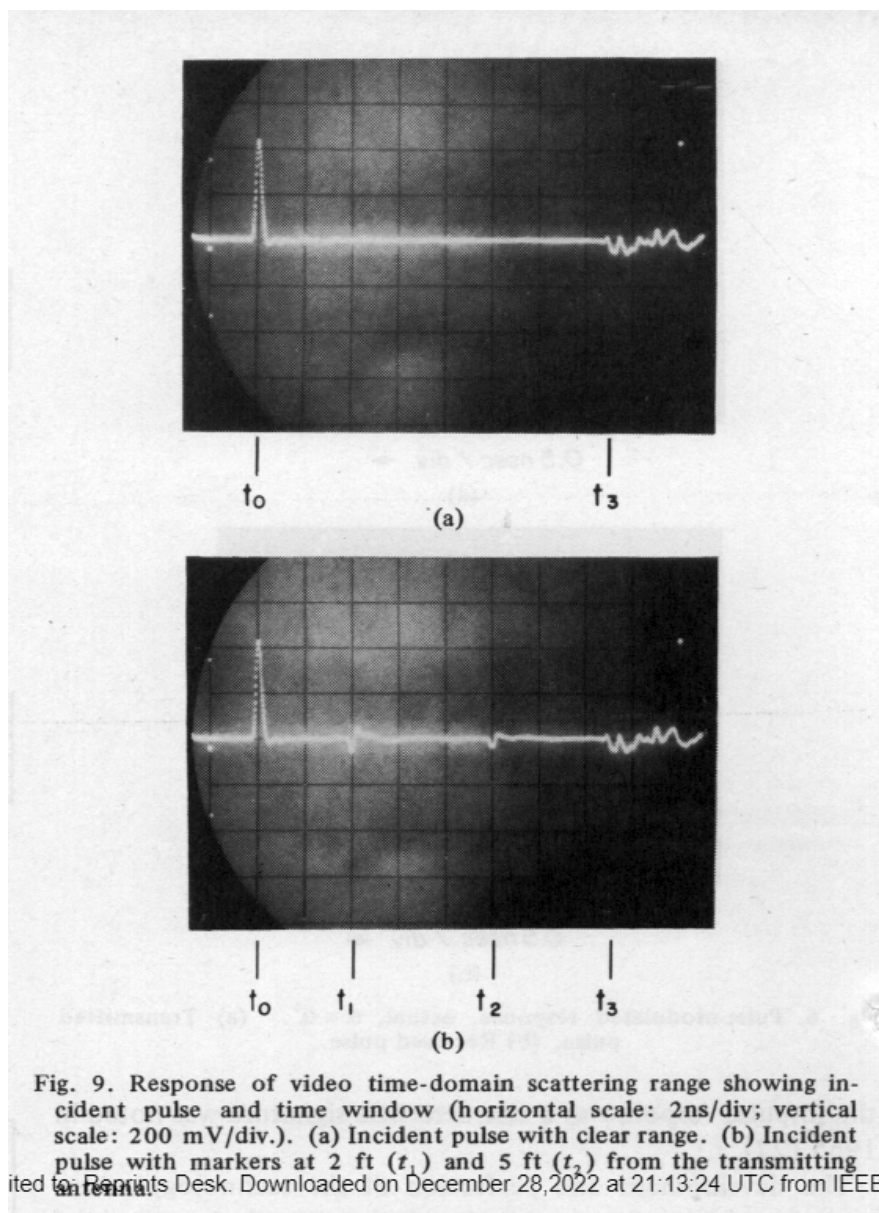
The authors wish to acknowledge the early support of this work by Dr. Bernard Oliver of the Hewlett-Packard Company who in 1962 loaned us a preproduction model of one of the first sampling oscilloscopes. We also wish to acknowledge the guidance and encouragement of both Professor Athanasios Papoulis of the Polytechnic Institute of New York and Professor Walter Weeks of Purdue University where these ideas began, and the many investigators in this country and abroad who also actively contributed to the work described in this survey paper and whom we have tried our best to reference. Finally, we wish to thank our editor, Alan Roberts, for his untiring assistance over the years in helping us order our thoughts.

