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MICROWAVE SHORT-PULSE BED-LEVEL DETECTOR

Annual Report for the Period January 1—December 31, 1981

By
Constantine A. Balanis
Donald M. Delauder

Work Performed Under Contract No. AC21-79ET15158

West Virginia University
Morgantown, West Virginia

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MICROWAVE SHORT-PULSE BED-LEVEL DETECTOR

ANNUAL REPORT
January 1 - December 31, 1981

CONSTANTINE A. BALANIS AND DONALD M. DELAUDER

DEPARTMENT OF ELECTRICAL ENGINEERING
WEST VIRGINIA UNIVERSITY
MORGANTOWN, W.VA. 26506

PREPARED FOR
U.S. DEPARTMENT OF ENERGY
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ABSTRACT

A short-pulse microwave system for measuring the bed-level within a fluidized-bed combustor, has been designed, built, and laboratory tested on static beds. The system is a short-pulse radar which operates in the frequency region of 6.75-10.95 GHz as a time-domain measurement system. Laboratory measurements of static bed-levels, for smooth and corrugated surfaces of metal plates and limestone sand, agree to an average of 2.0% of the actual heights. Additionally, the system was tested with a dielectric thermal protector, which did not compromise the accuracy of the measurements.

Analytical models have been formulated to provide insight into the operation of the system on a wide range of simulated targets without the necessity of performing expensive and difficult laboratory experiments. Two formulations have been used to describe electromagnetic scattering by a rough surface as a function of frequency: the space harmonic model and the physical optics model. A reconstruction technique has been devised which uses the scattering models and the spectrum of the transmitted pulse to synthesize the reflected pulse. The data generated by the models compare well to previously published data and to experimental results.

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I. INTRODUCTION

The process of fluidized-bed combustion is a technique for burning coal which is high in sulfur content, without releasing excessive pollutants into the atmosphere. The fluidized-bed combustion system also presents certain economic advantages over conventional burning and power generation schemes [1]. The fluidized-bed combustor is essentially an upright, enclosed chamber which allows for the introduction of granular solid fuel and air into the bottom of the chamber. The granular solids are typically high-sulfur coal and limestone (mixed). These solids rest initially on a perforated plate. During operation, air is blown through the perforations creating a turbulent mixture of air and granular solids which behaves as a fluid.

The objective of this part of the project is to investigate a time-domain microwave radar technique for measuring the bed-level within the combustor. This investigation includes the design, construction, testing, and evaluation in a simulated static environment in our laboratory of the candidate radar system. A frequency-domain system to meet the same requirements has been previously studied [2]. In addition, analytical models have been developed which predict the performance of the system, and they were utilized to examine the reflective characteristics of smooth and rough surfaces of metal and limestone. The computed data were

compared with measurements to validate the models and system concept as applied to this application, and to perform parametric studies to assess the capabilities and limitations of the system under various conditions.

The temperatures within the chamber during operation of the combustor can reach 1200⁰ F to 1500⁰ F. To remain operational and maintain its reliability, any instrumentation designed for this task must be protected from this hostile environment.

II. THEORY

The theory of operation of a pulse radar system is introduced in this section. An analytical model is developed which uses the characteristics of the transmitted pulse, the antenna gain, and the target surface roughness to determine the power level and the shape of the reflected pulse. The analytical model will prove useful in determining capabilities and limitations of the pulse radar system.

A. Pulse Radar Principles

The basic principle in a pulse radar system is that in a given medium electromagnetic waves travel at known speeds. This phenomenon can be used to determine how far a wave has traveled if the time to traverse this distance is known. The electromagnetic parameters of the media, through which the wave propagates, must also be known. Thus, the distance traveled is given by

$$d = vt = \frac{1}{\sqrt{\mu\epsilon}} t \quad (1)$$

where

d = distance (m)
 v = velocity (m/s)
 μ = permeability of medium (H/m)
 ϵ = permittivity of medium (F/m)
 t = time of travel(s)

In air, the wave velocity is approximately equal to 3×10^8 m/s. Therefore, (1) reduces to

$$d = (3 \times 10^8) t \quad (2)$$

These principles can be extended to the problem of range measurement (level detection) using a pulse radar. The pulse radar uses electromagnetic pulses as timing marks. That is, the time that the pulse leaves the radar is noted. The pulse reaches the target at some later time, and it is reflected toward the radar. The pulse travels the same distance back to the radar, where its time of arrival is also measured. A diagram illustrating this sequence appears in Figure 1. The difference of the two times is the total travel time, and it is the time taken for the pulses to travel twice the range to the target. Therefore, the distance equation is modified and can be written as

$$R = \frac{1}{2} v (t_a - t_d) = \frac{v \Delta t}{2} = 1.5 \times 10^8 \Delta t \quad (3)$$

where Δt is the time difference and R is the target range.

Practical considerations require that the radar pulse is generated repetitiously with a repetition period T . Also, the pulse exhibits a finite duration τ . The pulse duration should be much less than the repetition period to ensure good resolution between the transmitted and reflected signals. The pulse should also be shaped such that a reference point for timing is easily chosen. A waveform

Figure 1. Diagram of radar system pulse propagation.

which meets these requirements is shown in Figure 2a. Some convenient reference points for timing might be the leading or trailing edges of the pulse. If the pulse were sharply peaked, the peak could be selected as the timing reference point. Figure 2b illustrates a typical display of transmitted and reflected pulses with the trailing edge shown as the timing reference.

A final constraint on the pulse repetition period is determined by the maximum intended target range. If the pulses are transmitted with a repetition period which is too short, the reflected pulse will not return to the radar before another pulse is radiated toward the target. This dilemma is often referred to as range ambiguity, because it becomes difficult to determine which reflected pulse corresponds to a given transmitted pulse. An illustration of this condition is provided in Figure 3. For the example shown, it would not be possible on an oscilloscope to determine which time difference is correct, that is, Δt_1 or Δt_2 . The repetition period should clearly be chosen to be at least as long as the greatest expected round trip travel time that will be encountered during operation of the radar system. Range ambiguity is not limited to pulse radar systems as it can also exist in a variety of frequency-domain systems.

B. Pulse Power Reflection

In this section the theory of the electromagnetic wave scattering by a rough surface is presented. Two different analytical models are

a. transmitted pulse.

b. transmitted and reflected pulses.

Figure 2. Radar pulse trains.

Figure 3. Example of range ambiguity for pulse radar system.

utilized, and their results are compared. This theory is then used to develop a method of reconstructing the reflected pulses based on the spectrum of the transmitted pulse. The results provided by the theory facilitate the calculation of reflected power levels and pulse shapes.

1. Electromagnetic Scattering

Two formulations have been used to obtain the scattered power from a rough surface. One method, usually referred to as the space-harmonic model [3], [4], assumes a perfectly conducting sinusoidal surface, and it is exact. The other formulation, usually called physical optics model [5], [6], is valid for any geometrical shape and electrical parameters of the surface. However, this model is not exact in that a physical optics approximation is used; hence, its name. The analytical details of both will now be discussed.

Space Harmonic Scattering Model

The space harmonic scattering model represents the scattered electromagnetic field as a superposition of discrete space-harmonic plane waves. The appropriate geometrical configuration is displayed in Figure 4. The electromagnetic field is assumed to be horizontally polarized in the y-axis direction. This model can also be developed for vertical polarization, but the horizontal case is sufficient for our application. Horizontal polarization is also often called TE (for transverse electric) polarization, while vertical polarization is referred to as transverse magnetic, or TM, polarization.

Fig. 4. Sinusoidal surface geometry for space-harmonic scattering model.

