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STATUS REPORT IN EDDY CURRENT
THEORY AND APPLICATION

by

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STATUS REPORT IN EDDY CURRENT
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I. INTRODUCTION

This status report on Eddy Current Theory and Application summarizes new developments and improvements in nondestructive testing techniques by means of eddy currents. It represents the work of the type carried out by the Nondestructive Testing Group of the Metallurgy Division during the past two years.

Eddy current testing techniques are still in their infancy due to a lack of basic knowledge concerning eddy currents and the lack of adequate instrumentation. The information contained in this report will assist in supplying some of the basic knowledge and will be the basis for providing some of the needed instrumentation.

II. THE IMPEDANCE OF A COIL NEAR A CONDUCTOR

A. Introduction

In the nondestructive testing of metals by eddy current methods⁽¹⁻⁴⁾ a small probe coil is brought close to the surface of the metal, and the apparent change in impedance of the coil is determined by various methods. Certain items of information may be extracted from this change in impedance; among them are (1) the thickness of a plated metal layer on the base metal, (2) the presence of occlusions or voids in the metal, and (3) the composition of an alloyed metal. The probe coil itself may have an air core, or it may have a magnetic core made of ferrite. Apparently very little information has been published on the changes in impedance of such a coil near a metal, and this paper is an effort to supply some of this information. This study is concerned entirely with non-ferrous metals.

Forster⁽⁹⁾ has made a theoretical analysis and experimental tests of the effect of a metal rod or tube as the core of a circular coil. An analysis and some experimental results will be presented here for the case of a coil with its axis perpendicular to the surface of a plane piece of metal. When the coil is energized, the resulting eddy currents flowing in the metal produce an apparent change in impedance or a reflected impedance change in the coil. This reflected impedance is a function of a number of variables, including the thickness of the metal, its conductivity and composition, the distance between the coil and the metal and the shape, size, number of turns and core material of the coil. The impedance will also depend upon the frequency if sinusoidal electrical currents are used, and upon the pulse shape and length if pulsed currents are employed.

B. Analysis

The principal parts of the analysis are given here, while the details are available in Appendix I. This analysis of a coil near a conducting surface assumes a coil of negligible width and thickness compared to its diameter. The inductance of such a coil isolated from all other objects is⁽⁶⁾

$$L_0 = \mu N^2 (2a - r) \left[\left(1 - \frac{1}{2k^2}\right) K(k) - E(k) \right] \quad (1)$$

where

a = radius of coil

N = number of turns

r = radius of wire

μ = permeability of the space surrounding the coil

$$k^2 = \frac{4a(a - r)}{(2a - r)^2}$$

$K(k)$ = complete elliptic integral of the first kind

$E(k)$ = complete elliptic integral of the second kind.

This inductance is reduced whenever a conducting plate is brought near the coil. If the axis of the coil is perpendicular to the plate, and the plate is a perfect conductor, the inductance of the coil will be given by

$$L_C = L_0 - \frac{2 N^2 \mu \sqrt{a(a-r)}}{k_1} [(1 - 1/2 k_1^2) K(k_1) - E(k_1)] \quad (2)$$

where

$$k_1 = \sqrt{\frac{4 a(a-r)}{(2a-r)^2 + (2d)^2}}$$

and

d = distance between the coil and the metal surface.

Now let it be assumed that the metal plate is of infinite thickness, but of finite conductivity. It is the finite conductivity of the plate which causes a reflected resistance in the coil and modifies the change in inductance from that given by the above equation. A correction will be added to take this into account.

The first step in finding the resistance reflected into the coil is to obtain an expression for the current density on the surface of the metal, assuming perfect conductivity. The usual skin-effect resistance (which is derived on the basis of a plane electromagnetic wave impinging on the metal) is multiplied by the square of the current density to obtain the power density. The power density is then integrated over the entire surface of the metal to obtain the total power dissipated. The resistance reflected into the coil is this power divided by the square of the current in the coil.

To obtain a correction for the change of inductance caused by the finite conductivity of the sample, the usual skin-effect internal impedance is substituted into the expression for the power dissipated over the whole plane instead of the skin-effect resistance alone. When this expression for real and reactive power is divided by the coil current squared, a complex impedance is obtained. The real part of this impedance is taken as the reflected resistance (R), and the imaginary part is the correction for the change (L_m) of inductance caused by the finite conductivity of the metal. The total change (L) of inductance caused by the proximity of the metal is

$$L = (L_0 - L_C) + L_m$$

where $(L_0 - L_C)$ is obtained from Eq. (2) and L_m from Eq. (9) of Appendix I. By using the skin-effect internal impedance for the case of one metal plated on another, (6) it is possible to determine how this type of composite conductor affects the impedance of the coil. The above analysis is subject to the approximations made, namely:

1. The current distribution assumed is that for a perfect conductor; this becomes more nearly true as the frequency increases.

2. The usual skin-effect internal impedance is derived, using the case of a plane wave impinging on the metal. This assumption is more nearly true when the coil is at a great distance from the metal.
3. The idea of substituting the skin-effect internal impedance in place of the skin-effect resistance is not completely justifiable, but becomes a better approximation as the distance between the coil and metal becomes greater.

However, this analysis does give results over a certain range of coil-to-sample spacing which are of the correct magnitude, and in certain cases gives very close approximations. The analysis gives successively poorer results as the spacing decreases.

C. Experimental Work

The experimental data were originally taken, not as a direct check on the approximate analysis, but as preliminary work with a view toward developing practical instruments for nondestructive tests on metals using the eddy current method. For this reason the probe was not constructed to approximate the idealized coil of the analysis, but was constructed as a practical probe for eddy current testing. Nevertheless, for probe-to-sample spacings great enough or for the higher frequencies, the probe used to obtain the experimental results included in this paper evidently represented a fair approximation to the idealized coil of the analysis.

Figures 1 and 2 show comparisons between experimental and analytical curves. L is the probe inductance under any conditions, and L_0 is the probe free space inductance. The change R is the difference of the probe coil resistance with the metal in proximity and the same resistance with the metal absent. It is normalized with respect to the free space inductance. The distance between the probe coil and the metal is indicated in mils on both the theoretical and experimental curves. The divergence between both curves becomes greater as the distance between the probe and metal approaches zero. Figure 2 is the same plot as that of Figure 1 except that the frequency is 6 kc instead of 1 kc. The agreement between the theoretical and experimental data is much better for the higher frequency or for smaller values of the skin depth.

Figure 3 is a comparison of the theoretical curve for a thick piece of copper at a frequency of 6 kc with the experimental curve for a piece of copper whose thickness is about one-third of the skin depth. Unfortunately, the experimental curve for a thick piece of copper was not taken, but would lie somewhat below the experimental curve for the one-third skin depth in a similar fashion to that for aluminum in Figure 2. This seems to indicate

Figure 1. Variation of the Impedance of a Probe with Spacing in Mils for Aluminum at a Frequency of 1 kc.

Figure 2. Variation of the Impedance of a Probe with Spacing in Mils for Aluminum at a Frequency of 6 kc.

Figure 3. Variation of the Impedance of a Probe with Spacing in Mils for a Thin Piece of Copper at a Frequency of 6 kc.

that the closer the current is confined to the surface the better is the agreement between experimental and theoretical results. All of these sets of curves diverge as the probe-to-sample spacing becomes too small. At small spacings, the approximations made in the analysis fail; also, the probe used on the experiments cannot be regarded as a very good approximation to the ideal coil of the analysis.

Measurements were made using an Owen-type A.C. bridge constructed of laboratory decade capacitors and resistors. A commercial audio oscillator furnished the low-distortion sine wave needed. The probe used consisted of 40 turns of No. 34 wire wound $1/64$ inch deep and $3/64$ inch wide on a $1/2$ -inch piece of polystyrene rod. It was necessary to use as small an input voltage as was possible in the detection in order to reduce the heating of the probe to a minimum. Changes in the ambient temperature resulted in heating or cooling of the probe, and the consequent change in the probe parameters appears to be one of the major difficulties in the experimental work.

Figure 4 is an experimental plot of the impedance components of a probe placed above 24-ST aluminum for different frequencies and different probe-to-metal spacings. The abscissa is the normalized change in probe resistance ($R/\omega L_0$); for a constant probe-to-metal spacing, this normalized change has a maximum value of a certain frequency. For example, from

Figure 4. Impedance Plot of a Probe for 24-ST Aluminum with Varying Frequency and Probe-to-Metal Spacing.

Figure 4 the maximum value of ($R/\omega L_0$) for a 5-mil spacing occurs at about 3 kc. It is also interesting to notice that the change in probe-coil resistance R itself has a maximum at another higher value of frequency. These maxima can also be obtained from the analysis, although the calculations are tedious. The normal depth of penetration of current in the metal is the same at a given frequency regardless of the scale of the coil. Except as noted, these curves are all for samples very thick as compared to this nominal skin depth. According to the analysis, two probes of different sizes but of dimensions proportional to each other should produce the same curves. Experimentally, the two curves do resemble each other

but are not quite the same. The explanation seems to be that actual probes have finite dimensions, whereas the analysis assumes a vanishingly small thickness and width of winding.

As an example of this effect the maxima observed in Figure 4 do not occur at exactly the same frequency if similarly shaped probes of different dimensions are used. In Figure 5 the same type of curves are plotted; this time, the conductivity of the sample, rather than the frequency, is the parameter. Since the curves of Figure 5 were all taken at the same frequency of 12 kc, the abscissa in this case represents the change in resistance divided by a constant. Notice how R at any given spacing passes through a maximum. For the probe used, this maximum occurred when a metal with a conductivity somewhat poorer than lead was used as a sample. The distances in mils represent the constant probe-to-metal spacing.

In Figure 6 are plotted the identical curves for brass and zirconium as were plotted in Figure 5. The small loop between them represents increasing thicknesses in mils of brass shims placed on zirconium, while maintaining a constant probe-to-sample spacing. The larger loop represents the same situation except for a smaller probe-to-sample spacing.

Figure 5. Impedance Plot of a Probe at a Frequency of 12 kc with Varying Probe-to-Metal Spacing and for Various Metals.