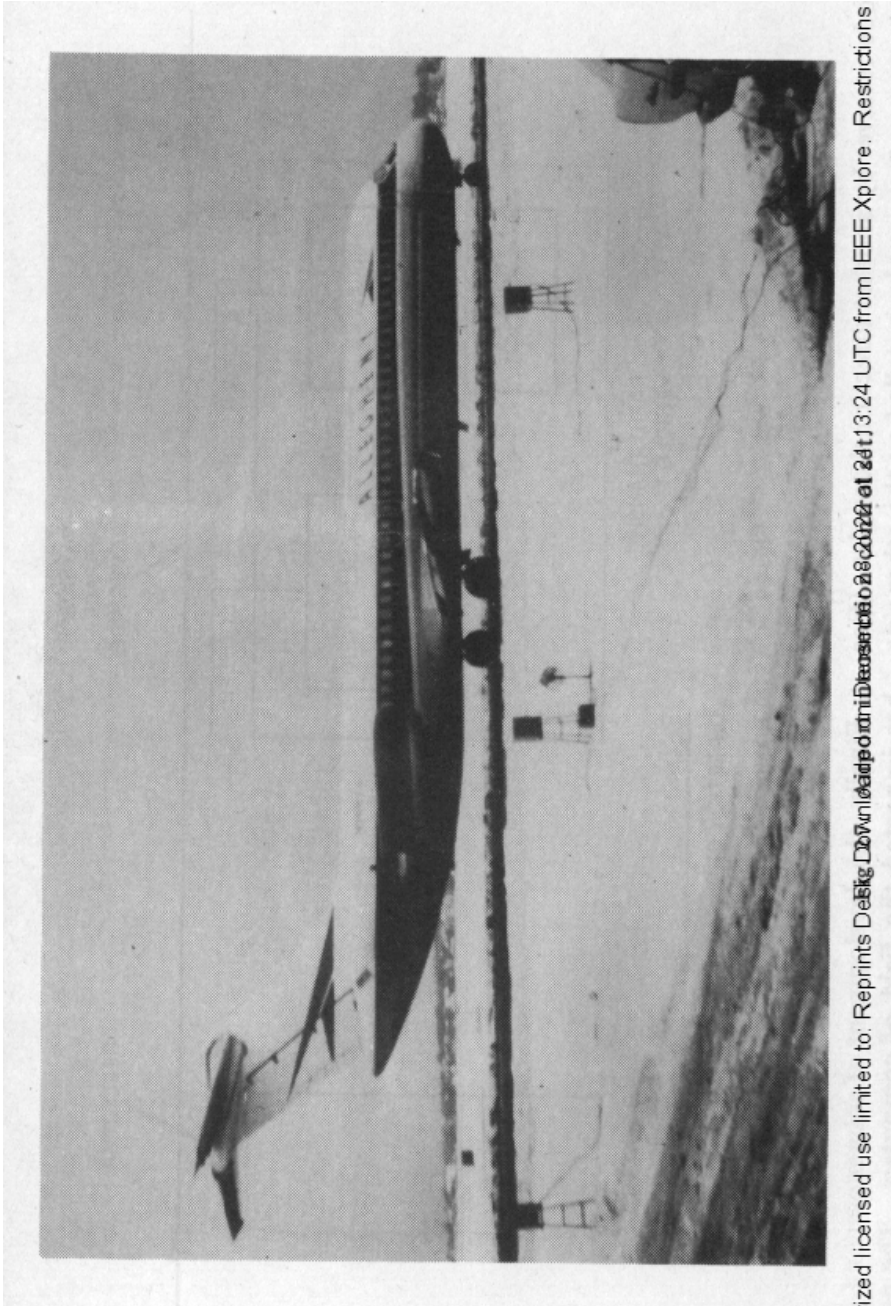
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