The space harmonic solution involves writing the total field, above the $z = 0$ plane, as a sum of the incident and scattered waves. That is,

$$\psi(x,z) = \psi_0(x,z) + \psi_{sc}(x,z) \quad z \geq 0 \quad (4)$$

where

ψ_0 = incident field

ψ_{sc} = scattered field

ψ = total field

The incident plane wave can be written as

$$\psi_0(x,z) = e^{ik(\alpha_0 x - \beta_0 z)} \quad z \geq 0 \quad (5)$$

where

$$k = \frac{2\pi}{\lambda} = \text{wavenumber}$$

$$\alpha_0 = \sin\theta_i$$

$$\beta_0 = \cos\theta_i$$

The representation of the scattered field is a superposition of plane waves weighted by the scattering coefficients of each of the plane waves. This is written as

$$\psi_{sc}(x,z) = \sum_{n=-\infty}^{\infty} R_n e^{ik(\alpha_n x + \beta_n z)} \quad z \geq 0 \quad (6)$$

where

$$\alpha_n = \sin\theta_n, \beta_n = \cos\theta_n \quad (6a)$$

θ_n is the scattering angle of the n^{th} space harmonic. The angle θ_0 is the specular angle; that is, the angle at which all of the scattered radiation would be directed if the surface were flat. The scattering coefficient, R_n , is the unknown quantity to be solved for. To accomplish this, the scattering angles θ_n must first be found. Given a surface which is sinusoidal and described by

$$h(x) = -\frac{d}{2} \left[1 + \cos\left(\frac{2\pi}{\lambda}x\right) \right] \quad (7)$$

the scattering angles can be solved for by using the familiar grating equations of

$$\alpha_n = \alpha_0 + n\lambda \quad (8a)$$

and

$$\beta_n = \sqrt{1 - \alpha_n^2} \quad n = \pm 1, \pm 2, \dots \quad (8b)$$

Assuming a sinusoidal surface shape ensures that the R_n can be solved for exactly. Desanto [3] has shown that by applying the boundary condition that the field on the surface must vanish and by using Green's theorem, the scattering coefficients are

$$R_m = -\frac{\beta_0}{\beta_m} i e^{i\tau_m^+} \left[J_m(\tau_m^+) + \sum_{n \neq 0} a_n J_{m-n}(\tau_m^+) \right] \quad (9)$$

$$m = 0, \pm 1, \pm 2, \dots$$

where the a_n 's are solutions of the system

$$J_m(\tau_m^-) + \sum_{n \neq 0} a_n J_{m-n}(\tau_m^-) = 0, \quad m = \pm 1, \pm 2, \dots \quad (9a)$$

$J_m(\tau_m^\pm)$ is the m^{th} order Bessel function with

$$\tau_m^\pm = \frac{\pi d}{\lambda} (\beta_0 \pm \beta_m) \quad (9b)$$

The solution for the scattering coefficients is seen to involve an infinite order system of equations. Generally, the only scattering coefficients which are of interest are the ones with β_n real. This corresponds to radiating space harmonics. When β_n is imaginary, the scattering coefficients correspond to evanescent space harmonics. Evanescent fields are those which decay rapidly away from the surface, and they do not contribute to the real radiated power. To obtain a practical solution, the infinite set of equations is truncated to order $2M$. The truncation introduces an error into the solution for R_n .

To maintain the error at acceptable levels, an energy conservation criterion is calculated on the radiating scattered fields. That is, the sum of the power in the propagating scattered space harmonics must equal to the power in the incident wave. If the energy conservation check fails, the order of the system can be increased.

The normalized energy density in the n^{th} space harmonic at a plane above the surface is

$$|\Gamma_n|^2 = |R_n|^2 (\beta_n/\beta_0) \quad (10)$$

The energy conservation error due to the truncation is then

$$\epsilon_m = |1 - \sum |\Gamma_n(M)|^2| \quad (11)$$

where the summation is taken only over the propagating harmonics.

$\Gamma_n(M)$ is the approximate Γ_n found using the $2M$ order truncation.

Using this simple criterion, the value of M necessary for accurate computation can be determined. It has been observed that M must increase as the depth of the surface roughness increases. Typically, M varies from 8 to 18. It is possible that for a given surface the solutions will not converge for any value of M .

Using the coefficients R_n and assuming a constant plane wave illumination of a surface of width L (assumed large in terms of a wavelength), it has been shown [4] that the scattered power is given by

$$\sigma(\theta) = \sum |Q(kL(\sin\theta - \sin\theta_n)) R_n|^2 \quad (12)$$

where the summation is taken only over the propagating modes and

$$Q(kL\alpha) = \frac{\sin(kL\alpha)}{kL\alpha/2} \quad (12a)$$

Using (12), the scattering pattern for the surface can be calculated. The assumptions made in developing this analytical

model were:

- 1) perfectly conducting sinusoidal surface
- 2) surface is two-dimensional with corrugations in one dimension only
- 3) electric field parallel to corrugations (horizontal polarization)
- 4) surface length L is large in terms of a wavelength

There is, however, no restriction on the depth, d , and the period, λ , of the surface.

Physical Optics Model

To investigate the scattering characteristics of nonconducting surfaces, a physical optics model was implemented and it was used to complement the computations of the space harmonic scattering model. Figure 5 displays the geometrical configuration used in the development of this model.

The physical optics model used here is derived from a full-wave exact solution [6]. The full-wave solution consists of a complete expansion of the fields into vertically and horizontally polarized waves using exact boundary conditions and assuming an arbitrary scattering surface configuration. The complete expansions are substituted into Maxwell's equations. Upon applying the boundary conditions and Green's theorem, Maxwell's equations are converted into a set of generalized telegraphists equations which are solved iteratively. Although the resulting solutions are exact,

Figure 5. Scattering geometry for physical-optics model.

programming difficulties occur.

The solution can be simplified greatly by invoking the physical optics approximation. This approximation assumes that the minimum radius of curvature of the surface is not too small (fractions of a wavelength) and that there is no shadowing. Shadowing occurs when a portion of the surface does not receive direct radiation from the incident wave because of radiation blockage by another portion of the surface. The conditions of no shadowing are met for a sinusoidal surface when [5]

$$1 \ll \frac{2\ell^3}{d\lambda^2} \cos(\theta_i + \tan^{-1}(\frac{\pi d}{\ell})) \quad (13)$$

When θ_i is small and $\ell/d \geq 5$, this condition is met. When (13) is not satisfied, the approximation fails gradually instead of collapsing immediately. The equations of the physical optics model will now be presented.

The essential results of the full-wave solution are contained in a definite integral for the scattering coefficients, in which the integrand is given (for deterministic surfaces) by a closed form expression based on the incidence and reflection angles, electrical parameters, and the rough surface characteristics. The scattering coefficient matrix is written as

$$\underline{C} = \frac{1}{A_y} \iint_{A_y} \underline{I}^f(C_o^{in} \underline{E}) \underline{I}^i \cdot e^{ik_o(\hat{n}^f - \hat{n}^i) \cdot \underline{r}_s} d\underline{A} \cdot \hat{n} \quad (14)$$

where

A_y = projection of the rough surface onto the reference plane

$k_0 = \omega \sqrt{\mu\epsilon}$ is the wavenumber

$\hat{n}^f$ and $\hat{n}^i$ = tangent vectors in the directions of the scattered and incident fields, respectively

$\underline{r}_s$ = vector from the origin to points on the surface

$\hat{n}$ = unit vector normal to the surface

C_0^{in} is the matrix of elementary surface scattering coefficients (a matrix is used so all polarizations can be accounted for). This matrix essentially accounts for scattering by an elemental horizontal rough surface. The matrix operator I^i decomposes the waves, which are horizontally and vertically polarized with respect to the reference plane, into vertically and horizontally polarized waves with respect to the local plane of incidence. The operator I^f performs the reverse transformation. Thus, the operation $I^f(C_0^{in}E)I^i$ effectively converts the matrix $C_0^{in}E$ into a matrix of scattering coefficients in which the rough surface is regarded as a continuum of elementary surfaces of varying slope and height.

