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Instruments

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A CAPACITANCE GAGE FOR
DIMENSIONAL MEASUREMENT

J. J. Henry

UNION CARBIDE CORPORATION
NUCLEAR DIVISION
OAK RIDGE Y-12 PLANT

operated for the **ATOMIC ENERGY COMMISSION** *under* **U. S. GOVERNMENT** **Contract W-7405 eng 26**

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ABSTRACT

A capacitance distance gage has been designed that has a variable frequency output suitable for high-accuracy, noncontact dimensional inspection with digital readout or direct computer processing. This gage further provides improved performance for medium to high-resistance materials and can be compensated to provide a low temperature drift. A gage stability of ± 15 microinches in the measuring range from zero to 60 mils has been realized using a 3/8-inch-diameter sensing probe tip, with a 5° F ambient temperature variation.

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SUMMARY

A high-precision capacitance-type noncontact gage, for dimensional inspection and product certification, has been designed at the Oak Ridge Y-12 Plant.^(a) This gage uses an RC integrating-oscillator measuring circuit which provides a simple linear reciprocal equation for calibration, allowing easy use with digital data processing (either by computer, or a hard-wired linearizing configuration). Two types of probe circuitry were developed. One form uses a unijunction transistor and the other uses a field-effect transistor (FET) and Schmitt trigger. The first circuit is lower in cost and simpler in circuit configuration, while the second is more stable and more uniform in its calibration and characteristics.

Calibration equations for this type of oscillating gage have been derived and compared with those of other variable oscillator gages. The FET probe showed conformity to within ± 15 microinches for a 0.100-inch range with less than a 100-microinch drift for the temperature range from 10 to 75° C.

(a) Operated for the US Atomic Energy Commission by the Union Carbide Corporation's Nuclear Division.

INTRODUCTION

Capacitance gages have been used for noncontact dimensional inspection for a number of years. Several analog circuits such as the capacitance bridge, the resonant slope detector, and the guarded impedance meter are used in these measuring circuits. The circuit is used to measure an extremely small capacitance between the probe tip and part to be inspected, with the output signal displayed on an analog meter or recorder. The basic means employed to measure these very low capacitances uses one of the following techniques: (1) the amplitude of an AC voltage or current is varied by the capacitor, converted to a DC signal, and displayed; or (2) an oscillator frequency is varied by the capacitor, this frequency is then converted to a DC signal, and displayed. All of these systems are basically analog instruments and, even in the existing oscillator types, it is necessary to convert from an analog to a digital signal for computer processing. Most of the existing oscillator types are active only for a limited range in the operating region, and they show gross nonconformity near contact with the part. This problem makes accurate absolute dimension measurement extremely difficult due to the difficulty in obtaining an absolute zero. With the advent of the digital computer for data processing, it is advantageous to use a digital readout, such as a variable frequency, directly into the computer in order to avoid the errors introduced in converting first to a DC signal and then from analog to digital. As previously mentioned, however, existing variable-frequency capacitance gages lack the accuracy, stability, and conformity necessary to advantageously use digital data processing.

An ideal circuit for obtaining the desired variable frequency could be obtained using the RC-integrator pulse-type oscillator (such as a unijunction oscillator) if a stable, extremely low-capacitance (high-impedance) circuit could be devised. This circuit would extrapolate to zero frequency at contact which, even if not measurable (due to the capacitor shorting out), would be extrapolatable due to the expected conformity to the theoretical equations. Hence, the main problem involved in designing this circuitry is one of circuit configuration and component or circuit selection for optimum stability, conformity, and precision. Since this circuitry demands a high input impedance, it also has the further advantage of better accuracy when medium to high-resistance parts are being inspected.

This report is concerned primarily with the design of RC-integrator-type gage probes. Experience has been acquired with both a unijunction circuit and an integrator/Schmitt trigger circuit, and with the design and application of gage probes.

THE CAPACITANCE GAGE

GAGE DESIGN

Variable Oscillator Capacitance Gages

There are several ways in which a variable capacitor can change the frequency of an oscillator. The capacity of a parallel plate placed a short distance, d , above a conducting surface is given by the formula:

$$C_p = \frac{KA}{d}, \quad (1)$$

where:

C_p represents the capacitance,

K the dielectric constant,

A the area of the plate, and

d the separation distance.

It can be readily seen from this formula that the capacitance varies from infinity (when $d = 0$) to an extremely small value (when $d \gg A$) and approaches zero when d approaches infinity.