Figure 6. Impedance Plot of a Probe at a Frequency of 12 kc Showing the Variation with Thickness in Mils of Brass on a Zirconium Base.

The brass shims were pressed firmly against the zirconium, but it is believed that if the brass were plated or clad on the zirconium, the resulting curves would differ slightly from those reproduced because the air layer between the brass and zirconium would disappear. Theoretical curves for the loops were also calculated and were found to have similar shapes, except that they did not extend quite as far to the right. Also for small thickness of brass, the theoretical curves crossed over the curve for zirconium alone.

D. Conclusions

The analytical and experimental results illustrate the variation of the impedance of a probe coil near the surface of a metal. Since the analysis here presented seems to be effective only under certain conditions, a more exact analysis, that would take into account the coil geometry, the coil-to-sample spacing, the frequency and the sample conductivity, would be preferable. Once a more exact analysis is made, it will be useful in understanding and applying to practical uses the phenomena noted in this study. It is hoped that the results presented will be useful in the design and development of instruments using probe coils near conducting surfaces.

III. MINIMIZING THE EFFECT OF THE PROBE-TO-METAL SPACING IN EDDY CURRENT TESTING*

A. Introduction

In the nondestructive testing of metals by eddy current methods, a central problem has always been the reduction of the effect of varying probe-to-sample spacing on the test results. In the testing of samples of circular cross section by means of circular test coils, changes in coil-to-sample distance may be caused by diameter variations of the sample, while in the testing of plane samples, such as a plate with a point-type probe, the variation in probe-to-sample spacing may be caused by irregularities of the surface of the sample or by vibrations of the probe or sample. This chapter will be concerned exclusively with the testing of plane samples. Production test situations require constant movement of the probe parallel to the surface of the sample, and it is preferred that no contact occur between the probe or probe holder and the metal sample.

Some previous work on the reduction of the effect of varying probe-to-sample spacing has been done in Germany⁽⁹⁾ and in this country.^(10,11) This paper will be concerned with presenting another method of reducing this effect. This method uses a phase-angle meter which is corrected automatically for changes caused by spacing variations. Still another method is presented in Chapter VI.

B. Bridge Circuit

Theoretical and experimental work^(10,12) has shown how the impedance of a probe may vary in proximity with a plane piece of nonferromagnetic metal. Figure 7, curve A is typical of the type of curve which may be obtained as the probe-to-sample spacing is varied. The ordinate is the fraction of the free space inductance obtained when the metal is in proximity; the abscissa is the incremental change in resistance from the free space value divided by a constant, the free space reactance. The axis of the probe was maintained perpendicular to the plane of the metal. Curve B of Figure 7 is typical of the type of curve obtained when the probe-to-sample spacing is maintained constant and the conductivity is varied.

Any information to be gained from a test must be carried by changes in the probe resistance and inductance. Measurements of small changes in probe resistance and inductance are slow and tedious, so the information to be derived from a test should be presented in another, more convenient form. One way to do this is to connect the probe as an element of an alternating current bridge. This bridge can then be balanced with the

*See bibliography listing 14.

Figure 7

Plots of Normalized Probe Inductance Versus the Normalized change in Probe Resistance. In Curve A, distance is varied, while in Curve B, resistivity is varied.

probe in a particular test situation. Any change in the probe's situation in relation to the metal, such as a change in probe-to-sample spacing, or a change in sample conductivity, will disturb the balance of the bridge and cause a sinusoidal voltage to appear at the detector terminals. In the system presented here, an Owen bridge was used and balanced with the probe far away from any sample. The reason for this will be shown presently. As is seen in Appendix II, the impedance looking into the bridge output terminals can be given by

$$Z = \frac{K}{\left(1 - \frac{L}{L_0}\right) + j \frac{\delta R}{\mu L_0}},$$

assuming that the bridge is balanced with the probe at a great distance from a metal and then brought close to it. The quantity K is nearly constant for small changes in the inductance and resistance of the probe. The phase angle of the quantity in the denominator of the expression for Z may be represented by the set of phasors shown in Figure 8, where probe-to-sample spacing is varied. This phase angle also varies when changes in the other test variables, such as sample conductivity, are made. This is shown in Figure 9. In Figure 8, note how the phase angle change is relatively small

over a considerable range of probe-to-sample spacings. This is the reason for balancing the bridge with the probe far away from any sample; it is an advantage inherent in this system.

Figure 8

Diagram Showing How the Phasor $\left(1 - \frac{L}{\omega L_0}\right) + j \frac{\delta R}{L_0}$ Varies with Probe-To-Metal Spacing.

Figure 9

Diagram Showing How the Phasor $\left(1 - \frac{L}{\omega L_0}\right) + j \frac{\delta R}{\omega L_0}$ Varies with Sample Resistivity.

In Figure 10, curve A, the phase angle of the bridge output voltage relative to the driving oscillator voltage is plotted versus the relative magnitude of the voltage developed across the bridge output terminals. We shall henceforth denote this angle θ and the relative magnitude by E_1 . In Figure 10, curve A, a composite sample was formed by laying a 10-mil thick piece of brass shim stock on a thick piece of zirconium and clamping the two together. This type of sample is impractical but illustrative. To obtain Curve A, the probe-to-sample spacing was varied. In Figure 11 the probe-to-sample spacing was maintained constant, while the thickness of the brass shim on the zirconium was varied.

Figure 10

θ , The Phase Angle of the Bridge Output Compared to the Input Voltage Plotted Versus the Bridge Output Voltage (E_1). The sample used to obtain curves was composed of a 10-mil thick piece of brass shim on zirconium. Curve B represents the way the phase angle of the amplification characteristics of the compensator should change with varying bridge output voltage for perfect compensation.

Figure 11

Curve Showing the Variation of θ with E_1 for Samples Using Different Thickness of Brass Shims on Zirconium. The Probe-to-Metal Spacing was Maintained Constant.

C. Compensating Network

Consider a network, such as the one shown in Figure 12, with an amplification characteristic

$$K/\alpha = \frac{E_2/\beta}{E_1/\theta},$$

where $\beta = \alpha + \theta$. The system of minimizing the effect of varying probe-to-sample spacing presented in this paper requires a network in which α varies in such a way with changes in probe-to-sample spacing that the effect is to maintain β constant. Figure 10, curve A, gives a typical relationship that may exist between θ and E_1 as the probe-to-sample spacing is varied. What is done is to use E_1 to control the α of the network so as to maintain the sum of α and θ constant. The angle β then carries the test results, but they are not dependent on probe-to-sample spacing changes. If α for this compensating network is plotted versus E_1 , it will appear as curve B in Figure 10. Note that the sum of curves A and B produces a straight line.

Figure 12. Block Diagram of the System.

A compensating network such as the one mentioned will eliminate one of the effects of varying probe-to-sample spacing. Another effect to be considered involves the question of whether an irregularity that should be detected in a sample produces the same output from the test instrument at one probe-to-sample spacing as at another. In other words, will the sensitivity of the instrument change at different probe-to-sample spacings? An illustration will make this problem clearer. Take the case presented in Figure 10. Suppose the normal probe-to-sample spacing is 50 mils and the normal thickness of brass on zirconium is 10 mils. If the thickness of brass changes to 13 mils, a change in β of perhaps 1 degree might result. But, suppose that for some reason the probe-to-sample spacing is not 50 mils but rather 40 mils, will β still change 1 degree? It has been found that it will over a useful range of probe-to-sample spacings in this and many other situations. An example is given in Figure 13. Note how the change in β stays nearly constant when a 3-mil thick piece shim of stainless steel is placed upon a thick piece of zirconium while probe-to-sample spacing is varied from 60 to 100 mils, as shown in Figure 13.

Figure 13

θ Plotted Versus Varying Probe-to-Sample Spacings. Thick zirconium alone was used in the upper curve, while a 3-mil thick piece of stainless steel shim was used on the zirconium to obtain the lower curve.

Still another effect which should be considered is in a sense opposite to the one just discussed. Taking the same example, suppose the 3-mil change in the thickness of the brass has just occurred, giving a total of 13 mils of brass. The compensating circuit has been set to compensate for the effect of probe-to-sample spacing using the zirconium and 10-mil thick brass shim as a sample. Will the compensating circuit still compensate properly with the sample now composed of zirconium and the 13-mil brass shim? It will as long as the change in the characteristics of the metal sample is not too great. In the example just cited, compensation

would still be satisfactory for 13 mils of brass on zirconium, but probably not for 23 mils of brass. A curve could be drawn for 13 mils of brass on zirconium similar to curve A in Figure 10 and it would run nearly parallel to curve A.

A block diagram of the system is shown in Figure 12. A simple compensating circuit is shown in Figure 14. The plate resistance of the 6SL7 in connection with C_1 provides the actual phase shift. The input voltage (E_1) is rectified and applied to the grids of 6SL7. The plate resistance of the 6SL7 changes as the input voltage changes, thus changing the phase of the input voltage relative to the output voltage (α). Depending on the test situation more elaborate circuits may be needed. Two other possible schemes involve the controlling of the amount of feedback current flowing in an amplifier feedback loop, thus changing the over-all amplification characteristic of the amplifier, or varying the saturation in the case of a saturable reactor, thus changing its inductance to provide a phase shift which varies with the saturating current.

Figure 14. Simple Compensating Circuit.

Figures 15 and 16 give examples of the use of this system in measuring plating thickness. These examples are illustrative, but probably not too practical. Figure 15 shows a plot of the phase angle of

Figure 15

Phase Angle of the Output Compensator Plotted Versus Plating Thickness for Copper Plate on Stainless Steel.

Figure 16

Phase Angle of the Output Compensator Plotted Versus Plating Thickness for Nickel Plate on Uranium.

the output from the compensator plotted versus plating thickness. Using curves like these, the output from the phase meter could be continuously recorded, thus furnishing a permanent record of the plating thickness of the material. The separate curves of Figure 15 were obtained at three different probe-to-metal spacings of 5, 13 and 21 mils, respectively. This illustrates the use of the compensator in keeping the measurements nearly independent of probe-to-metal spacing. Many commercial phase-measuring instruments have relative accuracies of one degree at 20 kc, and some are much better than that. Assuming the 1° figure, the accuracy of measurement in Figure 15 is about 0.25 mil. Generally, the amount of care necessary in the design of the compensating circuit depends on the accuracy of the phase-measuring instrument available. Figure 16 shows the measurement of nickel plating on uranium. Nickel is ferromagnetic; this makes the measurement of the plating thickness much easier. Again three separate curves are given for the 5, 13 and 21-mil probe-to-metal

spacings. Probe-to-metal spacings of up to 200 mils are possible, but with some loss in sensitivity. In the case of the nickel-plated uranium, a phase-measuring instrument with a relative accuracy of 1° should give a sensitivity of at least 0.1 mil.