The integral formulation for the scattering coefficient matrix given above is valid for both the exact full-wave solution and for the physical-optics approximation. It is possible, however, to simplify the scattering coefficient equation when using the physical-optics approximation. Then

$$\underline{C} = C_o^i F_o \begin{bmatrix} C_R^f V C^i + S_R^f H S^i & C_R^f V S^i - S_R^f H C^i \\ S_R^f V C^i - C_R^f H S^i & S_R^f V S^i + C_R^f H C^i \end{bmatrix} I \quad (15)$$

where

$$C_o^i F_o = \frac{\cos^2 \theta_o^s}{\cos\left(\frac{\theta_o^f + \theta_o^i}{2}\right) \cos\left(\frac{\theta_o^f - \theta_o^i}{2}\right)} \quad (15a)$$

$$\cos \theta_o^s = [(1 + C_o^i C_o^f - S_o^i S_o^f \cos(\theta^f - \theta^i))/2]^{1/2} \quad (15b)$$

$$C_o^i = \cos \theta_o^i, C_o^f = \cos \theta_o^f, S_o^i = \sin \theta_o^i, S_o^f = \sin \theta_o^f \quad (15c)$$

$$R^V = \frac{\cos \theta_o^s - \cos \theta_1^s (\eta_1/\eta_o)}{\cos \theta_o^s - \cos \theta_1^s (\eta_1/\eta_o)} \quad (15d)$$

$$R^H = \frac{\cos \theta_o^s - \cos \theta_1^s (\eta_o/\eta_1)}{\cos \theta_o^s - \cos \theta_1^s (\eta_o/\eta_1)} \quad (15e)$$

$$\cos \theta_1^s = (1 - \sin^2 \theta_o^s / n_r^2)^{1/2} \quad (15f)$$

$$\eta_r = (\epsilon_1 \mu_1 / \epsilon_o \mu_o)^{1/2} \quad \eta_o = (\mu_o / \epsilon_o)^{1/2} \quad (15g)$$

$$\eta_1 = (\mu_1 / \epsilon_1)^{1/2}$$

$$C^i = \frac{C_o^i + C_o^f - C_o^i (1 + \cos 2\theta_o^s)}{s_o^i \sin 2\theta_o^s} \quad (15h)$$

$$S^i = \frac{S_o^f \sin(\theta^f - \theta^i)}{\sin 2\theta_o^s} \quad (15i)$$

$$C^f = \frac{C_o^f + C_o^i - C_o^f (1 + \cos 2\theta_o^s)}{S_o^f \sin 2\theta_o^s} \quad (15j)$$

$$S^f = \frac{S_o^i \sin(\theta^f - \theta^i)}{\sin 2\theta_o^s} \quad (15k)$$

$$I = \frac{1}{A_y} \iint_{A_y} e^{ik_o(\underline{n}^f - \underline{n}^i) \cdot \underline{r}_s} dx dz \quad (15l)$$

The integration for the approximate case is much simpler than for the full-wave solution. It is possible that the rough surface will scatter the wave in a polarization which is different from the incident wave polarization. This effect is known as depolarization and is accounted for by using a matrix representation for the scattering coefficient $\underline{C}$. For instance, if a vertically polarized wave is scattered horizontally, the scattering coefficient accounting for the depolarization is given by the upper right element of the matrix $\underline{C}$. If the first superscript of $\underline{C}$ is the incident polarization and the second represents the scattered polarization, then the possible polarizations are indicated by

$$\underline{C} = \begin{bmatrix} C^{VV} & C^{VH} \\ C^{HV} & C^{HH} \end{bmatrix} \quad (16)$$

The term of interest for this study is C^{HH} which represents the horizontal wave polarization. The factors R^V and R^H are the Fresnel reflection coefficients for vertical and horizontal polarization, respectively. These are the factors which introduce the electrical parameters of the surface into the analytical model. The surface characteristics are introduced in the model by $\underline{r}_s$. That is, if the surface is described by $h(x,z)$, then

$$\underline{r}_s = x\hat{a}_x + h(x,z)\hat{a}_y + z\hat{a}_z \quad (17)$$

where $\hat{a}_x$, $\hat{a}_y$, and $\hat{a}_z$ are unit vectors in the indicated directions.

Radar Range Equation

Two analytical models have been described which yield scattering coefficients for rough surfaces. In addition to the scattering coefficients, the radar range equation must be used to predict the power received by the radar system. The radar equation for the ratio of received power, P_r , to transmitted power, P_t , is given by [7]

$$\frac{P_r}{P_t} = \sigma \frac{G_0^2 \lambda^2}{(4\pi)^3 R^4} \quad (18)$$

where

- σ = power scattering coefficient
- R = target range
- λ = wavelength
- G_0 = antenna power gain

The power scattering coefficient, σ , is the same as that described in the space harmonic model, but it is the square of the scattering coefficient described in the physical optics model. Also, it should be noted that σ is a function of frequency.

The radar system that was designed for this investigation utilizes a pyramidal horn antenna. The gain, G_o , is a function of the antenna construction materials, the impedance match of the antenna to its feed line, and it is strongly dependent on the antenna dimensions, which are shown in Figure 6. The gain can be written as

$$G_o = e_t D_o = e_r e_{cd} D_o \quad (19)$$

where

e_t = total antenna efficiency

e_r = reflection efficiency

e_{cd} = conduction-dielectric efficiency

D_o = directivity

For most aperture antennas the conduction-dielectric losses are very small so that $e_{cd} \approx 1$. In addition, most aperture antennas, and especially horns, are also well matched to their transmission lines so that $e_r \approx 1$. Thus the gain is approximately equal to the directivity.

Figure 6. Antenna dimensions for calculating directivity.

In general, the horn antenna directivity is given by [8]

$$D_0 = \frac{8\pi\rho_1\rho_2}{a_1b_1} \left\{ [C(u)-C(v)]^2 + [S(u) - S(v)]^2 \right\} \left\{ C^2\left(\frac{b_1}{\sqrt{2\lambda\rho_1}}\right) + S^2\left(\frac{b_1}{\sqrt{2\lambda\rho_1}}\right) \right\} \quad (20)$$

where

$$u = \frac{1}{\sqrt{2}} \left(\frac{\sqrt{\lambda\rho_2}}{a_1} + \frac{a_1}{\sqrt{\lambda\rho_2}} \right) \quad (20a)$$

$$v = \frac{1}{\sqrt{2}} \left(\frac{\sqrt{\lambda\rho_2}}{a_1} - \frac{a_1}{\sqrt{\lambda\rho_2}} \right)$$

$C(x)$ and $S(x)$ are the cosine and sine Fresnel integrals, respectively.

2. Pulse Reconstruction

To determine the power level and shape of a pulse scattered by a given surface, a technique has been formulated which reconstructs the reflected pulse using the scattering coefficients found in the previous section and the transmitted pulse spectrum. The basis of this method entails calculating the scattered power at each frequency of the spectrum of the transmitted pulse. Then, from this calculated scattered power spectrum, the reflected pulse waveform can be reconstructed in the time domain. The mathematical details of the reconstruction procedure follow.

