MICROWAVE EDDY-CURRENT TECHNIQUES
FOR QUANTITATIVE NDE

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ABSTRACT

The objective of this research program is to develop microwave scattering techniques for obtaining quantitative information about the characteristic dimensions of surface cracks in metals. Work carried out during the past year has emphasized experimental techniques. The use of an IF bridge and a near-field dielectric probe to improve the sensitivity of our 100-GHz eddy-current measurement system is described. The results of measurements made on fatigue cracks in 2024-T3 aluminum using this improved system also are presented. The interpretation of these data to provide information about crack depth is explained in terms of a simple model. Finally, a theoretical approach for modeling the scattering from elliptically shaped cracks is outlined.

INTRODUCTION

In the low-frequency eddy-current testing of metals, currents are caused to flow in a test specimen by placing it in the magnetic field of an induction coil. The flow of currents is affected by the electrical properties and shape of the test specimen, and by the presence of discontinuities and defects. In turn, these currents react on the exciting coil and affect its impedance. Thus, the presence of a defect is determined by monitoring the test coil impedance.

Such eddy-current tests are conducted typically at frequencies of less than 1 MHz, where induction fields predominate and the electromagnetic wavelength is greater than 200 m. However, in quantitative NDE, where it is desired to obtain the defect dimensions from an analysis of the measured data, the use of such low frequencies does not provide the degree of sensitivity to changes in defect dimensions that is necessary for obtaining an accurate determination of these dimensions from an inversion of the eddy-current data. The problem of obtaining sufficient accuracy becomes more difficult as the flaws become smaller.

This problem would be alleviated if higher frequencies were used in eddy-current inspection. Thus, the work reported here addresses the possibility of conducting eddy-current measurements in the microwave-frequency regime (1 GHz to 100 GHz). Previous work using frequencies in the range 10 GHz to 30 GHz has shown that good sensitivity to small cracks can be obtained, and that a clear correlation exists between crack depth and the detected signal.

In using microwave frequencies, the radiation fields associated with the sensors become an important consideration, and the physics involved is best described in terms of fields and waves. For example, a defect should be thought of as producing a change in the scattering of electromagnetic waves from the metal surface. It should also be noted that since microwave frequencies cause currents also induced in the test object to flow essentially on the surface (i.e., the skin depth is typically less than 1 μm at 100 GHz), microwave eddy-current techniques for testing of metals are limited to surface inspection, e.g., to detection and characterization of surface-breaking cracks.

Work carried out in the first two years of this program has resulted in a significant improvement in our understanding of microwave eddy-current testing, and of eddy-current testing in general. We have developed a theoretical model for the far-field backscattering from a rectangular slot in a metal plate, and calculations using this model were found to be in good agreement with measurements made at 100 GHz. Given the cross-sectional dimensions of the slot, we were able to estimate slot depths accurately from single-frequency measurements.

Work carried out in the past year has emphasized the improvement of experimental techniques and the measurement of actual fatigue cracks. Recent improvements in our 100-GHz eddy-current measurement system, and the results of measurements carried out using this system, are described in this report. Also, a theoretical method of analyzing the scattering from elliptically shaped cracks is outlined.

MICROWAVE EDDY-CURRENT SYSTEM

The 100-GHz eddy-current system that we have developed was described in last year's annual report. In that system, we utilized a microwave bridge using an orthomode coupler for minimizing the backscattered signal produced by an unflawed flat metal plate. An orthomode coupler is a microwave component which transmits a linearly polarized wave traveling in the forward direction (from the source toward the test object), and couples out an orthogonally polarized wave traveling in the backward direction (from the test object toward the receiver). A crack can be detected by such a system because a crack converts part of the incoming wave to the orthogonally polarized wave, while a perfectly aligned flat plate does not. Background rejection achieved in this way (polarization filtering) does not depend on the distance between the probe and the plate, and thus is analogous to the behavior of the differential probe that is widely used in low-frequency eddy-current testing.

*References are listed at the end of the report.
Unfortunately, no component is perfect, and it proved very difficult to achieve more than 30 dB of rejection using the orthomode coupler augmented by tuning screws. Also, the tuning screws could not be made small enough in terms of a wavelength at 100 GHz, and so the tuning they provided was relatively coarse and difficult to repeat. For these reasons, we decided to modify the system so that the background signal could be cancelled after the signal had been down-converted to 3 kHz, where we expected that the desired amount of background rejection could be obtained more easily. A minor disadvantage of this type of scheme is that the nulling is dependent on the distance between the probe and the metal surface.