Because of the lack of a sufficiently accurate theoretical analysis of the problem, the optimum frequency to be used in measurements in a particular situation must be found by experiment. This involves very little trouble in the system described here because of the ease with which the frequency is varied. Choice of frequency is usually not too critical; usually there is a relatively wide band of frequencies which all give about the same sensitivity.

D. Conclusion

The main advantage of this system is that it is adaptable to continuous or assembly-line processes, since the probe is maintained a significant distance away from the sample and effects of distance variations are largely cancelled out. Its operation could easily be made almost completely automatic. It is also relatively simple and very stable. A disadvantage lies in the fact that phase angle-measuring instruments are expensive. They operate by precisely measuring very small time intervals. As yet, there are few commercial instruments capable of measuring time intervals of less than one microsecond accurately, so this imposes a limitation on the sensitivity of the system.

IV. A PULSED EDDY CURRENT SYSTEM FOR GAGING PLATING THICKNESS²⁰

A. Introduction

The nondestructive measurement of cladding thickness when the cladding metal and the base metal are nonmagnetic has presented a major problem for a number of years. There have been several different approaches to the problem, such as: ultrasonics, back scattering of beta or gamma rays, and eddy currents. At the present time, the eddy current method appears to have the greatest chance for success. Several eddy current instruments have been developed, but there is still a need for improved instruments.

This eddy current method utilized two different techniques in application. The first,^(12,14) employs sinusoidal eddy currents of a single frequency. This method uses comparatively simple, well-known techniques and instruments, but inaccurate phase measurements and thermal drifts are serious limitations. The second uses pulsed eddy currents containing a spectrum of frequencies. This method uses more complex techniques and instruments, but the phase-measurement problem is eliminated, and the thermal problem created by current in the probe coil is minimized.

In the first pulsed eddy current instrument,^(3,15) the cladding thickness was read from the screen of a cathode ray oscillograph. This instrument suffered from these disadvantages: it was not useful for continuously scanning a metal; the coating thickness information was presented on the face of an oscilloscope and could not be recorded by a recording voltmeter. A second instrument,⁽¹⁶⁾ overcame these difficulties to some extent by the use of photocells placed on the screen of the oscilloscope. It was desirable, however, to eliminate the use of the oscilloscope and photocells if at all possible, and a new system for transferring the information to a recorder was developed. It is the purpose of this paper to describe this system and to present the results of tests made with it.

B. Operation

A small coil near the surface of a metal to be tested established a magnetic field in the metal, and this field induced eddy currents in the metal. The magnitude and direction of the eddy currents depends upon the coil geometry, the coil excitation voltage, the distance from the coil to the metal, and the resistivity of the metal. A change in cladding thickness is a change in resistivity. The impedance of the coil changed with a change in the magnitude and direction of the eddy currents.

A small ferrite-cored coil, with its axis perpendicular to the surface of the metal, was used to project the field into the metal. The coil

windings were concentrated near the end of the probe to obtain maximum sensitivity to a given change in coating thickness. A cross section of the probe is shown in Figure 17. When the cladding thickness changed, the impedance of the coil changed. To obtain the cladding thickness information, this change in coil impedance was measured with the aid of a bridge circuit.

Figure 17. Probe Cross Section

The bridge circuit in Figure 18 was excited by discharging a charged capacitor through the thyatron. When the input pulse from the pulse generator triggered the thyatron, the 1000-micromicrofarad capacitor discharged through a pulse-shaping network and the primary of the transformer. This provided a bridge excitation pulse shaped like the positive half of a sinusoidal wave. The secondary of the transformer was loosely coupled to the primary and shielded from the primary.

The pulse projected into the coated metal should meet certain length requirements. The length is directly related to the basic time,⁽¹⁴⁾

$$T = t^2 \mu_1 \sigma_1 \quad ,$$

where t is the coating thickness. Representative values of T are given in Table I. For best results, the input pulse from the probe should be approximately five times T . From Table I this would necessitate an extreme range

of pulse lengths. Some compromise is necessary unless only one coating thickness and coating metal is to be used.

Table I
VALUES OF T
(Microseconds)

Coating Metal	Coating Thickness	
	5 mils	25 mils
Aluminum	0.75	18.8
Zirconium	0.045	1.12
347 Stainless Steel	28.1	702.0

Figure 18. System Circuit Diagram

If a slight unbalance is present, an unbalanced voltage exists at the bridge output terminals. A standard sample of metal was placed on one probe, and another sample of metal was placed on the other probe. Balancing adjustments on the bridge were made, and the bridge output voltage was nearly zero. An unbalance was added by changing the test probe air gap, and the output voltage pulse was as indicated in Figure 19a. One

Figure 19

Waveforms From a Cathode Ray Oscilloscope

major difficulty was variation of the bridge output voltage with probe-to-metal spacing. It was discovered experimentally that one point on the bridge output pulse did not vary in amplitude with the probe-to-metal spacing, as indicated in Figure 19b. This point will be called the crossing point. The amplitude of the crossing point does vary with the cladding thickness.⁽¹⁶⁾

An analysis of the bridge circuit was made to show that the crossing point does exist. The details of the analysis are given in Appendix III and the results are shown in Figure 20.

Here e_0 is the bridge output voltage and K' is a constant which does not change the shape of the curves. There was one

point on the calculated curves that did not vary with the probe-to-metal spacing. This checked with the experimental results in Figure 19b.

To determine the cladding thickness, a sample of the bridge output voltage had to be taken at the crossing point. The sampling technique, in other words the technique for separating the probe-to-metal spacing information and the cladding thickness information, will be described in the next two sections.

C. The Sampling Pulse

To generate a sampling pulse, an 884 thyratron was operated as a switch to discharge a charged transmission line through the characteristic impedance of the line, as indicated

Figure 20

Calculated Bridge Output Voltage

in Figure 18. The pulse from the bridge circuit and the sampling pulse were synchronized by firing both thyratrons with one pulse generator. The sampling pulse was positioned in time by a variable delay line in the grid circuit of the thyatron.

The duration (t_0) of the sampling pulse was calculated from the value of the total storage capacitance (C_s) and the characteristic impedance (R_0) of the line. The network was assumed lossless and was charged to a voltage E . The amplitude of the pulse was $E/2$. Since the energy originally stored in the capacitance (C_s) was delivered to the load resistance (R_0) during discharge,

$$\frac{1}{2} C_s E^2 = \left(\frac{E}{2} \right) \left(\frac{E}{2R_0} \right) t_0$$

or

$$t_0 = 2 R_0 C_s$$

The duration (t_0) for 9 inches of RG-58/U coaxial cable was 2.29×10^{-9} second.

The actual pulse shape and length were largely determined by the 884 thyatron. The ionization time was about 10×10^{-9} second and this increased the length of the pulse. The deionization time of the 884 thyatron had a very small effect on the pulse length. The transmission line discharged very rapidly, and when the plate voltage dropped to zero, the tube extinguished. Figure 19c shows the sampling pulse. The pulse length was about 15×10^{-9} second and the amplitude was about 15 volts.

D. The Mixer and Voltmeter

The mixer in Figure 18 summed the bridge output pulse and the sampling pulse. The high input impedance reduced the interaction of the two pulses to a minimum. The cathode resistance has to be small to preserve the original waveforms. Much of the energy of the sampling pulse was transferred to the cathode of the mixer through the interelectrode capacitance (C_{gk_1}). The output of the mixer is as indicated in Figure 19d.

The sum of the pulse voltage from the bridge circuit at the crossing point and the sampling pulse voltage gave the indication of coating thickness. The peak voltmeter in Figure 18 measured this sum. During the time of the sampling pulse, a charge built up on the 150-micromicrofarad capacitor. During the off part of the cycle the charge decayed at a rate given by the time constant of the capacitance and the input resistance of the 6SF5 triode. Since the input resistance of the 6SF5 triode was very large, the time constant for the decay was very large, and the charge on this

capacitor decayed very slowly. As a result, the peak of the sampling pulse was measured with no interference from the peak of the bridge output pulse. The difference amplifier in Figure 18 provided a high gain and a simple method for balancing out the direct voltage that contained no coating thick-

ness information. With a proper selection of voltmeter, the coating thickness can be read directly in mils. The results for nonmagnetic stainless steel coated on brass are given graphically in Figure 21 where d is the probe-to-metal spacing. It should be noticed that the readings are nearly independent of the probe-to-metal spacing.

Figure 21

Output of the Entire System

One difficulty that was experienced was a seemingly random variation in the output indication. The use of regulated power supplies helped this considerably, but there was still a small amount of the random variation left. It is probable that further work on the supplies would eliminate most of this difficulty.

E. Conclusions

The calculated bridge output voltage indicated the presence of the crossing point. The major error in the calculation was introduced by the assumptions of the probe characteristics. A true equivalent circuit for the probe has not been found and this will have to be done before exact calculations can be made.

This eddy current instrument measures cladding thickness fairly independently of probe-to-metal spacing. Pulsed eddy currents have one major advantage over sinusoidal eddy currents, namely: a much higher peak current can be applied to the probe coil without heating the coil if the proper pulse length and repetition rate have been selected. The one major disadvantage of this pulse system is: the pulses must be very stable in amplitude and time.

V. EDDY CURRENT TECHNIQUES FOR TESTING LIQUID METAL BONDING²¹

A. Introduction

The primary purpose of the Experimental Breeder Reactor-II (EBR-II), which is initially a U^{235} fueled, unmoderated, sodium-cooled reactor with a power rating of 60,000 kilowatts (heat), is to demonstrate the engineering feasibility of a fast reactor system which will be economically competitive with existing power sources within the United States. At a later date it is expected to use plutonium as a fuel.