The waveform of an even periodic function can be written as

$$f(t) = \sum_{n=-\infty}^{\infty} a_n \cos(n\omega_0 t) = \sum_{n=-\infty}^{\infty} a_n \cos(2\pi f_n t) \quad (21)$$

and it can be used to represent the transmitted pulse. In (21), f_n is the n^{th} frequency component of the pulse, and the a_n 's are amplitude weighting coefficients which are collectively known as the spectrum, which is measured in the laboratory. Since physical pulses have finite bandwidths, the measured spectrum will have a lower frequency limit f_j and an upper limit f_k . Then (21) becomes

$$f(t) = \sum_{n=j}^k a_n \cos(2\pi f_n t) \quad (21a)$$

Reflection of the transmitted waveform by the target will modify the amplitude weighting coefficients. The new coefficients will be $a_n^s = \sigma_n a_n$, where σ_n is the power scattering coefficient at the n^{th} frequency component. Also, the detection system will filter the pulse. If a time constant τ_f is assumed and the detection system is modeled as a simple RC filter, the filter produces an attenuation [9] as a function of frequency of the form

$$A(f) = [1 + (2\pi f \tau_f)^2]^{-1/2} \quad (22)$$

Since the diode detector is a square law device, the output voltage is proportional to input power. If the pulse waveform is

written in units of power, comparisons of relative pulse strength to the oscilloscope display will be possible. The pulse power $p(t)$ can finally be written, accounting for the above modifications, as

$$p(t) = \left[\sum_{n=j}^k a_n^S A_n \cos(2\pi f_n t) \right]^2 \quad (23a)$$

or

$$p(t) = \left[\sum_{n=j}^k \sigma_n a_n A_n \cos(2\pi f_n t) \right]^2 \quad (23b)$$

III. SYSTEM DESIGN

In this section, the design of the pulse radar system is presented and discussed. The design of the system used for measuring the power spectrum (scattered power as a function of frequency) is also presented.

A. Pulse Radar System

As indicated in the theory section, the primary parameter of interest in a pulse radar system is the round-trip time for the pulse to travel to the target and back again to the radar. Thus, in addition to the generation of the pulses, the radar system must also incorporate an accurate display device. The device chosen for displaying the pulses is a sampling oscilloscope. It provides the means for direct measurement of the time differences between pulses, and it also possesses the bandwidth necessary for measuring and displaying very short pulses.

A pulse radar system for measuring short distances can be configured in block diagram form as shown in Figure 7. The pulse is generated by the pulse generator, amplified, and then transmitted via the transmitting antenna. The transmitted pulse is sampled by a directional coupler and displayed directly on the sampling oscilloscope. The sampled pulse is called the reference pulse since it

Figure 7. Block diagram of dual antenna radar system.

will be used as the timing reference for the transmitted signal. This system would be acceptable in applications which do not place a restriction on the size of the radar system

A radar system design which utilizes the same antenna for transmitting and receiving is shown in Figure 8. This design is generally more compact, and it is more appropriate for monitoring the bed-level in the fluidized-bed combustor. In addition to requiring only one antenna, this design replaces the directional coupler with a circulator. Its operation is essentially the same as the dual antenna system. The sample of the transmitted pulse is attained by circulator leakage. While an ideal circulator attempts to pass all of the power entering port 1 to port 2, practically there is some leakage of power from port 1 to port 3 providing a sample of the transmitted power at port 3. The leakage is approximately 1% of the transmitted power. The reflected pulse passes from port 2 to port 3 in the circulator. Thus, the need for a dual antenna system is eliminated.

The pulse generator is perhaps the most critical element of the system. While the distance from the antenna to the bed in a fluidized-bed combustor can vary according to the application, three meters is considered nominal. For this distance, the round-trip time is about 20 nanoseconds. Therefore, the maximum repetition

Figure 8. Block diagram of single antenna radar system.

rate should be $\text{pr}_{\text{max}} = 1/(20\text{ns}) = 50 \text{ MHz}$. Also, to ensure high resolution, the pulse width τ should be considerably smaller than 20ns. The width of the pulse leaving the system antenna should therefore be no more than 2ns.

The pulse generator circuit design is discussed in detail in Appendix A. The method chosen to generate the pulse uses the ability of a step-recovery diode to switch rapidly from a high-impedance to a low impedance state to generate a voltage impulse. This impulse is then amplified by the traveling wave tube amplifier to provide a radar system pulse whose width, measured by the oscilloscope at the output of the amplifier, is 1.8ns. The pulse is then filtered by a waveguide with a cutoff frequency of about $f_c = 6.56 \text{ GHz}$. The pulse is then comprised of a broad spectrum of frequencies above this cutoff frequency. The use of microwave frequencies enables the antenna to be small enough to fit easily inside the fluidized-bed combustor chamber while still providing high gain (greater than about 20 dB).

All of the components which follow the waveguide (which is considered a high-pass filter) also modify the spectrum of the pulse. The traveling wave tube amplifier is designed for operation in the frequency range 7.0 - 12.4 GHz, though it nonetheless broadens the pulse from 1.0ns to about 1.8ns. The antenna, isolators, and circulator are all designed for X-band (8.2 - 12.4 GHz)

operation, though they work quite well as low as the 6.56 GHz cut-off frequency. These waveguide components seem to provide no additional spreading of the pulse beyond that introduced by the traveling wave tube amplifier. However, the leakage path through the circulator is very dispersive, causing the reference pulse-width to be about 2.5ns.

The bed-level, as measured and computed by (3), neglects internal differences in the paths taken by the reference and reflected pulses. If the bed-level is referenced to the aperture of the antenna, the reference lengths shown in Figure 9 must be included in the bed level calculation. The appropriate equation is

$$h = R + (\ell_1 - \ell_2 - 2\ell_3 - \ell_4)/2 = R - \ell_{\text{ref}} \quad (24)$$

where

h = bed-level distance
 R = measured target range

The ℓ 's denote reference paths.

B. Power Measurement System

A power measurement system was designed and constructed to measure power as a function of frequency. The power measurement system provides information useful for determining the accuracy of analytical scattering models.

Figure 9. Radar system block diagram including reference paths.

A block diagram of this system is shown in Figure 10. The microwave signal source is used to produce a signal which can be tuned from 8.2 GHz to 9.6 GHz. This wide range of frequencies was chosen to provide an adequate level of confidence in the comparison of the analytical models to the measurements. The directional coupler placed after the signal source is used to sample the transmitted power and to provide a reference level. The sampling level (better known as the directivity) is -30 dB. Following the coupler is an isolator which prevents reflected signals from affecting the signal source. The other directional coupler is used to sample the reflected power at a level of -10 dB.

The power detection system uses two square-law detectors which have been matched to each other to raise the accuracy of the measurements. Since these are envelope detectors, the signal must be modulated for the detectors to operate properly. The standard modulation for which power measurement equipment has been designed is a 1.0 kHz square-wave. In addition, the power meters are specifically calibrated to square-law type detectors. To determine the ratio of transmitted to received power, both of the meters are referenced to the transmitted power at the beginning of a measurement cycle. Thus the power ratio is simply the difference of the two readings, since the meters are calibrated in dB.

Figure 10. Functional block diagram of power measurement system.

IV. RESULTS

In this section the results of the experimental measurements of the short-pulse radar system and those of the analytical models for pulse reconstruction are presented.

A. Range Measurement (level detection)

To assess the performance of the short-pulse radar system, measurements were made in a laboratory environment using static metal and limestone surfaces as targets. The metal targets were flat plates of aluminum while the limestone targets were boxes of different cross-sectional areas filled with limestone sand.

A typical waveform of the reference pulse is shown in Figure 11. The larger pulse is the desired reference waveform while the smaller pulse, following the reference pulse, is attributed to a reflection by the antenna. In general, the reference pulse is wider than the transmitted pulse, because it is obtained via the circulator leakage. This path possess discontinuities and lossy characteristics which are not present in the direct path, and they introduce dispersion and pulse width and waveform degradation. While the transmitted pulse is only about 1.0ns wide, the reference pulse has a width of about 2.5ns, and it is sufficiently narrow to provide a good timing reference at the peak of the pulse.

reference pulse
↓
internal reflection

Figure 11. Typical reference pulse display from radar system.