A block diagram of the modified measurement system is shown in Fig. 1. In addition to the 3-kHz nulling networks, we also added a circular polarizer. This component provides about 10 dB of backscatter rejection from a flat plate without the need for any fine tuning. Since the circular polarizer combined with a rectangular waveguide functions as a polarization filter like the orthomode coupler, this contribution to backscatter rejection is independent of the probe-to-surface spacing. Use of a circularly polarized wave also is advantageous because the crack signal is independent of crack orientation; however, this experimental convenience entails a 3-dB loss in signal power over that produced by an optimally aligned, linearly polarized incident wave.

We found that the sensitivity of the modified system was about the same as for the old system, but that the adjustments needed to achieve this sensitivity were much easier to carry out and to maintain. The gain of the 3-kHz amplifiers was about the same in both systems. With the improved background rejection in the modified system, we had expected to be able to increase this gain, but we were limited by ground-loop pickup and the finite isolation of the directional coupler. Of course, it should be possible to reduce such stray pickup with further effort.

![Fig. 1 Microwave Eddy-Current System](image-url)
The sensitivity of an eddy-current system is also affected by the ability of the probe to convert as large a fraction of its driving current as possible into eddy currents flowing in a region no larger than the flaw being examined. In the system tested last year, the probe used was a microwave lens (polystyrene) having a focal length of 7.6 cm and a focal-spot size of 3 mm. This probe provided nearly plane-wave excitation of the EDM slots, and thus was an important factor in the good agreement obtained between theory and experiment. The spot size of this probe was about optimum for the 2.5- and 1.25-mm-long slots that were examined. However, the scattering patterns for these slots are all broader than the lens pattern; thus, some of the scattered energy is lost, making this probe less efficient than we would like.

One way of obtaining a more efficient probe is to use a near-field probe, such as an open-ended waveguide, and bring it close to the flaw. Such a probe still concentrates its fields in a small region slightly larger than the waveguide cross section, but also captures more of the energy scattered by the flaw. It should be noted, however, that the attendant increase in sensitivity is accompanied by increased difficulty in performing a theoretical analysis of the probe-flaw interaction.

The sensitivity of a near-field probe can be increased further by using dielectric loading to increase the ratio of magnetic field to electric field at the surface of the metal being inspected. One method of dielectric loading is to attach a resonant dielectric sphere to the end of the waveguide. Resonance in the sphere is required so that, by proper positioning of the sphere, we can match the incident wave in the waveguide to the resonant mode in the sphere. An excellent dielectric material for this purpose would be sapphire (ε = 9). However, we could not easily obtain sapphire spheres of the correct size to be resonant at our system’s operating frequency of 100 GHz, and so we did not pursue this approach. However, we believe this approach to be feasible.

Another method of dielectric loading is to couple the air-filled waveguide to a dielectric waveguide. We constructed such a dielectric waveguide out of Delrin (ε = 3.7). A good match between the air-filled guide and the dielectric guide was obtained by tapering one end of the dielectric guide to a point. A photograph of the dielectric-waveguide probe is shown in the right-hand side of Fig. 2. The outer diameter of the Delrin probe is 1.1 mm. For comparison, the far-field lens probe is also shown in that figure.
Figure 3 provides a comparison of the eddy-current responses obtained when the lens probe and open-ended waveguide probe were used to examine the EDM slots studied previously. The two upper photographs were obtained using the far-field lens probe, and the two lower photographs were obtained using the near-field, circular-waveguide probe. The abscissa and ordinate on each photograph are proportional to the real and imaginary parts of the probe reflection coefficient, respectively. Each trace was obtained by mechanically moving the slots through the region illuminated by the probe. The background signal has been suppressed so that no lift-off trace is generated when the plate containing the slots is moved, i.e., the lift-off trace has been reduced to a point. When a slot enters the illuminated area, the spot on the storage-oscilloscope display moves away from its resting point and then returns to that point when the slot leaves the illuminated area. The maximum excursion and angular position of the resulting trace are indicative of the size of the slot.