In the usual capacitance gage, the conducting surface is grounded and a grounded guard shield is used around the probe to reduce the effects of external object motion. This arrangement introduces a fixed stray capacity. The electronic circuitry will also have some input capacitance and there will be some necessary input wiring capacity. If all these unavoidable input circuit capacitances are represented as a fixed single capacity, C_s , then:

$$C = C_p + C_s, \text{ and} \quad (2)$$

$$C = \frac{KA + dC_s}{d}, \quad (3)$$

where C represents the total input capacity.

If the capacitor is then used to vary the frequency of an oscillator, the effects of varying " d " can be predicted from the oscillator tuning equation and the known oscillator characteristics.

When the capacitor is used as one of the elements in a Wein-bridge RC oscillator (eg, C_2 in Figure 1), the tuning equation requires zero phase shift and nearly equal amplitudes at the bridge midpoints (the amplifier inputs). The zero phase shift requirement demands that the phase angle through the two halves of the tuned side of the bridge must be equal and opposite. This requirement is met only when:

$$\frac{X_{C_1}}{R_1} = \frac{R_2}{X_{C_2}}, \quad (4)$$

but X_C is equal to $1/2\pi FC$. Hence, substituting and rearranging terms gives:

$$F = \frac{1}{2\pi \sqrt{R_1 C_1 R_2 (C_p + C_s)}} \quad (5)$$

If $R_1 R_2 C_1$ is equal to the constant, W , then:

$$F = \frac{\sqrt{d}}{2\pi \sqrt{W C_s d + W K A}}, \quad \text{and} \quad (6)$$

if $a = 2\pi \sqrt{W C_s}$ and $b = 2\pi \sqrt{W K A}$, then:

$$F = \sqrt{\frac{d}{ad + b}} \quad (7)$$

As Equation 7 shows, the theoretical tuning equation is the square root of a reciprocal equation which should be weighted toward smaller values of d and becomes relatively insensitive as d approaches A in value.

Figure 1. WEIN-BRIDGE OSCILLATOR.

The Wein-bridge circuit is further complicated due to its gain balance requirements. The limited range of the thermistor gain adjusting circuit (T and R_3 of Figure 1) demands that, for oscillation, the value of R_2C_2 be somewhere near the value of R_1C_1 . This restriction prohibits operation near $d = \text{infinity}$ unless a large capacitor is added to C_s and the oscillator will stop oscillating when d approaches zero.

A similar analysis of an LC oscillator also shows that:

$$F = \frac{1}{2\pi\sqrt{L(C_p + C_s)}} \quad , \text{ or} \quad (8)$$

$$F = \frac{1}{\sqrt{\frac{4\pi^2 LKA}{d} + \frac{4\pi^2 LC_s d}{d}}} \quad , \text{ or} \quad (9)$$

$$F = \frac{\sqrt{d}}{\sqrt{4\pi^2 L(KA + C_s d)}} \quad , \quad (10)$$

which has similar characteristics to the Wein bridge.

If, however, the capacitor is integrated toward some fixed voltage, V_h , through a resistor, R , until it reaches an upper detection voltage, V_p , then is discharged toward another fixed voltage, V_L , through another resistor, R_2 , until it reaches a lower detection voltage, V_q (as represented in Figure 2), whereupon the process is repeated, the output frequency of this oscillator is the inverse of one period.

Figure 2. TIMING WAVEFORMS FOR AN R-C INTEGRATOR OSCILLATOR.

From the charging equation:

$$V_p - V_q = (V_h - V_q) (1 - e^{-T_1/R_1C}) , \text{ or} \quad (11)$$

$$\frac{V_h - V_p}{V_h - V_q} = e^{-T_1/R_1C} . \quad (12)$$

Since V_h , V_p , and V_q are constants of calibration, they may be adjusted as desired to calibrate T_1 as a function of C :

$$-\frac{T_1}{R_1C} = \text{Log}_e \frac{V_h - V_p}{V_h - V_q} , \text{ or} \quad (13)$$

$$\frac{T_1}{R_1C} = M \text{ (a constant) when } M = -\text{Log}_e \frac{V_h - V_p}{V_h - V_q} . \quad (14)$$

From the discharging equation:

$$V_p - V_q = (V_p - V_L) (1 - e^{-T_2/R_2C}) , \text{ or} \quad (15)$$

$$\frac{T_2}{R_2C} = N \text{ (a constant)} , \quad (16)$$

where:

$$N = -\text{Log}_e \frac{V_p - V_q}{V_p - V_L} .$$

Hence, $T_1 = MR_1C$, $T_2 = NR_2C$, and $T = T_1 + T_2 = (MR_1 + NR_2)C$, or $T = HC$, where H is a constant determined by fixed components and calibration voltages.

$$F = \frac{1}{HC} = \frac{1}{\frac{HKA}{d} + \frac{dHC_s}{d}} , \text{ or} \quad (17)$$

$$F = \frac{d}{a + bd} , \quad (18)$$

where $a = HKA$ and $b = HC_s$.