Typical waveforms of pulses reflected by smooth metal and limestone targets of 3'x3' cross-sectional areas are displayed in Figure 12. The response from the metal surfaces was always found to be considerably stronger than the response from the limestone targets. Early measurements indicated that the distance measurement error was less than about 10%. As the pulse generator was refined, the maximum error was reduced to less than 4.5%. The present average error is about 2.0%. Furthermore, there appears to be no correspondence between the error and the target type (metal or limestone).

Since the actual static fuel beds within a fluidized-bed combustor are not necessarily flat, experiments have also been performed for targets which were rough. Once again, the error in the distance measurement for rough surfaces, of both limestone and metal, did not correlate with the roughness of the targets for the surface roughness used. It should be noted that the intensity of the signal reflected by a given rough target was always weaker than that reflected by the same smooth surface at the same distance. For the limestone rough surfaces, the reflected pulse was often too weak to be detected with the available equipment. Similar indications were obtained from coal targets, due mainly to the weak reflective properties of coal.

a. Metal

b. Limestone

Figure 12. Typical display of reflections from metal and limestone smooth targets.

The performance of the system as a function of the wave polarization was also examined. In Table 1, a comparison of the range measurement for flat and corrugated targets is shown for horizontal (TE) and vertical (TM) polarizations. For a sinusoidally corrugated surface as shown in Figure 4, horizontal polarization denotes a wave with the electric field parallel to the corrugations, while vertical polarization occurs when the electric field is in the plane of incidence (xy plane). The surfaces were square plates with dimensions of 3'x3' and with sinusoidal corrugations in one direction only. According to the data of Table 1, polarization changes do not affect the error. Figure 13 displays the response waveform from the TE and TM corrugated targets. The signal intensity from the TE polarized target always appeared slightly stronger than for the TM case.

Experiments have also been performed on the smooth metal and limestone surfaces which were surrounded with cylindrical metal enclosures of square cross-sections. The enclosures were introduced to simulate the effect of the fluidized-bed combustor chamber walls. Three different enclosure cross-sectional sizes were being tested: 1'x1', 2'x2', and 3'x3'. The test results showing the radar accuracy are compiled in Table 2. Apparently, the introduction of the walls does not impair the precision of the distance measurement. However,

Table 1. Range measurements showing effect of polarization.

Target	Range To Target		
	Actual (in)	Measured (in)	%Error
smooth metal	74.4	70.6	5.1
corrugated metal, TM	74.8	71.2	4.7
corrugated metal, TE	74.8	71.2	4.7
smooth metal	63.3	61.5	2.8
corrugated metal, TM	63.0	60.9	3.3
corrugated metal, TE	63.0	60.3	4.3

a. TM case

b. TE case

Figure 13. Radar pulses from corrugated target for both vertical and horizontal polarizations.

Table 2. Comparison of range measurements with and without enclosures.

a. 3'x3' enclosure

enclosure	target	actual range	measured range	% error
no	metal	60.5	59.5	1.65
yes	metal	60.5	59.5	1.65
no	limestone	60.5	60.7	-0.33
yes	limestone	60.5	60.7	-0.33

b. 2'x2' enclosure

enclosure	target	actual range	measured range	% error
no	metal	43.0	41.8	2.79
yes	metal	43.0	42.4	1.39
no	limestone	43.0	41.2	4.18
yes	limestone	43.0	41.2	4.18

c. 1'x1' enclosure

enclosure	target	actual range	measured range	% error
no	metal	49.5	48.9	1.21
yes	metal	49.5	48.9	1.21
no	limestone	49.5	48.9	1.21
yes	limestone	49.5	48.9	1.21

*all ranges are given in inches

the pulse intensity for the smallest enclosure was stronger than for either of the two wider sizes. Also, the reflected signal appeared much sharper when the smallest of the simulated chambers was used. An example of this phenomenon is shown in Figure 14. This effect may be due to the waveguiding action by the walls of the enclosure which prevent significant spreading of the pulse energy.

During the testing of the short pulse radar system, it was noted that the strength of the signal reflected by a given target at times increased as the distance to the target increased. This behavior has been attributed to a multipath interference effect. To investigate this further, a series of measurements was taken over a closely spaced range of distances for smooth metal and limestone targets. The experimental results showed that the strength of the scattered pulse did indeed exhibit cyclical variations with distance, thus confirming the suspected multipath condition. However, while the pulse intensities varied periodically with range, no correlation was observed between the distance and the measurement error. The findings are summarized in Table 3. Also, the variation of the signal strength with distance was never significant enough to reduce a previously measurable signal to the system noise level. Ultimately then, the multipath action can be discounted.

a. No enclosure

b. With enclosure

Figure 14. Influence of 1'x1' enclosure on radar pulses for flat metal target at $h = 1.26\text{m}$.

Table 3. Measurement error versus distance for smooth metal and limestone targets.

a. Metal target

actual distance	measured distance	% error
26.0	25.9	-0.38
29.5	28.8	-2.37
33.3	31.8	-4.36
37.5	35.9	-4.27
40.1	40.1	-0.06
42.4	41.8	-1.36
44.3	43.6	-1.47
46.0	45.4	-1.30
47.5	45.4	-4.42
49.6	48.3	-2.67
51.9	50.1	-3.42
53.4	53.0	-0.70
55.0	54.8	-0.36
56.9	56.6	-0.48

b. Limestone target

actual distance	measured distance	% error
30.5	31.2	2.30
32.5	32.4	-0.31
35.1	35.3	0.50
37.9	37.7	-0.46
41.4	41.2	-0.42
43.3	43.0	-0.58
45.0	44.8	-0.44
47.8	48.3	1.15
49.3	50.1	1.73
51.3	51.9	1.27
53.6	53.6	-0.05
55.9	55.4	-0.85
58.5	59.5	1.71

*all distances are given in inches

Since the system is eventually intended for use in a fluidized-bed combustor, it is important to provide protection for the antenna from the high operating temperatures extant in the operating system. To accomplish this, the antenna can be covered by a high-grade dielectric thermal protector. The material chosen for this task is 0.4375 inch thick all silica (LI-1500) manufactured by the Lockheed Corporation [10]. This material was originally designed as a candidate thermal protector for the NASA Space Shuttle. Its electrical characteristics are such that it appears transparent at X-band frequencies to a temperature of about 2200⁰ F. Tests using the dielectric material with the short-pulse radar system indicate that the thermal protector does not degrade the radar system operation or its range measuring accuracy. This particular dielectric was also chosen for the frequency domain radar system previously investigated under this contract [2].

To check the accuracy of the analytical models, measurements of power return versus frequency were also made. In addition, the spectrum of the transmitted pulse was determined experimentally for use in synthesizing the reflected pulses from the analytical formulations. Since both of these measurements relate directly to the scattering models, the results of these experiments will be presented in the next section.

B. Analytical Results

To verify the experimental results and better understand the scattering characteristics of the targets as applied to a short-pulse radar system, this section presents the results of the analytical models. As an initial check on the analytical formulations, the theoretical results are compared with the measurement of scattered power versus frequency. The analytical models are then used to synthesize the reflected pulses (as described in section II.B.2.), which in turn are used to estimate the shape and strength of the reflected signal that can be expected for a given target. Thus the analytical results can be used as predictors of the experimental values. In addition, they can provide added confidence in the measurements.

The analytical models were first tested by comparing scattering patterns generated by the models to known patterns. A scattering pattern is a graph of the scattered power as a function of the scattering angle θ_n , as shown in Figure 4, for a given set of rough surface characteristics. An example of a scattering pattern generated by the space harmonic model, for a surface with $d = 0.0132\text{m}$, $\ell = 0.0530\text{m}$, $L = 1.125\text{m}$, $\lambda = 0.0375\text{m}$, and $\theta_i = 45^\circ$, is shown in Figure 15. Though the incident energy arrives at an angle of 45° , the reflected energy is directed toward all angles, hence the term

Figure 15. Typical scattering pattern generated by the space harmonic model.