The eddy-current responses obtained using the lens probe are essentially the same as those obtained previously with the orthomode-coupler scheme. Recall that all the slots are 0.25 mm wide, but vary in length and depth. Slots 1, 2, and 3 are 2.5 mm long, while Slots 4, 5, and 6 are 1.25 mm long. Slots 1 and 4 are 0.25 mm deep, Slots 2 and 5 are 0.5 mm deep, and Slots 3 and 6 are 1.0 mm deep (nominal). Slots 1, 2, and 3 produce responses that are well separated in phase, but the traces produced by Slots 4, 5, and 6 essentially lie on top of one another. These latter traces can be resolved (as was done previously) by tilting the plate to produce an observable lift-off trace. The responses from Slots 1, 2, and 3 have widely different phases because these slots are long enough to support a propagating waveguide mode that can reflect from the bottom of the slot. On the other hand, Slots 4, 5, and 6 support only evanescent modes (at 100 GHz).

The results obtained using the waveguide probe (see the lower photographs in Fig. 3) were somewhat unexpected. Qualitatively, the waveguide probe was indeed found to be more sensitive than the lens probe. The surprising feature of the waveguide eddy-current responses was, however, the asymmetry of the traces; i.e., their open-loop or kidney shape. We believe that this behavior is caused by coupling between two modes in the circular waveguide. The size of our waveguide is such that two modes can propagate: the TE_{11} (dominant mode) and the TM_{01}. The TE_{11} mode is the one eventually detected in the receiver but, in the near field, the slot apparently can couple these two modes and thus change the effective impedance seen by the dominant mode. The electric field in the TM_{01} mode is not symmetrical with respect to a plane containing the waveguide axis, and so the effective impedance depends on which side of this axis the slot is located.

The corresponding eddy-current responses obtained using the dielectric waveguide probe are shown in Fig. 4. These responses are very similar...
to those obtained using the lens probe (the traces obtained in the two cases are obviously rotated with respect to one another, but this difference is due simply to a different choice of phase reference). It is notable that the traces in Fig. 4 show very little of the open-loop shapes exhibited by the response of the open-ended waveguide. We attribute this result to the fact that only one mode can propagate in the dielectric waveguide, and thus there is very little excitation of higher-order waveguide modes by the slot.

MEASUREMENTS ON FATIGUE CRACKS

Three specimens of 2024-T3 aluminum containing fatigue cracks induced by tension-tension cycling were supplied to SRI by Dr. Otto Buck of the Rockwell International Science Center. The lengths of the cracks in the three specimens were measured under a microscope, and these crack lengths are listed in Fig. 5. The eddy-current responses of each of these cracks were obtained using a conventional eddy-current instrument and coil probe. Fig. 5 shows the eddy-current responses obtained at 483 kHz. Note that there is a qualitative correlation between the amplitude of each response and the crack length (and presumably the crack depth). However, the phase of each response seems to depend very little on the crack size. We will see that at microwave frequencies the phase of the eddy-current response can be a strong function of the crack size.

The 100-GHz eddy-current responses of these fatigue cracks were obtained by mechanically translating the aluminum samples past the fixed dielectric-waveguide probe. The cracks were relatively tight (the crack opening was on the order of 1 μm), and we found that they could not be detected with our system. Therefore, we built the four-point bending jig shown in Fig. 6. By turning the screw in the jig and bending the aluminum sample, we could control the degree of crack opening in the manner depicted in Fig. 7.

Photographs of the crack opening in sample A4 as a function of the sample’s surface displacement produced by bending are shown in Fig. 8. We began to detect the crack with our 100-GHz eddy-current system when the crack opening was about 3 μm, which corresponds to the middle photograph in Fig. 8. The right-hand photograph shows the maximum crack open-
Next, the sample was scanned slowly along a line containing the center of the crack. During this scan, the lift-off signal changes because the distance between the sample surface and the probe changes. This lift-off signal roughly follows a circular path on the oscilloscope, of which only a small part can be seen in the display (the diameter of the lift-off circle depends on the gain of the system). As the probe passes over the crack, the trace departs from its circular path (the amount of the departure from the circular path determines the degree of lift-off discrimination). Finally, as the probe moves past the crack, the response retraces its path, since the sample is bent symmetrically.