Equation 18 is then the general calibration equation for either a unijunction oscillator or for an integrator/Schmitt trigger oscillator. As can be seen from the equation, this is a simple reciprocal equation without the complexity of the square root function and hence should have better resolution as d approaches the probe area in size. In addition, the oscillator may operate down to zero frequency and, if an impedance is used that is high enough, up to $d = \text{infinity}$. (Operation to $d = \text{infinity}$ is limited only by the frequency response of the circuit.)

An alternate form of Equation 18 is:

$$F = \frac{1}{\frac{a}{d} + b} \quad (19)$$

Here, it is readily seen that if d approaches infinity then:

$$F_{\infty} = \frac{1}{b}, \quad (20)$$

where F_{∞} is the frequency (in Hz) measured when d is large compared to the probe area. Thus, one constant in the equation is directly measurable by a rather simple procedure—just move the probe far away from the part.

Transposing Equation 18 in terms of d (with F and F_{∞} known) yields:

$$d = \frac{aFF_{\infty}}{F_{\infty} - F}, \quad (21)$$

which may be written:

$$d = aF_{\infty} \left(\frac{F}{F_{\infty} - F} - \frac{F_{\infty} - F_{\infty}}{F_{\infty} - F} \right). \quad (22)$$

This term equals zero.

Rearranging Equation 22 yields:

$$d = aF_{\infty} \left(\frac{F_{\infty}}{F_{\infty} - F} - \frac{F_{\infty} - F}{F_{\infty} - F} \right), \text{ or} \quad (23)$$

$$d = aF_{\infty} \left(\frac{F_{\infty}}{F_{\infty} - F} - 1 \right). \quad (24)$$

Since F_{∞} and a are constants of calibration, it then follows that:

$$d = Mx - N \text{ (a straight line)}, \quad (25)$$

where:

$$M = aF_{\infty}^2,$$

$$N = aF_{\infty}, \text{ and}$$

$$x = \frac{1}{F_{\infty} - F}.$$

It is readily seen that by measuring F_{∞} and d at two points (which may have some small zero error, Δd , such that $d = d_i + \Delta d$) that both the zero error and the constant, a , may be obtained from simultaneous solution of the linear equation (Equation 25) which exactly fixes the calibration equations for the gage. Excellent conformity to these theoretical equations has been observed with gages developed at the Y-12 Plant. In one case, conformity to within 15 microinches was obtained using a 3/8-inch-diameter probe tip and a 0 to 0.100 inch range. This conformity is estimated to be as good as the calibrating micrometer positioning accuracy and hence it is assumed that the actual gage conformity was even better.

Unijunction Probe Circuit Design

In order to provide the highest input impedance and maximum operating frequency, a low-leakage, low-input capacity unijunction transistor such as a Type 2N4853 must be used. Preliminary investigations with this transistor showed that charging resistors as high as two megohms were usable, and that the output pulse amplitude varied directly with the probe input capacitance at these low values of capacity. Hence, as frequency increased, amplitude decreased. A typical probe circuit is presented in Figure 3. Components Q_1 , R_1 , R_2 , R_3 , C_1 , and C_{1a} comprise the measuring circuit. The rest of the circuit is a buffer amplifier which is needed because of the decreasing signal when d is increased. The distance-measuring capacitor, C_1 , is lumped with stray capacities to determine the R_1C_1 time constant of the integrator oscillator circuit. When the voltage across C_1 rises to the peak voltage of the unijunction emitter circuit, the unijunction discharges C_1 through R_3 , generating a positive output pulse. When the capacitor (C_1) is discharged to the valley voltage, the unijunction turns off and C_1 charges up through R_1 again, repeating the cycle. Since the output pulse amplitude is proportional to the value of C_1 , ($C_1 = C_p + C_s$), it was found necessary, when using the 2N4853 unijunction, to increase the stray capacity slightly in order to ensure pulse amplitudes which were large enough to count at high values of the separation distance, d . When this adjustment is accomplished, the F_{∞} decreases proportionately. It also affects the slope at larger values of d . The addition of the required "stray" capacity was accomplished by wrapping a small insulated wire, or "gimmick", around the probe tip support post and grounding the other end of the wire to the common probe. The wire is then slowly unwound and snipped off until the output amplitude is approximately 1/2 volt at F_{∞} . The "gimmick" is represented by C_{1a} in the circuit diagram (Figure 3). A typical output signal range of 0 to 400 KHz was obtained with the probe circuit mounted on an epoxy-fiberglass circuit board, and potted into a 3/4-inch-diameter brass tube using epoxy potting compound. A typical response curve using a 3/4-inch-diameter tip and a 3 by 3-inch flat-plate part is seen in Figure 4.