"scattering" is used. If the surface were flat and infinite, all of the scattered energy would be concentrated in the $\theta_n = 45^\circ$ direction (often called the specular direction). It is observed that the major lobe in the specular direction is stronger than either of the other major lobes for this surface, though this will not necessarily be the case for all surfaces. It is also quite possible for the radiation in the backscatter direction (toward the source) to be very weak compared to that in the specular direction, contrary to the pattern of Figure 15. An example of weak backscatter is given in Figure 16, for a surface with $d = 0.00375\text{m}$, $\lambda = 0.375\text{m}$, and with all other characteristics the same as before. In addition to the weak radiation in the backscatter direction, only one major lobe exists for this surface compared with three for the previous surface. Predictions of the detailed scattering properties of a given surface are generally not possible on an intuitive basis, and they can usually only be obtained by rather complex analytical models such as these.

The scattering patterns generated by the space harmonic model compare well to previously published data [4]. As an additional comparison to the published data, the scattering coefficients have also been validated [11]. Furthermore, the energy conservation criterion shows that the energy error is typically much less than 0.1%, and it provides good confidence in the space harmonic scattering model.

Figure 16. Scattering pattern produced by the space harmonic model showing weak backscatter.

Scattering patterns generated by the physical optics model have been used to check the formulation of this model. These patterns can be compared to the space harmonic model patterns for the case of a perfectly conducting surface. Figure 17 shows a scattering pattern produced by the physical optics equations for the same surface used in Figure 16. Both of the plots have been normalized to the maximum value of the respective major lobes. A direct comparison of the two figures shows that they agree well in the specular region, which is the space of interest. However, the models do not match as well near the grazing angles ($\theta_n = \pm 90^\circ$) and in regions of low intensity. This is expected since the physical optics expression is approximate, and because the scattered power is of very low intensity in the grazing region. Thus the differences between the two models would be more evident in this range. Otherwise, the physical optics results provide a good match to the space harmonic results. The jump in the pattern of Figure 15 at about -32° is as yet unexplained and unresolved.

For the fluidized bed-level monitoring application, the designed radar system is operated with the antenna pointed downward for both transmitting and receiving. That is, the incident and scattering angles are both 0° . To generate data for this configuration, the analytical models were used to compute the reflected power level

Figure 17. Scattering pattern generated by the physical optics model.

as a function of frequency for several different targets. The measured data of power versus frequency were used to validate the analytical formulations and to instill a degree of confidence in the developed models. This was felt to be the most direct means of comparing the models to measured results. In Figure 18, the scattered power is plotted against the frequency for a metal surface with $d = 0.01\text{m}$, $\lambda = 0.0635\text{m}$, and $L = 0.66\text{m}$ using data from the experiment and from the space harmonic model. Figure 19 shows the same experimental data compared to the results from the physical optics model. Both of the theoretical curves were normalized to the measured power reflected at 8.2 GHz.

The dips in the experimental curves are attributed to a standing wave effect. The hypothesis was tested by performing the same measurements on a flat metal surface at two different distances. The same ripple structure was noted, and (as expected) its frequency of repetition increased as the distance to the target increased. Therefore, the standing wave explanation appears tenable.

With the analytical models generating results which are compatible with measurements, it was possible to use these models to synthesize the reflected pulses. The objective of the pulse synthesis (for which the scattering models have been included) is to provide a method for determining the effect of various surfaces on the operational capability of the radar system without performing time

Figure 18. Power backscattered by TE corrugated metal 1.07m from aperture, showing experimental results with space harmonic model results.

Figure 19. Power backscattered by TE corrugated metal target 1.07m from aperture, comparing experimental results with physical optics model results.

consuming and financially unattractive experiments.

Since the pulse reconstruction procedure requires a knowledge of the spectrum of the transmitted pulse, this spectrum was measured with a spectrum analyzer. The spectrum was measured at the output (ports) of the circulator (see Figure 8) for the transmitted pulse data and at the leakage port (port 3) of the circulator for the reference pulse data. The transmitted pulse and reference pulse spectra are shown in Figure 20 and Figure 21, respectively. Clearly, the leakage path through the circulator greatly attenuates the higher frequency components of the pulse, making the signal less broadband. This explains why the reference pulse is wider than the transmitted pulse. Using the transmitted pulse spectrum as the input to the pulse reconstruction algorithm, the reflected pulses can now be synthesized analytically.

The reconstruction algorithm can be tested by comparing the synthesized reference pulse width to that measured on the oscilloscope. In Figure 22 the synthesized reference and transmitted pulses are shown. The plots of these waveforms show some ringing which is not observed on the oscilloscope, because the rise times are too short for the oscilloscope to respond. Thus the pulses displayed on the oscilloscope appear as smooth envelopes with the same shape as that of the waveforms in Figure 22. A measurement of the pulse width for the reference pulse yields 2.5ns using the oscilloscope

Figure 20. Transmitted pulse spectrum.

Figure 21. Reference pulse spectrum.

a. Reference pulse

b. Transmitted pulse

Figure 22. Synthesized reference and transmitted pulses.

and 2.2ns using Figure 22. These values compare favorably, particularly considering the lack of adequate knowledge of the system bandwidth. The comparison of the above results, as well as others, generate confidence in the pulse reconstruction technique. From Figure 22, the pulse width for the transmitted signal is approximately 1.0ns. Attempts to measure this with the oscilloscope have yielded a width of 1.8ns. Thus the limited bandwidth of the detection system becomes very evident as the pulse width decreases. Hence, the actual width of the transmitted pulse is considered to be closer to the synthesized pulse value than to the measured one. Both of the reconstructed waveforms of Figure 22 exhibit the sharp peak which has been common throughout the experiments. Some examples of synthesized waveforms scattered by the radar targets will now be given.

Since the purpose of the pulse reconstruction is to determine the quality of the response from a given target, a baseline waveform must be chosen. The reflected pulse selected for this baseline is the response from a flat metal target, which is considered to be the surface that produces the "best" waveform. Figure 23 shows the synthesized waveforms for such a target for both of the analytical models (space harmonic and physical optics models). A target distance of 1.5m was assumed for these and all of the other reconstructed pulses. The two waveforms shown are identical in

a. Space harmonic model

b. Physical optics model

Figure 23. Reflected pulse from flat metal plate reconstructed using each of the analytical models.

every respect. Additionally, the reflected pulse is identical in shape to the transmitted pulse of Figure 22, though the levels are different due to spreading of the energy between the antenna and the target. The pulses which are reconstructed will be compared to the baseline reflected pulse displayed in Figure 23.

If the target is a rough surface, the reflected pulses should be weaker than that for a smooth surface. To illustrate that, Figure 24 shows the reflected signal for a weakly rough metal surface with $d = 0.00375\text{m}$ and $\lambda = 0.375\text{m}$. Once again, the pulse shape is not changed but the level is about 0.9 dB below the baseline (flat metal surface) level. Also, the results from the two analytical models compare exactly. If the metallic surface is made rougher with $d = 0.0100\text{m}$ and $\lambda = 0.0635\text{m}$, the pulses in Figure 25 will be generated. Now, the strengths predicted by the two models begin to diverge, because the physical optics approximation is no longer completely valid. That is, the space harmonic formulation shows a strength which is about 7.6 dB below the baseline value while the physical optics model predicts a value 9.5 dB below the baseline. However, the two models produce waveform shapes which are exactly the same and which exhibit no significant spreading when compared to the transmitted pulse. Thus the space harmonic model can be used to obtain the true level for metal surfaces and the

a. Space harmonic model

b. Physical optics model

Figure 24. Reconstructed waveforms reflected by a corrugated metal target with $d = 0.00375\text{m}$, $\lambda = 0.375\text{m}$, and $L = 0.66\text{m}$.

a. Space harmonic model

b. Physical optics model

Figure 25. Synthesized pulses scattered by a corrugated metal surface with $d = 0.010\text{m}$, $\lambda = 0.0635\text{m}$, and $L = 0.66\text{m}$.

physical optics model can be rescaled to this level to yield pulse strengths for nonconducting targets such as coal and limestone.