The uranium-fissium alloy fuel pin for EBR-II is 14.22 inches long and 0.144 inch in diameter. This fuel pin is encased in a thin-walled (0.009 inch), stainless steel tube, approximately 17 inches long and 0.175 inch in diameter. The 0.006-inch annulus between the fuel pin and the wall of the tube is filled with sodium which serves as a bonding and heat transfer medium. A stainless steel spacer wire, 0.049 inch in diameter, is attached to the stainless steel tube.

A pyrometallurgical procedure, because of its inherent suitability for incorporation in a compact, on-site, low-inventory, economic pilot plant, will be used for reprocessing the fuel. Such a method does not involve complete decontamination of the fuel. Because of the intense radiations, it will be necessary to refabricate and assemble the fuel elements and subassemblies by remotely controlled operations in heavily shielded facilities. This factor has influenced the details of the design and construction of the fuel-element assemblies.

The presence of voids or gas pockets in the sodium bonding layer between the uranium rod and the tubing wall will impair the heat-transfer characteristics of the fuel element. Heat-transfer calculations indicate that the maximum size of void acceptable for EBR-II service is a round void $3/16$ inch in diameter or a rectangular void not greater than $1/8$ inch in width. The fact that the fuel element may not be symmetrically located in the stainless steel can is not considered to be objectionable. Likewise, variations in the inside diameter or the outside diameter of the stainless steel cladding are not considered objectionable.

A number of EBR-II fuel elements were tested by two eddy current systems: the Dumont Cyclograph, and an eddy current testing instrument developed at the Argonne National Laboratory. A description of the Cyclograph can be found in literature references.^(17,18) A description and theory of operation for the instrument developed at Argonne National Laboratory is described in Chapter III, entitled "Minimizing the Effect of Probe-to-Metal Spacing in Eddy Current Testing." A test frequency of 25 kc was found to be most suitable, although the choice of frequency was not at all

critical. Both circular-type coils and point probe-type coils were investigated. Because of the way of specifying the maximum void size permissible in various directions along the fuel element, the circular-type test coil was chosen.

B. Results

Some results using the point probe are shown in Figure 22. This figure shows that the point probe was able to detect a void on one side of the fuel element from positions, not only directly above the void, but also with the probe at angles as much as 180 degrees away, measuring around the circumference of the fuel element. None of the voids in pin M-3 covered more than about 60 degrees around the circumference. The top trace was obtained using the circular coil and the lower four traces were obtained using the point probe which had a diameter of $1/4$ inch. Each successive test was made with the fuel element rotated 90 degrees from its previous position, starting at an arbitrary zero degree point. The voids in these traces are the short peaks and valleys. The long rises and falls are caused by variations in the thickness of the sodium layer under the probe, due to the uranium rod being warped or off center. This effect is partially cancelled because of symmetry when the circular test

Figure 22. Defect Indications

coil is used. The traces in Figure 22 were obtained by the instrument developed at Argonne National Laboratory. At frequencies in the neighborhood of 25 kc, the eddy currents flowing in the stainless steel have a very minor effect compared to those flowing in the sodium because of the thinness of the can and the much greater conductivity of the sodium. The test coil effectively "sees" through the can.

Some of these fuel elements were stripped by the simple expedient of cutting the can open at one end and then peeling off the stainless steel in a spiral, using a pair of long nose pliers. It was necessary to identify and photograph the voids as quickly as they appeared, since the sodium oxidized very rapidly and voids quickly became unidentifiable or changed in appearance. Because of the crudeness of the stripping procedure, it was difficult to determine if all the voids extended from the can to the uranium, but apparently the majority did. Figures 23, 24, and 25 show typical traces.

Both eddy current instruments detected every void of significant size. As far as the detection of voids was concerned, correlations between stripping results and the test was practically complete except for voids at the extreme end of the uranium rod. Occasionally a small void located near a large void would not appear on the trace, or two small voids would appear as one because of the limited definition of the test coil.

C. Conclusion

As can be seen from Figures 23, 24, 25e and 25f, there is a definite correlation between the amplitude of the void indication and the length of the void around the circumference, and, naturally, a correlation between the length of the void indication on the recorder trace and the actual length of the void along the rod axis. As usual, the finite size of the test coil and its area of influence made interpretation of the recorder traces difficult for the smaller voids, and blurred the edges of the larger ones as shown by the testing instrument. The trace of the void on the recorder paper is designated by a letter so that it can be compared with the photograph of the void also designated by the same letter. These recordings were made using the circular-type coil.

As mentioned previously, because of the intense radiations it will be necessary to test the refabricated fuel element in a heavily shielded enclosure, while the fuel element is handled by remotely controlled operations. The radiation level in the process cell is expected to be of the order of 10^6 roentgens per hour. The temperature of the fuel element due to gamma heating will be approximately 100°C . Fortunately only the test coil of the eddy current instrument need be placed in the process cell.

DISTANCE ALONG THE PIN

Figure 23. Typical Tracing and Defect

DISTANCE ALONG THE PIN

Figure 24. Typical Tracing and Defect

Figure 25. Typical Tracing and Defects

In order to determine the effect of irradiation on the test probe, typical probes were made and placed in the Gamma-Irradiation Facility at the Argonne National Laboratory. Coil forms were made from Micarta. After 10^9 -roentgen radiation the Micarta became brittle, but was still serviceable. Formvar-covered wire (#40) was used for winding the coils. There were no observable or measurable effects on the wire or its insulation during irradiation.

Because the sodium will be molten it will be necessary to test the fuel element while held in a vertical position. Since the test coils are temperature sensitive, it will be necessary to be certain that all fuel elements are tested at the same temperature. A small oven will be provided to assure that elements will be at the same temperature before testing.

Because the test of the EBR-II fuel elements for voids in the sodium layer is a production application, the test equipment must be designed for maximum reliability, stability and simplicity of operation. Automatic operation is contemplated so that interpretation of the test results must depend on the instrument, not on the operator.

VI. A MULTI-FREQUENCY EDDY CURRENT TESTING SYSTEM

A. Introduction

The application of eddy current methods to the problems of nondestructive testing has always been hindered by the difficulty of separating the useful information produced by the test from the unnecessary information which is always present. In testing metal specimens, such as reactor fuel elements, which are usually paramagnetic, it is usually necessary to suppress the effect of variable spacing between test coil and metal over a useful range of spacings. A number of methods have been used to accomplish this (see Chapter III). Some of these systems lack sufficient sensitivity to undertake many of the important tasks to which eddy current methods are applicable; others utilize elaborate and expensive laboratory equipment, which causes the testing system to be unduly complicated and expensive. Still others depend on specially designed test coils or probes useful only at one frequency. Unless a large number of test coils or probes are available, the flexibility of the system is seriously limited.

The system described in this paper was developed primarily for the measurement of the thickness of the cladding or plating of one metal on another. However, it has many other applications, such as the measurement of wall thickness of metal tubing with extreme accuracy, and the detection of subsurface cracks and voids in metals. This system has a much greater sensitivity than that described in Chapter III.

Eddy currents of two or more separate frequencies are induced to flow in the sample to be tested. One frequency is used to obtain the information required, the other frequency is used to compensate for the effect of variations in test coil-to-metal spacing and other undesirable effects. An application of this idea will be described in detail and should be sufficient to make the technique clear.

B. An Application of the Dual Frequency System

This test problem required an accurate measurement of the thickness of the Zircaloy-2 cladding on EBR-I Mark III fuel elements. These rods measured about 0.404 inch in diameter and were tested in various lengths. They were clad with a nominal 20 mils of Zircaloy-2 on a uranium-2 w/o zirconium core. The resistivity of the core was approximately $72 \mu\text{ohm-cm}$, while that of the cladding material was in the neighborhood of $52 \mu\text{ohm-cm}$. Since measurements of this type require relatively great sensitivity, an instrument which does not provide some method of compensating for varying probe-to-metal spacing cannot be used in a production application. It was expected that the cladding thickness would always range between 15 and 25 mils. Experiments showed that frequencies between 25 and 50 kc were suitable for the measurements. It was also determined that eddy currents of a frequency of 260 kc were relatively ineffective for the measurement of clad thicker than 15 mils.

C. Basic Features of the System

It is simpler to consider each channel of the system separately. In the block diagram, Figure 26, the elements of each channel are shown according to their function and not necessarily according to position in the circuit diagram. In the high-frequency channel, a sinusoidal voltage of 262 kc is generated and applied across the probe winding. More will be said about the probe later. The probe forms part of a bridge circuit which is balanced when the probe is far away from any metal. Any change in the field about the probe causes an apparent change of the impedance seen looking into the terminals of the probe. This unbalances the bridge and causes a voltage to appear at its output terminals. Changing the distance from the probe to the metal produces a voltage which is amplified and rectified. Figure 27 shows how this voltage may vary, depending on the spacing from the probe to fuel element. The high-frequency channel output is practically independent of the cladding thickness since a frequency of 262 kc is too high to be effective for the measurement.

All that has been said concerning the high-frequency channel can be said for the low-frequency channel, except that it produces a voltage which is a function of the same variables as the high-frequency channel plus an additional variable, the cladding thickness. In addition, the low-frequency channel amplifier is designed so that its gain may be varied by means of a D.C. voltage which is normally supplied from the high-frequency channel output. The high-frequency channel output voltage varies with probe-to-metal spacing and changes the gain of the low-frequency channel just enough to compensate for the effect of varying probe-to-metal spacing on the low-frequency channel output. Figure 28 shows a typical curve of low-frequency channel output plotted versus distance from a fuel element. The change in low-frequency channel sensitivity produced by a change in cladding thickness from 21 mils to 18 mils is plotted in Figure 29, as a function of probe-to-metal spacing. From Figure 28 and 29, it can be deduced that the voltage produced by the high-frequency channel must compensate for at least two separate effects that could cause an error in the measurement of the cladding thickness. One is the actual change in the magnitude of the low-frequency channel output caused by varying spacing. The other is the dependence upon probe-to-metal spacing of the low-frequency channel sensitivity to a given change in cladding. Figures 30 and 31 show how Figures 28 and 29 look after probe-to-metal compensation has been added. Details of the compensation process are given in Appendix IV.