The striking feature of the crack signals is not so much the amplitude dependence on effective crack depth, but the phase dependence. For smaller crack depths, the crack-signal trace moves almost directly opposite to the lift-off trace. As the crack depth increases, the angle between the crack trace and the lift-off trace increases, progressing very rapidly through $90^\circ$. This angle opens up to $180^\circ$, and then reverses its direction back toward $0^\circ$. The complete phasor display that would be seen if the entire lift-off circle were visible is illustrated in Fig. 10.

This behavior can be explained in terms of a propagating mode inside the crack. Such a propagating mode can exist in this case because the crack length is greater than half a wavelength. At one frequency, the approximate equivalent circuit for such a mode is a simple short-circuited transmission line shunted at its input by a radiation conductance, $G$. From our previously developed theory, we know that scattering from the crack is proportional to the input impedance of this equivalent network. Therefore, the phase of the crack signal at the metal surface, $\theta_s$, is given by

$$\theta_s = \tan^{-1}\left[(1/GZ_0)\cot \delta d\right]$$

(1)

where $Z_0$ is the characteristic impedance of the propagating mode and $\delta$ is its propagation constant. The depth of the crack is $d$. In the phasor display, $\theta_s$ is the angle measured counter-clockwise from the crack phasor to a normal to the lift-off circle, as illustrated in Fig. 10. The variation of $\theta_s$ with crack depth calculated from Eq. (1) is shown in Fig. 11 for two values of $1/GZ_0$. The "high-Q" case, $1/GZ_0 = 10$, is more realistic for open cracks according to calculations using the theory developed last year. Qualitatively, the high-Q variation of phase is indeed what was observed during the gathering of data shown in Fig. 9. Furthermore, the calculations also show that the zero crossing of the phase is primarily determined by crack depth, and is.
Fig. 7 Illustration of the Variation of Effective Crack Depth with Bending Displacement

Fig. 8 Crack Opening as a Function of Surface Displacement, $S_o$

$S_o = 0$ in.  $S_o = 0.062$ in.  $S_o = 0.107$ in.
Fig. 9 100-GHz Eddy-Current Response of a Fatigue Crack in Sample A4 as a Function of Relative Surface Displacement, $S$
not a strong function of crack length or opening.

This result suggests that crack depth can be measured accurately with this technique, provided that the crack is sufficiently open to be detectable and have a high Q, and provided that the crack length is larger than one-half wavelength. Rather than bend the specimen, in practice one would vary the frequency (the measurement system would have to be constructed to have sufficient bandwidth) until the crack signal is normal to the lift-off circle. At this point, \( \beta d \) is an odd multiple of \( \pi/2 \), the ambiguity being resolvable from a knowledge of the crack length. Then, assuming that a crack model is available to provide an accurate value for \( \beta \), the depth is determined. An important feature of this technique is that it is self-calibrating, since the phase of the crack signal is measured relative to the lift-off circle.

**THEORETICAL SCATTERING FROM AN ELLIPTICAL CRACK**

In the theory developed last year we found that the scattering produced by a crack was inversely proportional to a suitably defined crack admittance. This crack admittance is the sum of a radiation admittance and a cavity admittance. If the crack is several times deeper than it is wide, we can assume that the radiation admittance does not depend on the interior geometry of the crack, but only on the geometry of the crack opening. The radiation admittance of a thin rectangular aperture like that shown in Fig. 12(a) has been calculated assuming the electric field in the crack mouth is given by

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*The modeling of elliptical cracks is discussed in the next section.
ELLIPTICAL COORDINATES IN Z-X PLANE

Fig. 12 Coordinate Systems for an Elliptical Crack

\[ \mathcal{E}(x,y,0) = \frac{1}{b} \sin [k(a-|x|)] \]  \hspace{1cm} (2)

and need not be discussed further. In Eq. (2), \( k \) is the free-space wave number and is equal to \( \frac{2\pi}{\text{free-space wavelength}} \).

Assuming this same aperture field, the cavity admittance is given by

\[ Y_c = -\int_a^b dx \int_0^b dy \frac{H_x^c}{b} \sin [k(a-|x|)] \]  \hspace{1cm} (3)

where \( H_x^c \) is determined from

\[ -j\omega_0 \mathcal{E}_x^c = \nabla \times \mathcal{E}_y^c . \]  \hspace{1cm} (4)

Here, \( \mathcal{E}_x^c \) is the interior electric field that is excited by the negative of the assumed aperture electric field. In other words, this interior electric field must satisfy the interior boundary conditions and be equal to the negative of the assumed aperture electric field in the aperture.