As an example of the reconstruction capability for nonconducting surfaces, Figure 26 shows the pulses reflected by both coal and limestone surfaces for the same surface roughness that was used in Figure 25. Again, the shape of the pulse has not changed, though the levels are very different. Signals returned from coal are seen to be about 3.3 dB below the limestone return, which in turn is about 4.3 dB weaker than the signal reflected by the same surface when it is made of metal.

As the surface is allowed to become rougher, the reflected pulse strength decreases, though not necessarily in a monotonic fashion. An example of this is shown in Figure 27, in which the synthesized pulse for a metal surface with $d = 0.015\text{m}$ and $\lambda = 0.0635\text{m}$ is plotted. This surface has the same period as the target of Figure 25, but the depth has been increased by 0.5 cm. Though the change in the depth was not physically large, the pulse strength decreased by 9.7 dB from the case with $d = 0.010\text{m}$. The depth variation of 0.5 cm is small on a physical scale, but it is a significant fraction of a wavelength at the harmonic frequencies of the pulse. Therefore, the dramatic decrease in the pulse strength for the deeper surface can be explained as a probable resonance effect which was not present for the shallower surface.

a. Limestone

b. Coal

Figure 26. Reconstructed pulses scattered by rough limestone and coal surfaces with $d = 0.0100\text{m}$, $\lambda = 0.0635\text{m}$, and $L = 0.66\text{m}$.

Figure 27. Pulse reflected from sinusoidally rough surface with $d = 0.015\text{m}$, $\lambda = 0.0635\text{m}$, and $L = 0.66\text{m}$.

A comparison of the measured pulses with the analytically synthesized pulses shows that though the reconstruction technique does not provide an absolute value of the pulse strength, it is quite useful in providing a relative measure of the signal level. For instance, the analytical methods show that the signal strength for limestone surfaces, rough or smooth, is always at least 4.3 dB below the response for the same surface made of metal. In addition, a coal surface will produce a response which is 7.6 dB below the metal surface reflected signal level. This explains why the interface detection, with the present configuration of the radar system used, was not as apparent for coal and rough limestone surfaces. It also suggests that if the transmitted pulse strength were raised by roughly 10 dB, it would be possible to use the short-pulse radar to detect the bed-level for these surfaces.

The analytical reconstruction results also confirm that the reflected pulse always retains its sharply peaked shape, regardless of the scattering surface type. Initially, it was expected that some surfaces would cause the reflected pulse to be distorted; however, the measured results indicate that the scattered waveform maintains its basic shape. This result, confirmed by experiment and analytically, is encouraging, because it indicates that a suitable timing reference will always be available. The most likely

explanation for this is that the transmitted signal is very broadband and a large number of the spectral lines would need to be attenuated for the signal to change its shape significantly.

CONCLUSIONS

A short-pulse time-domain radar system has been designed, constructed, and tested for use as a bed-level detector in a fluidized-bed combustor. Testing of the system has been performed in a static laboratory environment. The results indicate that the bed-level monitor is accurate to within 4.5% of the actual target range, with a 2.0% average error, over a variety of static metal and limestone test beds. A dielectric cover has been examined as a thermal protector for the antenna, and its presence does not degrade the system performance. The bed-level detector performs equally well when the targets are surrounded by metal enclosures, used to simulate combustor chamber walls, with cross-sectional areas as small as 1'x1'. For coal and rough limestone surfaces, the targets provide reflected pulses which are not as apparent as those from the other targets (using the presently available detection equipment).

Two analytical models have been developed to predict the pulse shape and strength, after it is scattered by a target. These models reduce the amount of laboratory testing, and they can be used to examine the performance of the system for a variety of targets. The analytical results found in this investigation compare well with each other in the regions of interest, and with those published in

the literature. Furthermore, the computed results of scattered power versus frequency, for a given target, compare favorably with those measured in the laboratory. Thus, within the bounds of applicability, the models can be utilized with confidence to perform parametric studies and to assess the effectiveness of the system.

The analytical results verify the pulse radar measurements and indicate that the limestone targets will always produce returns which are at least 4.3 dB below the strength of the pulse reflected by the same metal target shape. Moreover, the coal targets give reflected signals which are 7.6 dB weaker than the reflection by the metal target. The models also verify the encouraging result that the reflected pulse maintains the same shape as the transmitted pulse for all of the surfaces which have been investigated.

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APPENDIX A

Pulse Generator Design

This section describes in detail the design of the impulse generator. The characteristics of the step-recovery diode are discussed within the framework of pulse generation. Both the impulse generator circuit and the peripheral hardware are presented.

Frequency multiplication by non-linear devices is a well-known phenomenon. The step-recovery diode is such a device, and it has often been used as a frequency-multiplier or harmonic generator. The step-recovery diode develops a low device impedance for forward stored charge and a high impedance for reverse charge storage. Thus, the impedance varies as a function of time depending on the stored charge. A model has been developed [12] which represents the diode by two linear circuits depending on the stored charge. The model is idealized in the sense that during conduction (the forward stored charge interval) the diode capacitance is assumed infinite, while in the depletion mode (the reverse stored charge interval) the capacitance is assumed finite. The time required to switch between the two states is assumed to be zero and the minority carrier lifetime, τ , is assumed infinite. Figure A.1 shows the comparison of the ideal and actual circuit models for the step-recovery diode. The actual values of the above parameters

$$\begin{aligned}\tau &= \infty \\ C_F &= \infty \\ R_S &= 0\end{aligned}$$

1. ideal

$$\begin{aligned}R_F C_F &= \tau \\ R_S &\approx 1/2 \, \Omega\end{aligned}$$

2. actual

a. Forward conduction equivalent

1. ideal

2. actual

b. Depletion equivalent

Figure A.1. Ideal and actual equivalent circuit models of an SRD.

will be discussed following a description of the impulse generator circuit. It will be seen that the rapid switching from one impedance state to another is the quality necessary for impulse generation.

Operation of the impulse generator circuit will first be described qualitatively, with the detailed design equations presented later. The circuit to be analyzed is that of a basic impulse generator shown in Figure A.2. The input voltage is assumed sinusoidal, while the diode bias voltage V is a constant value. The inductance L is referred to as the drive inductance. As is expected, the inductor current passes through positive and negative cycles due to the sinusoidal input voltage. Typical inductor current and output voltage waveforms are shown in Figure A.3. When the current is positive, the diode is in the forward stored charge state. The amount of charge stored in a given cycle is equivalent to the area under the positive part of the current waveform of Figure A.3. As the current goes negative, the stored charge is swept out of the device. At the point when the areas (positive and negative) are equal, all of the stored charge has been removed and the diode then enters the depletion mode. This is the point at which the step-recovery diode abruptly switches from a low to a high impedance, thereby creating a discontinuity in the inductor current.