D. Probes

The probes used in this application were wound on 1/4-inch or 3/16-inch cylindrical ferrite forms. The winding consisted of a single layer of #40 enameled wire. Single-layer windings usually possess smaller temperature coefficients of inductance than do multi-layer

Figure 26. Block Diagram of the System

Figure 27. The High Frequency Channel Output Voltage Plotted as a Function of Probe-to-Metal Spacing.

Figure 28. Amplified Low Frequency Bridge Output Voltage Plotted as a Function of Probe-to-Metal Spacing.

Figure 29. Low Frequency Channel Sensitivity Plotted as a Function of the Probe-to-Metal Spacing. No Compensation was Used. The Actual Change in Cladding was 3 Mills.

Figure 30. Low Frequency Channel Output Voltage Plotted as a Function of the Probe-to-Metal Spacing. Compensation Used.

Figure 31. Low Frequency Channel Sensitivity Plotted as a Function of the Probe-to-Metal Spacing. Compensation Used. The Actual Change in Cladding Thickness Was 3-Mils.

windings. A typical probe is shown in Figure 32 in position over a segment of fuel rod. Probe inductance is not critical; a number of probes of widely different inductances have been used in this particular test problem with no important adjustments necessary except balancing the bridges and the re-adjustment of the high-and low-frequency channel gains. This is the work of but a few minutes.

Figure 32. A Typical Probe Shown in Position Over a Fuel Rod. The Probe Is to the Immediate Left of the Small Follower Wheel.

E. Circuitry

The circuit diagram of the low-frequency channel is shown in Figure 33 and that of the high-frequency channel in Figure 34. Since the probe is a common element of each bridge, the low-frequency voltage is

Figure 33. Circuit Diagram of the Low-Frequency Channel.

Figure 34. Circuit Diagram of the High-Frequency Channel.

present in the output of the high-frequency bridge, and vice versa. These undesired frequencies must be filtered out. This is done in the early stages of the amplifiers, since the bridges are not balanced with respect to these frequencies and their relatively high amplitudes would overload later stages of the amplifiers. The low-frequency bridge is always balanced first since it affects the balance of the high-frequency bridge to a large degree, while the reverse effect is only slight. The bridges are balanced by means of the variable resistors; (R_1) and (R_3), in each bridge.

The crystal oscillators furnish the sine wave voltages needed to drive the probe. An unusual amount of frequency stability is not needed; crystals are used in this instrument because the amplifiers in earlier models were sharply tuned. The crystal oscillators have been carried over from the earlier models. RC oscillators have been used successfully. The oscillators should provide voltage of less than 5 per cent distortion. Changes in oscillator output amplitude are compensated for by the 1N34 diodes in each channel and their associated circuitry. Part of the oscillator output is rectified and fed to the 6AQ6 stage in each channel. Any change in oscillator output changes the gain of this stage so as to maintain the output of the 6AQ6 stage independent of routine variations in oscillator output.

The 262-kc voltage is filtered out of the low-channel amplifier by the low pass filter (L_1 C_3 C_4). The double-tuned transformer (T_1) performs an analogous function for the high-frequency channel. There are a number of other tubes which could have been used in the low-frequency channel instead of the 6L7's.

The output of each channel is fed to separate difference voltmeters with several ranges. Two of the ranges on each voltmeter are useful when balancing the bridges. The most sensitive scale on the low-frequency channel difference voltmeter is calibrated directly in mils of cladding, while the most sensitive scale of the high-frequency channel difference voltmeter is used to keep check on the probe-to-metal spacing. An output from the low-frequency channel can be connected to a recording voltmeter.

F. Mechanical Conveyor

The instrument and conveyor to move the rods past the probe are shown in Figure 35. Skewed wheels equipped with rubber tires spin the rod and at the same time provide it with motion along its longitudinal axis. The pitch of the wheels is set so that every portion of the rod is inspected, with some overlap included. No trouble was experienced in maintaining the fuel rods in the proper position for testing as they passed under the probe. Probe-to-metal spacing was easily held within the range over which nearly perfect compensation could be achieved. The probe inspected an area roughly equal to its cross-sectional area, since cladding thickness

changes which lie beyond the projection of the probe on the fuel rod influenced the measurements, while the probe was relatively insensitive to the thickness of the cladding which lay immediately beneath its center. Inspection of a fuel rod 19.8 inches long required about 30 seconds.

Figure 35. The Complete System.

G. Calibration

The calibration of a nondestructive testing instrument is always one of the most difficult phases of its development. Calibration of this instrument involved the following steps. (1) A certain amount of the cladding was machined from a fuel rod. The amount machined off was determined by a measurement of the cladding thickness with a measuring microscope before and after machining. (2) The sensitivity of the instrument was adjusted so that it measured the incremental change correctly, going from the machined part of the fuel rod to the unmachined part.

Two sources of possible uncertainty arise at this point. The machining creates a standard of a slightly different diameter than the actual rod. However, this amounts to only about 0.8% change in diameter, much too small to cause any significant error since the instrument compensates for small changes in diameter. The other source of uncertainty arises from the fact that the eddy current measurements cannot be made at the end of the rod because of end effects, while the optical measurements can be made only at the end. So it is necessary to assume that the cladding thickness at a small distance from the end is practically the same as it is at the end. (3) Once the sensitivity of the instrument was adjusted to the proper value, the instrument was used to find a fuel rod which happened to possess very uniform cladding. The cladding on the rod selected varied less than 0.5 mil over the entire length. This rod was sectioned and found

to have cladding 21 mils thick. It was used as the standard for all subsequent measurements. It was necessary to apply a -1-mil correction to the measurements made on rods tested after heat treatment since the heat treatment caused changes to take place in the core material and bond layer which did not exist in the "as-extruded" standard.

Figure 36 shows an actual recording of the cladding thickness of a fuel rod. This rod was cut from near the end of the extrusion. The heavy taper of the cladding made this rod particularly suitable for sectioning. Exact comparison between nondestructive measurements and destructive measurements is impossible since the probe measures the average cladding thickness over a certain area while measurements made by the measuring microscope are obtained point-by-point. The oscillations of the trace in Figure 36 were caused by a slight eccentricity of the cladding. The dots shown on the trace indicate the average cladding thickness obtained when measurements were made on the cross section of the rod which lay immediately underneath the center line of the probe at that time. The average cladding thickness was based on 4 measurements around the circumference. The probable difference between nondestructive and destructive measurements based on this tape is 0.25 mil.

Figure 36. A Pen Recording of the Cladding Thickness of a Rod.

H. Acknowledgement

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The authors also wish to acknowledge the work of Mr. Nelson Beck in designing the conveyor used here.

VII. AN EDDY CURRENT TESTING INSTRUMENT FOR THE DETERMINATION OF WIRE QUALITY WITH AN APPLICATION TO THE SPACING WIRES USED ON THE EBR-I MARK III FUEL ELEMENTS

A. Introduction

The design of the EBR-I Mark-III reactor included zirconium wire spacers for the fuel elements. These wire spacers were welded in place on the fuel elements. The welding process required a wire which was a good heat conductor; therefore, the wires had to be fairly free of cracks and voids. After each wire was welded in place, part of it was removed by a machining process. If cracks existed in the wire, the machining process often split the wire along the crack and damaged the fuel element.

There were several possible tests which could have been made to determine the quality of the wire. First, a visual examination was used to detect the defects. This system was slow, and the nature of the external defect did not give a description of internal defects. Ultrasonics has been widely used for detecting internal defects, but in this particular application the small diameter, 0.055 inch, of the wire prohibited the use of this test. Eddy currents proved to be an effective test for the cracks and

voids. A block diagram of the eddy current instrument is shown in Figure 37. The purpose of this paper is to describe the eddy current instrument and its application to this particular problem. The instrument used is a modification of the instrument discussed in Chapter III, except that only the amplitude variations of the bridge output were used.

Figure 37. Block Diagram

B. Theory

A small test coil establishes a magnetic field in the wire to be tested. This magnetic field induces eddy currents in the wire. The magnitude and direction of the eddy currents depend upon the test-coil geometry, the coil excitation voltage amplitude and frequency, the distance from the coil to the wire, and the electrical properties and quality of the wire. The impedance of the coil changes with a variation of the magnitude and direction of the eddy currents. Measurements of the change in impedance of the coil⁽¹²⁾ can provide information concerning any of the above parameters. In this application, the wire quality was the only variable. Determination of the other parameters will be discussed later in this chapter.

Test coil sensitivity is the magnitude of the impedance change for a standard defect. This can also be defined in terms of the bridge output voltage. Coil definition is inversely related to the volume covered by the eddy currents. Maximum sensitivity and definition are desired for any test situation. Coil geometry largely determines the sensitivity and definition; however, the maximum for each variable does not occur for one particular coil geometry. A compromise must be made in accordance with the test situation. The optimum coil geometry is determined experimentally.

There are two types of test coils. The first type is called a point probe, because the axis of the coil is perpendicular to the surface of the conductor to be tested. In the ideal situation the source of the magnetic field would be a point, but in the actual situation a definite area is covered by the coil. This type of coil is used when the conductor to be tested has a flat surface or a diameter which is larger than coil diameter. The second type of test coil is called an encircling coil because the coil encircles the conductor to be tested. The eddy currents cover the volume of the conductor which is inside the coil. This type of coil is used when the conductor to be tested has a small diameter or a small cross-sectional area. Each of these two types have advantages over the other type in application to particular problems. The proper type of coil is determined experimentally. An encircling coil was used for testing the zirconium wire.

C. Operation

The change in the impedance of the test coil was measured by the modified Owens bridge circuit shown in Figure 38. The circuit was composed of the test coil, two variable resistances, and two fixed capacitances. The two variable resistances were used for the final balancing of the bridge. Determination of the capacitance values will be discussed later. The bridge was balanced with the coil in air. When a wire was brought near the coil, an unbalanced voltage appeared at the bridge output terminals. This voltage contained information about the defects in the wire. The crystal-controlled oscillator shown in Figure 38 supplied a 20-volt peak-to-peak sinusoidal voltage to the primary winding of the LS-30 transformer. The variation in this voltage was ± 15 millivolts. This drift could have led to a considerable error in the results, so the amplitude stabilizing circuit shown in Figure 38 was used to eliminate practically all the error. In the stabilizing circuit, the voltage at the transformer input was rectified, and a portion of this rectified voltage was applied as a bias voltage on the grid of the first stage of the amplifier. When the correct amount of bias was added, the oscillator voltage could be varied over a considerable range without introducing a change in the amplifier output voltage.