In our previous work involving a rectangular crack, the interior electric field was found by expanding the negative of the aperture field in a series of transverse-electric rectangular-waveguide modes. The result of this expansion was

\[ \mathcal{Y}_c = -j \sum_{q \text{ odd}} \left( \frac{8}{\pi} \frac{n_0}{k_b} \right) \left( \frac{k_b a^2}{r_q} \right)^{3/2} \cos^2 (ka/2) \coth (r_q a) \]  \hspace{1cm} (5)

where

\[ (r_q a)^2 = (a_k)^2 - (ka)^2 . \]  \hspace{1cm} (6)

When the interior of the crack assumes an elliptical shape like that shown in Fig. 12(a), we need to find the corresponding solutions of the Helmholtz equation

\[ \frac{\partial^2 \mathcal{E}_x^c}{\partial x^2} + \frac{\partial^2 \mathcal{E}_y^c}{\partial z^2} + k^2 \mathcal{E}_y^c = 0 \]  \hspace{1cm} (7)

which satisfy the interior boundary conditions and can be used to expand the aperture field.

This solution is best obtained by transforming to elliptical cylinder coordinates \((u,v)\) where

\[ \xi = c \cdot \cosh u \hspace{1cm} (8a) \]
\[ \eta = \cos v \hspace{1cm} (8b) \]
and
\[ c = \sqrt{d^2 - a^2} . \]  \hspace{1cm} (8c)

In terms of the rectangular coordinates, \((z,x)\), we can write:

\[ z = \xi \eta \hspace{1cm} (9a) \]
\[ x = \sqrt{(\xi^2 - c^2)(1-\eta^2)} . \]  \hspace{1cm} (9b)

The coordinates, \((u,v)\), for an elliptical crack are shown in Fig. 12(b). Assuming that the variables are separable in the new coordinate system so that

\[ \mathcal{E}_y^c(u,v) = U(u) \cdot V(v) \]  \hspace{1cm} (10)

the Helmholtz equation separates into two equations:

\[ \frac{d^2 U}{du^2} - [h - \left( \frac{c_k^2}{2} \right) \cosh 2u] U = 0 \]  \hspace{1cm} (11a)
and
\[ \frac{d^2 V}{dv^2} + [h - \left( \frac{c_k^2}{2} \right) \cosh 2v] V = 0 \]  \hspace{1cm} (11b)

where \( h \) is a separation constant. Referring to Fig. 12(b), the boundary condition on the perfectly conducting interior wall is

\[ \mathcal{E}_y^c(u_a,v) = 0 . \]  \hspace{1cm} (12)

Also, the assumed aperture field will only excite modes that are symmetrical about \( x = 0 \). Hence,

\[ \mathcal{E}_y^c(u_0/2) = \mathcal{E}_y^c(u_3/2) . \]  \hspace{1cm} (13)

Therefore, \( \mathcal{E}_y^c \) is even in \( v \) with period \( 2\pi \). This condition determines the separation constant, \( h \).

The solutions to Eq. (11b) are known as Mathieu functions. The even solution with period \( 2\pi \) is designated \( c_{2n+1}(v,q) \), where \( q = c_k^2/4 \) and \( n = 0,1,2, \ldots \). Each of these Mathieu functions is associated with a unique separation constant, which is designated as \( h = a_{2n+1} \).
The solutions to Eq. (11a) are called modified Mathieu functions. Satisfaction of Eq. (12) requires that both independent solutions to this second-order differential equation for \( h = b_{2n+1} \) be used. Therefore, we set

\[
U(u) = A_{2n+1} C_{2n+1}(u,q) + B_{2n+1} F_{2n+1}(u,q)
\]

where \( C_{2n+1} \) and \( F_{2n+1} \) are modified Mathieu functions of the first and second kinds, respectively. The constants \( A_{2n+1} \) and \( B_{2n+1} \) are determined by Eq. (12). They are:

\[
A_{2n+1} = C_{2n+1} F_{2n+1}(u_a,q)
\]

and

\[
B_{2n+1} = -C_{2n+1} C_{2n+1}(u_a,q)
\]

with \( C_{2n+1} \) an arbitrary constant.