In general, the voltage across an inductor is given by

Figure A.2. Basic impulse generator circuit.

a. current

b. voltage

Figure A.3. Impulse generator current and voltage waveforms.

$$v = L \frac{di}{dt} \quad (A1)$$

where i is the current through the inductor. This equation states that a discontinuity in current results in a voltage spike across the inductor. This voltage spike is the impulse of interest. If the diode were not in the circuit at this point, the circuit would tend to ring. However as the voltage begins to return to a positive level, the diode conducts again and the cycle repeats itself. Thus, it is possible with this simple circuit to generate a pulse which is much shorter than the period of the sinusoidal voltage source. The generated pulse is rich in high frequency harmonics. The design equations for the pulse generator will now be discussed.

In practice, the pulse generator used is a modified version of the ideal model [14]. The changes are rather simple involving only a few extra components to better match the pulse generator to the voltage source and to provide a self-bias voltage for the diode. The modified circuit is shown in Figure A.4. The capacitor C_T resonates the inductance. The combination of L_m and C_m forms a matching network to match the input resistance of the sinusoidal source to the diode impulse generator input resistance.

The circuit design proceeds as follows:

- (1) Define the input frequency, f_{in} . In the bed-level monitor application, f_{in} is the maximum frequency of 50 MHz to avoid

Figure A.4. Actual impulse generator circuit.

range ambiguity. The period is

$$t_i = 1/f_{in} \quad (A2)$$

- (2) Choose the desired output pulse width, t_o . In this application t_o must be on the order of μns .
- (3) Calculate the multiplication factor

$$N = \frac{t_i}{2t_o} \quad (A3)$$

- (4) Find the total drive inductance

$$L_D' = \frac{4R_L^2 C_{TR}}{1 + \left(\frac{2\pi R_L C_{TR}}{t_o} \right)^2} \quad (A4)$$

where

R_L = load resistance

C_{TR} = diode reverse capacitance

The circuit drive inductance is $L_D = L_D' - L_p$

where L_p is the diode package inductance.

- (5) Calculate the damping factor

$$\xi = \frac{1}{2R_L} \left(\frac{L_D'}{C_{TR}} \right)^{1/2} \quad (A5)$$

(6) Determine the operation angle α

$$\tan \alpha = \frac{A_N G + B_N}{C_N G + D_N} \quad (\text{A6})$$

where

$$A_N = 2 \sin \frac{\pi}{N} + (2\pi - \frac{\pi}{N})^2 \sin \frac{\pi}{N} + 2(2\pi - \frac{\pi}{N}) \quad (\text{A6a})$$

$$B_N = 2 \sin \frac{\pi}{N} - (2\pi - \frac{\pi}{N})^2 \sin \frac{\pi}{N} + 2(2\pi - \frac{\pi}{N}) \cos \frac{\pi}{N} \quad (\text{A6b})$$

$$C_N = -2 + 2 \cos \frac{\pi}{N} + (2\pi - \frac{\pi}{N})^2 \cos \frac{\pi}{N} \quad (\text{A6c})$$

$$D_N = -2 + 2 \cos \frac{\pi}{N} - (2\pi - \frac{\pi}{N})^2 \cos \frac{\pi}{N} - 2(2\pi - \frac{\pi}{N}) \sin \frac{\pi}{N} \quad (\text{A6d})$$

$$G = \exp \left(\frac{\pi \xi}{\sqrt{1 - \xi^2}} \right) \quad (\text{A6e})$$

(7) Find the resonating capacitor C_T

$$C_T = \frac{1}{2} N^2 C_{TR} [1 + 2 \sin(\alpha - \frac{\pi}{N}) \sin \alpha] \quad (\text{A7})$$

(8) Calculate the impulse generator input resistance

$$R_{in} = \frac{\pi f_{in} L' D}{\cos \alpha \sin(\alpha - \frac{\pi}{N})} \quad (\text{A8})$$

(9) Determine the L-match network

$$L_m = \frac{\sqrt{R_{in}(R_g - R_{in})}}{2\pi f_{in}} \quad (A9a)$$

$$C_m = \frac{\sqrt{(R_g - R_{in})/R_{in}}}{2\pi f_o R_g} \quad (A9b)$$

where R_g is the output resistance of the source

(10) Finally, find the bias resistance

$$R_b = \frac{5\tau}{N^2 C_{TR}} \quad (A10)$$

where τ is the minority carrier lifetime for the diode.

The only parameters not specified above are the values of the blocking capacitors and the radio frequency choke. The blocking capacitors should be chosen to present a very low impedance at the input frequency. The inductor L_b is chosen to provide an impedance of several hundred ohms at the input frequency. The design procedure presented above is intended to provide an approximate set of values for the circuit components. Thus, in the design of the impulse generator, the desired results should be used more as guidelines than as actual values.

The diodes chosen were Microwave Associates MA43592 step-recovery diodes. The particular diode used in the final design

of the impulse generator has the following characteristics:

$$\begin{aligned}\tau &= 13 \text{ ns} \\ C_{\text{TR}} &= 0.214 \text{ pF} \\ L_p &= 0.4 \text{ nH}\end{aligned}\tag{A11}$$

The diodes were required to provide an impulse with significant harmonics higher than about 7 GHz, so that it would be amplified and propagated in X-band microwave hardware. An estimate of the highest frequency in the impulse is obtained by finding the inverse of t_s , the time required for the diode to switch impedance states. For the MA43592 step-recovery diode, the snap time is $t_s = 80\text{ps}$. Thus the upper limit on output frequency is about 12.5 GHz, which is sufficient for the application at hand.

To efficiently convert the input sinusoidal energy to the impulse, the diode minority carrier lifetime should be much longer than the input period. The maximum allowable input frequency is 50 MHz because of the range ambiguity limitation in connection with the short distances the bed-level monitor must measure. Hence, the 50 MHz value was chosen as the input frequency to minimize the losses due to minority carrier recombination. A lower frequency would have been more desirable from the standpoint of radar operation, but the resultant loss in output power would have been unacceptable.

Since the pulse width at the antenna of the radar system was required to be less than 2.0ns, the impulse generator was designed to produce a nominal pulse width of 250ps. This was felt to provide a reasonable margin for pulse spreading due to dispersive effects in the amplifier and waveguide components. Then, using $t_0 = 250\text{ps}$, the following circuit component values were calculated:

$$\begin{aligned} L_D &= 1.6\text{nH} \\ C_T &= 210\text{pF} \\ L_m &= 23.5\text{nH} \\ C_m &= 421\text{pF} \\ R_b &= 190\Omega \end{aligned} \tag{A12}$$

To minimize losses, the components were mounted on a teflon circuit board and connected by 50Ω microstrip lines. In addition, the SRD was mounted in a small hole and was secured by soldering directly to the microstrip conductor.

To generate an acceptable pulse, the circuit component values must be very carefully tuned. Otherwise the waveform will contain extraneous spikes and multiple reflections from impedance mismatches. Since the drive inductance is so small, this component was formed only by a short length (1/4 inch) of the microstrip connecting lines. The value of the inductor is adjusted by varying the length of the conductor between the diode and the capacitor C_T . After an

acceptable pulse is generated, further tuning can be performed by adjusting the diode bias resistor R_b .

In addition to the impulse generator circuit design, the devices which connect directly to the generator must also be carefully chosen. A block diagram of the pulse generator section of the radar system is shown in Figure A.5. To drive the diode so that the resulting output pulse is intense enough for the radar system application, the input signal is amplified to a level of about six watts. Attenuation of 6.0 dB is placed after the input amplifier to buffer it from the impulse generator circuit. At the output of the generator, the baseband pulse is high-pass filtered by an X-band waveguide adaptor whose cutoff frequency is 6.56 GHz. This filtering results in both a spreading of the waveform and in a great loss of power in the pulse, since most of the pulse energy is blocked by the filter. The signal is then amplified by the microwave amplifier. At the output of the amplifier, the measured pulse width is 1.8ns.

$f_o = 50 \text{ MHz}$

Figure A.5. Block diagram of pulse generator section of radar system.