The unbalanced voltage from the bridge circuit was amplified with the five-stage amplifier shown in Figure 38. A tuned circuit from the grid of the second stage to the ground was employed to eliminate the undesirable harmonics. The last four stages were coupled by several degenerative

Figure 38. Circuit Diagram

feedback loops to improve the stability. The amplifier output voltage was rectified and the D.C. output voltage was adjustable with the 500-kilohm variable resistance. The differential D.C. voltmeter in Figure 38 indicated the output voltage. The 50 and 5-volt scales were used for balancing the bridge and adjusting the output voltage for a given unbalanced condition. The test (T) scale was used for indicating the wire quality. When the meter was on the test scale, the variable resistance in the grid circuit of the 12AU7 balanced out a portion of the D.C. voltage. This allowed the remaining D.C. voltage to cover the entire meter range; thus the meter deflection was greatly increased for a given defect in the wire. A photograph of the complete instrument is shown in Figure 39.

Figure 39. Photograph of the Completed Instrument

D. Application

The procedure followed in testing the zirconium wire for cracks and voids is described in this section. This is a typical application, and the procedure outlined here can be followed in other applications for which this instrument can be utilized. The most important factor in determining the coil type was the diameter, 0.055 inch, of the zirconium wire. The point probe could not be made small enough for good sensitivity; however, the encircling coil proved to be very effective for detecting the cracks and voids. The cross section of this coil is shown in Figure 40 and

Figure 40. Test Coil Cross Section

a photograph of the coil is shown in Figure 39. Several different diameters, lengths, and number of turns were tried for the coil to obtain optimum sensitivity and definition. The diameter of the coil that was used was about 0.015 inch greater than the wire diameter. The coil was close to the wire, and this greatly improved the sensitivity. The length of the coil was 0.188 inch. Since the length affected both the sensitivity and definition, this length was a good compromise for this application. The coil had two layers of number 40 wire wound on a nylon coil form.

Variations in probe coil-to-wire spacing changed the impedance of the coil, and error was introduced. To minimize this error, the diameter of the center hole of the probe was made 0.005 inch greater than the nominal wire diameter. Essentially no error was introduced for a variation of this magnitude. A change in wire diameter had about the same effect as a change in spacing. The diameter of this wire varied ± 0.003 inch, and again no appreciable error was introduced. Changes in the operating temperature of the coil varied its impedance. If these changes were too drastic, the calibration of the instrument was affected. These changes in temperature came from changes in the ambient temperature or from excess coil current. Eddy currents of a considerable range of frequency were satisfactory for detecting flaws in the wire. This range can be determined theoretically⁽¹⁹⁾ or experimentally, and it varies with the particular permeability, conductivity and diameter of the wire to be tested.

The bridge output voltage amplitude for a wire without defects was largely determined by the bridge parameters. A maximum output voltage was desired. This maximum output voltage was attained by varying the ratio of C_1 to C_2 . The capacitances C_1 and C_2 were determined experimentally for each coil. The correct values for the coil in this application are shown in Figure 38. Since the magnitude of the impedance change of the coil depended upon the electrical properties of the wire, the output voltage of the instrument also depended upon these properties. The electrical properties of the zirconium were fairly constant and no complications were introduced.

The defects in the wire appeared as cracks or voids along the length of the wire. The size of the crack or void determined the magnitude of the unbalanced voltage at the bridge terminals. The magnitude decreased as the size of the defect increased. Several typical cross sections of wire and corresponding meter readings are shown in Figure 41. A cross section of the standard wire is shown in Figure 42. The first two cross sections in Figure 41 were not etched, and this accounts for the smooth appearances of their surfaces.

To calibrate the meter test (T) scale, several wires were checked, and a complete range of voltage readings was obtained from the rectifier output. The wires were then sectioned at the points where the readings were taken, and the wire qualities were determined. This gave two ranges of output voltage: 0.0 to 4.25 volts for the unacceptable wire, and 4.25 to 5.0 volts for the acceptable wire. To expand the range of acceptable wire, the test scale was calibrated for a full range of 2.5 volts. The value of the resistance (R_T) for this calibration was 0.2 megohm. The balancing adjustment in the meter circuit balanced out the undesirable 2.5 volts.

The procedure used to set up the instrument for operation contained these steps: The bridge was balanced with no wire in the test coil. With the meter on the 5-volt scale, the standard wire, which was without defects, was placed in the coil. The gain was adjusted until the meter read 5 volts. The meter was switched to the test scale, and the balance was adjusted until a full-scale deflection of the meter was obtained. The instrument was then ready to check wire.

The velocity of the wire through the probe was limited by the inertia of the meter needle. When the velocity was too great, the needle would not follow the output voltage and error was introduced. The velocity was approximately 1.5 inches per second in this application. This velocity can be greatly improved if the meter is replaced by an electronic relay circuit.

E. Conclusion

Several hundred feet of wire were checked, and approximately 33 per cent was found to be acceptable. The final test of the wire was the spot welding and machining process. The accepted wire passed this test. This instrument can be depended upon to give a measure of the integrated wire quality. In other words, several small cracks may give a meter indication that is the same as one large crack. In the application described in this paper, wire was rejected for small voids and porosity as well as for cracks. In production-line applications similar to the one described in this paper, the instrument can determine the quality of wire quickly and with good reliability.

STANDARD WIRE	METER READING
TEST WIRE	5.0
	0.0

STANDARD WIRE	METER READING
TEST WIRE	5.0
	0.5

STANDARD WIRE	METER READING
TEST WIRE	5.0
	2.0

STANDARD WIRE	METER READING
TEST WIRE	5.0
	2.5

Figure 41. Wire Cross Sections and Corresponding Meter Readings

	METER READING
STANDARD WIRE	5.0

Figure 42. Standard Wire Cross Section

VIII. AN APPLICATION OF THE MULTI-FREQUENCY EDDY CURRENT SYSTEM TO THE MEASUREMENT OF THE THICKNESS OF ZIRCONIUM CLADDING ON A URANIUM-MOLYBDENUM ALLOY

A. Introduction

Chapter VI described an instrument which can be used in a variety of applications. The theory of operation and one application of the instrument were discussed in that chapter. Another application of a similar instrument will now be described. In this application, the thickness of zirconium cladding on uranium-molybdenum fuel pins is measured. The nominal cladding thickness was 0.004 inch and the nominal outside diameter of the pins was 0.158 inch. The resistivity of the cladding and core were $69.3 \mu\text{ohm-cm}$ and $39.3 \mu\text{ohm-cm}$, respectively.

B. Operation

Circuit diagrams for this instrument are given in Figures 43 and 44. The depth of penetration of eddy currents for a frequency of 4.5 Mcps was limited to a small percentage of the cladding thickness. Figure 43 shows the circuit diagram of this high-frequency channel. This channel gave probe-to-metal spacing information with no interference from changes in cladding thickness.

The low-frequency channel, shown in Figure 44, was operated at a frequency of 256 kcps. Eddy currents for this frequency penetrated the fuel pins to a depth greater than the cladding thickness. This channel gave both the cladding thickness and probe-to-metal spacing information. The high-frequency channel output voltage was applied to the last stage of the low-frequency channel amplifier. A change in the high-frequency channel output voltage changed the gain of the low-frequency channel amplifier, and the effect of a change in probe-to-metal spacing was compensated out of the low-frequency channel output voltage. This output voltage was independent of probe-to-metal spacing over a range of about 0.010 inch. This range could probably be made larger by carefully designing the low-frequency channel amplifier. The design procedure is given in Appendix IV. The cladding was read directly from the differential D.C. voltmeter. The procedure for making the initial balance and adjustments of the instrument was described in Chapter VI.

C. Probe Design

The probe used in this application was a small ferrite-cored, single-layer coil of number 40 wire. The diameter of the ferrite was 0.0625 inch. To increase the impedance of the low-frequency bridge to the high-frequency bridge, the probe coil was center tapped so that one-half the coil was in series with the low-frequency bridge. This made balancing of the high-frequency bridge easier. A cross section of the probe is shown in Figure 45.

Figure 43. Eddy Current Tester ANL Model C-3 High Frequency Channel

Figure 44. Eddy Current Tester ANL Model C-3 Low Frequency Channel

Figure 45

Cross Section of the Probe
Showing the Method of
Center Tapping the Coil.

D. Conclusion

This instrument measured the 0.004 inch of cladding within ± 0.0002 inch. The area of a measurement was about 0.003 square inch. The instrument is suitable for fast production-line applications.

IX. APPENDIX

A. APPENDIX IAN ANALYSIS FOR THE IMPEDANCE OF A COIL NEAR A
CONDUCTING SURFACE

For a single coil of N turns isolated from all other objects, the self-inductance L_0 is given by⁽⁶⁾

$$L_0 = \mu N^2 (2a - r) \left[\left(1 - \frac{1}{2k^2}\right) K(k) - E(k) \right] \quad , \quad (1)$$

where

a = radius of the coil

r = radius of the wire

μ = permeability of the space surrounding the coil

$$k^2 = \frac{4a(a - r)}{(2a - r)^2}$$

$K(k)$ = complete elliptic integral of the first kind

$E(k)$ = complete elliptic integral of the second kind.

When a coil is placed near a perfect conductor, as shown in Figure 1a, an image coil coaxial with the first coil may be thought of as existing below the surface of the metal.^(7,8) The inductance of the first coil is then

$$L_c = L_0 - \frac{2\mu N^2 \sqrt{a(a - r)}}{k_1} \left[\left(1 - \frac{1}{2k_1^2}\right) K(k_1) - E(k_1) \right] \quad (2)$$

where

$$k_1 = \sqrt{\frac{4a(a - r)}{(2a - r)^2 + (2d)^2}}$$

d = distance between the coil and the surface of the metal.