In summary, the solution for the internal electric field is

\[
E_y(u,v) = \sum_{m=0}^{\infty} C_{2n+1} C_{2n+1}(v,q) Z_{2n+1}(u,q)
\]

where

\[
Z_{2n+1}(u,q) = C_{2n+1}(u_a,q) \quad \text{and} \quad \text{Fey}_{2n+1}(u,q)
\]

It can be shown that the \( Z_{2n+1} \) are orthogonal over the interval \([0,u_a]\), i.e.,

\[
\int_0^{u_a} Z_{2n+1} Z_{2n+1} \, du = 0
\]

if \( m \neq n \). Hence, the constants \( C_{2n+1} \) can be determined from the relation

\[
C_{2n+1} = \frac{\int_0^{u_a} \sin[k(a-c-sinh u)] Z_{2n+1}(u,q) \, du}{\int_0^{u_a} Z_{2n+1}^2(u,q) \, du}
\]

Series expansions for the Mathieu functions \( C_{2n+1}(v,q) \), \( F_{2n+1}(u_a,q) \), and \( \text{Fey}_{2n+1}(u,q) \) can be found in the literature.

The magnetic field in the crack mouth can be obtained from Eq. (4) expressed in elliptic coordinates, that is,

\[
-jw_0 \mu \mathbf{H}^c(\xi,0) = \frac{1}{2 \pi} \int_{\pi/2}^{\pi} \mathbf{E}^c(u,v) \, du
\]

From Eq. (9b) we find that

\[
dx = \xi \, du
\]

when \( n = 0 \). Hence, substitution of Eqs. (20) and (21) into Eq. (3) gives the result

\[
Y_c = [2j/\mu_0] \int_0^{u_a} \sin[k(a-c-sinh u)]
\]

This integral can be evaluated numerically, but such calculations have not yet been carried out.

SUMMARY AND CONCLUSIONS

Improvements in our 100-GHz eddy-current measurement system have been described, and the results of measurements made on fatigue cracks in 2024-T3 aluminum using this improved system have been presented. It has been shown that the phase of the 100-GHz eddy-current signal is a sensitive measure of crack depth, provided that the crack length is greater than about one-half wavelength. This phase measurement is a self-calibrating technique for determining crack depth. However, this technique requires that the crack be sufficiently open so that the component of crack impedance associated with energy propagating inside the crack is both detectable and larger than the skin-effect component of crack impedance, which is proportional to the surface impedance of the metal. To date, our 100-GHz system has not been sensitive enough to detect the skin-effect component.

In addition to this experimental work, we have also presented the theory necessary to compute the crack impedance associated with energy propagating in an elliptically shaped crack.

Therefore, we conclude that microwave eddy-current techniques can be used to accurately determine the dimensions of surface-breaking cracks in metals. However, more work is needed to improve crack models, system sensitivity, and particularly the practicality and flexibility of the physical implementation.

ACKNOWLEDGMENTS

Several other staff members at SRI made significant contributions to this work: Dr. C. M. Ablow provided the basic mathematical analysis of the elliptical crack, Dr. L. E. Eiselstein designed the four-point bending jig, and Dr. A. C. Phillips designed the 3-kHz nulling network.

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REFERENCES


Jim Martin, Chairman (Rockwell Science Center): Are there any questions?

Dick Elsley (Science Center): What is the skin depth in the aluminum?

Al Bahr (SRI): I don't have an exact number for that.

Jim Martin, Chairman: You said several microns.

Al Bahr: At 1 gigahertz, it's several microns, so at 100 gigahertz the skin depth is one tenth as large.

C.G. Gardner (University of Houston): You said when you change from linear polarization to circular polarization, you become insensitive to the crack orientation. What was the loss of sensitivity you associate with that?

Al Bahr: It's three db because you're lining up the polarization correctly only half of the time, i.e., the polarization is spinning around. So presumably you would be able to detect a particular crack with a smaller crack opening if the polarization were linear and aligned normal to the crack.

I should mention that another thing we are trying to do is to look at resonator probes in an effort to improve our sensitivity. We always need more sensitivity.

Frank Meunnenmann (Stanford University): Have you calibrated the actual impedance in ohms?

Al Bahr: No.

Jim Martin, Chairman: If there are no more questions, this concludes today's session.