The magnetic field intensity on the surface of the metal is

$$H_\rho = \frac{NI d}{\pi} \left\{ \frac{\left[\frac{a^2 + p^2 + d^2}{(a - p)^2 + d^2} \right] E(k) - K(k)}{\rho [(a + p)^2 + d^2]^{1/2}} \right\} \quad (3)$$

where

$$k = \sqrt{\frac{4ap}{(a + p)^2 + d^2}}$$

I = current in the probe coil.

Figure 1a

The Geometry of the Coil
Placed Near the Metal.

ρ and ϕ are polar coordinates on surface of the metal with the origin at the point where the axis of the probe coil meets the surface of the metal. The current density (J_ϕ) on the surface of the metal has the same numerical magnitude as the magnetic field intensity H_ρ of (2). The power dissipated in the whole plane is

$$P_d = \frac{1}{2} \int_0^\infty R_s (2\pi p) dp |J_\phi|^2$$

$$= \frac{R_s N^2 I^2 d^2}{\pi} \int_0^\infty \frac{\left\{ \frac{a^2 + p^2 + d^2}{(a - p)^2 + d^2} E(k) - K(k) \right\}^2}{p[(a + p)^2 + d^2]} dp \quad (4)$$

where R_s is the skin resistance of the metal, (6)

The reflected resistance into the coil is

$$R_r = \frac{P_d}{(I/\sqrt{2})^2} = \frac{2R_s N^2 d^2}{\pi} \int_0^\infty \frac{\left\{ \frac{a^2 + p^2 + d^2}{(a - p)^2 + d^2} E(k) - K(k) \right\}^2}{p[(a + p)^2 + d^2]} dp \quad (5)$$

If now in place of R_s , the skin effect internal impedance,

$$Z_s = R_s + jL_i\omega,$$

is used, then the impedance (Z) reflected into the coil is

$$Z = \frac{2Z_s N^2 d^2}{\pi} \int_0^\infty \frac{\left\{ \frac{a^2 + d^2 + p^2}{(a - p)^2 + d^2} E(k) - K(k) \right\}^2}{p[(a + p)^2 + d^2]} dp \quad (6)$$

To evaluate the integral of (6) put $x = (\rho/a)$ and $d = (b/a)$; (10) becomes

$$Z = \frac{2Z_s N^2 b^2}{\pi} \int_0^\infty \frac{\left\{ \frac{1 + x^2 + b^2}{(1 - x)^2 + b^2} E(k) - K(k) \right\}^2}{x[(1 + x)^2 + b^2]} dx \quad (7)$$

with

$$k = \sqrt{\frac{4x}{(1 + x)^2 + b^2}} \quad .$$

If now in (7) the integration is done with respect to k instead of x , (7) becomes

$$f(y) = \frac{\pi Z}{2Z_s N^2} = \frac{2y}{1 + y} \int_0^q \frac{\left\{ \left(\frac{1 - k^2/2}{1 - k^2} \right) E(k) - K(k) \right\}^2}{k \sqrt{\frac{1 - k^4(1 + y)}{(2 - k^2)}}} dk \quad (8)$$

where

$$y = b^2$$

$$q = \frac{2}{y} \sqrt{(-1 + \sqrt{1 + y})}$$

The integral $f(y)$ was evaluated on the electronic digital computer AVIDAC at the Argonne National Laboratory with the results given in Table I.

Table I

THE FUNCTION $f(y)$ AS EVALUATED
BY THE COMPUTER

y	f(y)	y	f(y)
00.01	15.03865	00.80	00.51158
00.05	05.96961	01.00	00.38193
00.10	03.75141	02.00	00.13857
00.15	02.76572	03.00	00.07164
00.20	02.18480	04.00	00.04381
00.30	01.51567	06.00	00.02131
00.40	01.13766	08.00	00.01254
00.60	00.72687	10.00	00.00831

Now let

$$Z = R + jL_m \quad (9)$$

where R is the change in resistance of the probe caused by the proximity of the metal and L_m is the change in inductance caused by the fact that the metal is not a perfect conductor. Then the change (L) in inductance of the probe coil caused by the presence of the metal is

$$L = L_0 - (L_c + L_m) = (L_0 - L_c) + L_m \quad (10)$$

where $(L_0 - L_c)$ may be obtained from (4) and L_m from (7) and (8).

In measuring R and L experimentally, the inductance (L_0) and the resistance (R_0) of the probe coil were measured with the metal absent. The metal was then brought close to the coil and the inductance ($L_0 + L$) and the resistance ($R_0 + R$) measured. The difference of the measured quantities gave the change (R) in resistance and the change (L) in inductance which could be compared with the calculated values of R and L as obtained from (8) and (9).

It should be pointed out here that the analysis assumes that the frequency is low enough and the probes small enough so that the effect of radiation from the coil may be completely neglected. Also the change in inductance (L_0) with frequency is assumed negligible.

B. APPENDIX II

BRIDGE ANALYSIS

The current (i) flowing through Z_5 in a bridge such as the one in Figure 2a is given by⁽¹³⁾

$$i = \frac{Z_1 Z_3 - Z_2 Z_4}{\Delta} \cdot e \quad (11)$$

where

$$\Delta = \begin{vmatrix} -Z_3 & -(Z_3 + Z_4) & Z_3 + Z_4 + Z_6 \\ Z_2 + Z_3 + Z_5 & Z_2 + Z_3 & -Z_3 \\ -Z_5 & Z_1 + Z_4 & -Z_4 \end{vmatrix} \quad (12)$$

$$\frac{e}{i} = \frac{\Delta}{Z_1 Z_3 - Z_2 Z_4}$$

Figure 2a
The Bridge Circuit.

For the Owen bridge,

$$Z_1 = R + R_2 + j\omega L_0$$

where

$$R + j\omega L_0 = \text{probe impedance}$$

$$Z_2 = R_3 - \frac{j}{\omega C_1}$$

$$Z_3 = \frac{-j}{\omega C_2}$$

$$Z_4 = R_1$$

ω = angular frequency of source.

When the above substitutions are made in equation 12, the result is:

$$\frac{e}{i} = \frac{\Delta}{(R + j\omega L_0 + R_2) \left(\frac{-j}{\omega C_2} \right) - \left(R_3 - \frac{j}{\omega C_1} \right) R_1} \quad (13)$$

For balance,

$$R = \frac{R_1 C_2 - R_2}{C_1}$$

$$L_0 = C_2 R_3 R_1$$

or

$$R_3 = \frac{L_0}{R_1 C_2} \quad .$$

Suppose balance has been attained and a metal sample is brought near the probe. R_2 was zero.

$$\frac{e}{i} = \frac{\Delta}{R + \delta R + j\omega (L_0 - \delta L) \left(\frac{-j}{\omega C_2} \right) - \left(\frac{L_0}{C_2 R_1} - \frac{j}{\omega C_1} \right) R_1} \quad (14)$$

δR and δL are always positive.

Δ stays nearly constant for small changes in R and L .

Then

$$\frac{e}{i} = \frac{\Delta}{- \left(\frac{\delta L}{C_2} + \frac{j \delta R}{C_2} \right)} \quad (15)$$

$$= \frac{\omega C_2}{\omega \delta L + \delta R} \quad (16)$$

Since

$$\delta L = L_0 - L \quad ,$$

where L_0 is the inductance of the probe in air
and L is the inductance of the probe with the sample in proximity, then

$$\frac{e}{i} = \frac{\frac{-C_2}{L_0} \Delta}{\left(1 - \frac{L}{L_0} \right) + \frac{j \delta R}{\omega L_0}} = \frac{-C_2 \Delta}{(L_0 - L) + j \delta R} \quad (17)$$

C. APPENDIX III

BRIDGE CIRCUIT ANALYSIS

The equivalent circuit of the bridge circuit and bridge circuit pulser is shown in Figure 3a. The voltage e was calculated by replacing the transformer and the bridge with an equivalent circuit which was found by measuring the input impedance at the transformer primary. Then the equivalent circuit is as shown in Figure 4a. The expression for the Laplace transform of the voltage e is:

$$L(e) = \frac{\frac{1}{R_1 C'} E_0 S}{S^3 + \left(\frac{1}{R_1 C} + \frac{1}{R_1' C'} + \frac{1}{R_1 C'} \right) S^2 + \left(\frac{1}{R_1 R_1' C C'} + \frac{1}{C' L'} \right) S + \frac{1}{R_1 C C' L'}} \quad (18)$$

Figure 3a. Equivalent Circuit of Bridge Pulser and Bridge.

Substituting circuit values

$$L(e) = \frac{3.46 \times 10^6 E_0 S}{S^3 + 9.15 \times 10^6 S^2 + 12.76 \times 10^{12} S + 9.35 \times 10^{18}} \quad (19)$$

The Laplace transform of the primary current is:

$$L(i_p) = \frac{1}{L' S} L(e) \quad (20)$$

After substituting Eq. (19) into Eq. (20) and taking the inverse Laplace transform, the primary current (i_p) is:

$$i_p = 1.67 \times 10^{-3} E_0 e^{-0.755 \times 10^6 t} \sin(0.806 \times 10^6 t - 6.8^\circ) \quad (21)$$

for $t \geq 0.3$ microsecond. The Laplace transform for the approximate

primary current (i_p) is:

$$L(i_p) = \frac{-0.197 \times 10^{-3} E_0 S + 1.19 \times 10^3 E_0}{[(S + 0.755 \times 10^6)^2 + (0.806 \times 10^6)^2]} \quad (22)$$

Figure 4a. Equivalent Circuit for Calculating the Voltage Across the Transformer Primary.

The resistances R_1 and R_2 were assumed small enough to be neglected. The 10-kilohm resistances may be neglected, because they are large compared to the 1-kilohm resistances. The large input resistance R_d of the mixer was neglected.

With the bridge balanced, the voltage equation for the secondary is:

$$0 = M \frac{di_p}{dt} + L_s \frac{di_z}{dt} + i_z Z \quad (23)$$

where Z is the entire bridge impedance and

$$Z(S) = \frac{2LRS}{RCLS^2 + LS + R} \quad (24)$$

The Laplace transform of Eq. (23) was solved for the Laplace transform of secondary current. Multiplication by $Z(S)$ yields

$$L(e_2) = \frac{-MS L(i_p) Z(S)}{L_s S + Z(S)} \quad (25)$$

The circuit values were substituted in Eq. (22) and the inverse Laplace transform was used to obtain:

$$\begin{aligned} \frac{e_2}{k10^{-9}} = & 39.8 e^{-0.755 \times 10^6 t} \sin(0.806 \times 10^6 t + 46.6^\circ) \\ & + 88 - 8.45 \times 10^6 t \\ & - 1630 - 31.6 \times 10^6 t \end{aligned} \quad (26)$$

where

$$K = 0.394 \times 10^{-3} \frac{ME_0}{L_s C} .$$

This equation is approximated by

$$\frac{e_2}{K10^{-9}} = 45e^{-0.9 \times 10^6 t} \quad (27)$$

for $0.5 \text{ microsecond} \leq t \leq 2.0 \text{ microseconds}$. The Laplace transform of Eq. (27) is:

$$\frac{L(e_2)}{K10^{-9}} = \frac{45}{S + 0.9 \times 10^6} . \quad (28)$$

The bridge circuit was drawn as indicated in Figure 5a. Let the voltage e_s be any applied voltage, and e_0 the output voltage. The mixer input resistance was assumed to be 10^6 ohms. The loop equations were:

$$-e_s = 20 \times 10^3 i_1 - 10 \times 10^3 i_2 \quad (29)$$

$$0 = 10^4 i_1 + (10^4 + 10^6 + Z_2) i_2 - Z_2 i_3 \quad (30)$$

$$e_s = Z_2 i_2 + (Z_1 + Z_2) i_3 . \quad (31)$$

Figure 5a. Equivalent Circuit of the Bridge.

These equations were solved for the current i_2 , and i_2 was multiplied by the mixer input resistance to obtain

$$e_0 = \frac{10^6(Z_2 - Z_1) e_s}{2Z_1 Z_2 + 2.01 \times 10^6 (Z_1 + Z_2)} \quad (32)$$

The Laplace transform of this equation is:

$$L(e_0) = \frac{10^6 [Z_2(S) - Z_1(S)] L(e_s)}{2Z_1(S) Z_2(S) + 2.01 \times 10^6 [Z_1(S) + Z_2(S)]} \quad (33)$$

where

$$Z_i(S) = \frac{\frac{1}{C} S + \frac{R_i}{L_i}}{S^2 + \left(\frac{R_i}{L_i} + \frac{1}{RC} \right) S + \frac{R_i + R}{CL_i}} \quad (34)$$

The probe characteristics are assumed to be:

$$L_0 = 200 \text{ microhenries}$$

$$R_0 = 50 \text{ ohms}$$

in air. From previous sinusoidal investigations,⁽¹²⁾ the probe inductance and resistance varied with a change in probe-to-metal spacing according to:

$$L = L_0 - \Delta L$$

$$R = R_0 + \Delta R$$

where

$$R = K\Delta L$$

and K is a constant.

Let L_1 and R_1 equal L_0 and R_0 , respectively. Let it be assumed that a metal is close to probe two, and that the probe characteristics for three different probe-to-metal spacings are:

d	$L_2 (\mu h)$	$R_2 (\Omega)$
0	120	80
X'	130	75
X	140	70

These values were substituted in Eqs. (33) and (34). Let Eq. (28) be the applied voltage function. This was substituted in Eq. (33) in place of the Laplace transform of any voltage function e_s . The inverse Laplace transform was taken and for $d = 0$

$$\frac{e_0}{K'} = -0.105\epsilon^{-0.403 \times 10^6 t} + 0.236\epsilon^{-0.9 \times 10^6 t} , \quad (35)$$

where

$$K' = \frac{-45}{K10^{-9}} ;$$

for $d = X'$

$$\frac{e_0}{K'} = -0.081\epsilon^{-0.378 \times 10^6 t} + 0.192\epsilon^{-1.9 \times 10^6 t} ; \quad (36)$$

for $d = X$

$$\frac{e_0}{K'} = -0.059\epsilon^{-0.352 \times 10^6 t} + 0.151\epsilon^{-0.9 \times 10^6 t} . \quad (37)$$

These equations are plotted in Figure 20, p. 27 .

D. APPENDIX IV

A MULTI-FREQUENCY EDDY CURRENT TESTING SYSTEM

A procedure for achieving satisfactory probe-to-metal spacing compensation over a desired range of spacings will be illustrated by an example. A normal probe-to-metal spacing of 20 mils was chosen. Curves of high- and low-frequency channel output voltages are shown in Figures 27 and 28, respectively. Figures 6a and 7a are the same as Figures 27 and 28, but with the axis shifted. It was desired to compensate for distances between 10 and 30 mils from the metal, or between -10 and +10 mils on the shifted axis shown in Figures 6a and 7a. For simplicity, and since it will serve just as well as an example, a 3-term power series will be used to approximate the equation of the low-frequency channel output voltage plotted as a function of distance:

$$y = Ax^2 + Bx + K \quad (38)$$

The value of K was immediately determined since $y = K$ at $x = 0$. The other two coefficients were obtained by using the value of y at $x = +10$ and $x = -10$ from Figure 7a and substituting it in equation (38). The resulting simultaneous linear equations were solved, yielding

$$y = 0.00055x^2 - 0.0615x + 4.44 \quad (39)$$

In order to maintain the low-frequency channel output invariant with respect to probe-to-metal spacing, the compensating 6L7 tubes and high-frequency channel together must generate a function $G(y')$ such that

$$y G(y') = K_1 \quad (40)$$

where K_1 is a constant. Defined in this way, $G(y')$ will also maintain a constant sensitivity as a few quick calculations from Figures 28 and 29 will indicate. If it is desired to maintain the low-frequency channel output the same at 10-mil spacing as at 20-mil spacing, the low-frequency channel amplifier gain must decrease by a factor of $4.4/5.1 = 0.861$. If a 3-mil change in cladding occurs while the spacing is 20 mils, the instrument will indicate the correct change. But without compensation and with a spacing of 10 mils, the instrument would indicate 3.6 mils instead of the actual 3-mil change. To obtain the correct result, the amplifier gain would need to be reduced by a factor of $3/3.5 = 0.857$. These factors are so nearly the same that no important change in sensitivity occurs over the range over which equation (40) must be satisfied. While we have no proof that $y G(y') = K_1$, a constant sensitivity is maintained in every case. As a practical matter we have never encountered a case in which this was not true, and the idea has intuitive appeal.

Figure 6a This figure is identical to Figure 27 except that both axes are shifted.

Figure 7a This figure is identical to Figure 28 except that the X axis is shifted.

Now

$$G(y') = \frac{K_1}{y} = \frac{K_1}{0.00055x^2 - 0.061x + 4.44} \quad (41)$$

We wish to obtain $G(y')$ in terms of y' , the high-frequency channel output voltage. Note that in Figure 6a both the abscissa and ordinate have been shifted so that the high-frequency channel output voltage is zero at $x = 0$, which is actually a probe-to-metal spacing of 20 mils. The justification for doing this will be obvious later. We approximate the equation of the probe-to-metal spacing as a function of high-frequency channel output voltage by a power series:

$$x = -1.102 y'^2 + 13.74 y' \quad (42)$$

Substituting equation (42) in equation (41),

$$G(y') = \frac{K_1}{0.00055 (-1.102y'^2 + 13.74y')^2 - 0.061 (-1.102y'^2 + 13.74y') + 4.44}$$

If the relative gain of the 6L7 stages is given the value 1 at $x = 0$, then

$$G(y') = \frac{4.44}{0.00055 (-1.102y'^2 + 13.74y')^2 - 0.061 (-1.102y'^2 + 13.74y') + 4.44}$$

A plot of $G(y')$ is shown in Figure 8a.

Figure 8a The Gain Function $G(y')$ Plotted as a Function of the High Frequency Channel Output Voltage y' .

The next step is to use the curve of low-frequency channel output voltage versus the #1 grid to ground potential of the 6L7 stages, given in Figure 9a. In addition, as can be seen from the circuit diagram, a provision has been made to set the Q point of the #1 grid of these stages. This was the reason we could ignore the actual voltage developed across the high-frequency channel output at 20 mils probe-to-metal spacing. Mathematically, this amounted to shifting the axis. Selecting a portion of the curve of Figure 9a which will match $G(y')$ as closely as is necessary,

Figure 9a The Low Frequency Channel Output Voltage Plotted as a Function of the Grid-to-Ground Bias on the 6L7 Stages.

Evidently, c has been chosen too low. With $c = 2$, the relative gain of the 6L7 stages when $y' = 0.7$ is 0.875, while the gain at $y' = +0.78$ is 1.13. This is quite close to the gain needed at each of the end points and is as far as it is necessary to carry the procedure in this example. Since $c = 2$, the gain of the high-frequency channel should be halved. A positive voltage would need to be inserted in series with the output of the high-frequency channel to set the grid to ground potential of the 6L7 stages to -1 volt at $x = 0$. This analysis is based on approximate expressions for curves of the high and low-frequency channel output distance versus spacing. These could be made more accurate by including terms of higher powers in the expressions. The curve of Figure 9a could be changed by many different means if good compensation could not be attained otherwise.

the process is largely one of cut and try. From Figures 6a and 8a it can be noted that $x = -10$, $y' = 0.7$ and $G(y') = 0.88$, while at $x = +10$, $y' = 0.78$ and $G(y') = 1.12$. From Figure 9a, one can start by deciding to operate the 6L7 stages about the -1-volt point and decreeing that the relative gain of these stages at that bias will be 1. In effect, we are erecting a new set of axes at the (4, -1) point in Figure 9a. These are shown as heavy lines. Note also that the abscissa of this new set of axis is labeled y'' , where $y'' = y'/c$ and c is a constant. We are free to vary the gain of the high-frequency channel and are not limited to the value of output shown in Figure 6a. As a start, we set $c = 1$ so that $y'/c = y'$. Then the relative gain of the 6L7 stages at $y'/c = -0.7$ is 0.778, which is too low, while at $y'/c = +0.78$, the relative gain at $y'/c = 1.32$, which is too high.

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