As Figure 4 indicates, this gage has excellent response and sensitivity in the 0 to 0.100-inch range and is usable to 0.300 inch or more.

When fabricated from standard fixed components, it was necessary to adjust R_2 for the particular unijunction in order to obtain a minimum temperature drift.

Figure 3. UNI-JUNCTION PROBE CIRCUIT.

The matching of probe characteristics for similar calibration curves also proves difficult due to variation in both the unijunction input capacitance and the unijunction peak voltage. With the advent of good, high-quality field-effect transistors (FETs) with low input capacitance (such as the 2N4416), it appeared that matched probes with predictable temperature drift could be obtained using the FET input Schmitt trigger circuit.

Field-Effect Transistor Probe Circuit Design

Design of the FET probe is similar to the unijunction in that the capacitor is charged up toward a high level and then discharged toward a low level using a Schmitt trigger and electronic switch. A FET buffer amplifier is used to couple the RC circuit to the required low input impedance of the Schmitt trigger.

A functional diagram of the probe circuit is given in Figure 5. Ordinarily, Q_1 operates with about a + 1.2-volt bias from source to gate. As a result, when the voltage on C_p is about 4.8V, the input to the Schmitt trigger is at V_h ; when the voltage on C_p is about 3.4V, the input to the Schmitt trigger is at V_L . The Schmitt trigger turns off the electronic switch (ES) (the switch output goes to ground) whenever the input reaches V_h (C_p charges through Resistor R_1 and Thermistor T to 4.8V). The output remains off until the input is reduced to V_L (the voltage on C_p is discharged from 4.8V down toward ground through R_1 , T , and the electronic switch until it reaches 3.4V). The Schmitt trigger then turns the switch back on, applying the "calibrate" voltage to the input RC circuit and repeating the cycle.

A DC signal is taken from the electronic switch (+ when on, ground when off) and this constant amplitude square wave is the frequency output from the probe. The requirements for good stable operation of this circuitry determine the circuit

Figure 4. UNIJUNCTION PROBE RESPONSE CURVE.

configuration and operating conditions. The only known satisfactory input buffer field-effect transistor which is usable with this probe is the 2N4416. All other junction and metal oxide semiconductor FETs examined had too high an input capacitance, or input (source-to-gate) leakage, or both.

Figure 5. FET PROBE FUNCTIONAL DIAGRAM.

The Schmitt trigger is of a special design (see Figure 6) and requires well-regulated + 25 and - 15V power supplies for stable trip points (V_h and V_L). V_h is determined primarily by the 25-volt supply and the passive components R_3 , R_4 , and R_5 . This path can be made very stable. V_L is determined from V_h by the collector current of Q_1 in the "on" state. This path is made very stable by using a large resistance in the emitter circuit (R_7) and a well-regulated - 15V power supply.

The temperature characteristics of this probe can be adjusted by changing the value of R_2 (see Figure 5) in the source of the FET buffer amplifier. Excellent agreement between stray input capacities and temperature response using a standard value of R_2 was obtained with this probe. Two designs of this probe have been utilized, one employing a 4.7K source resistor (R_2) which optimizes the probe for wide temperature excursion and a probe using a 6.8K resistor which has a minimum temperature coefficient around room-temperature operation.

The "calibrate" voltage can be adjusted between + 6 and + 15 volts in order to adjust the value of F_∞ for a particular probe and allow for tolerance in the Schmitt trigger components. Best temperature stability is obtained with the calibrate voltage above + 10 volts (F_∞ high). A typical gage schematic and mechanical drawing are given in the Appendix.

Probe Mechanical Design and Potting

A typical probe design is seen in Figure 7. The electronics are housed inside a 3/4-inch-diameter tube, coupled to the tip with a Size 4-40 threaded stud and

Figure 6. SCHMITT TRIGGER SCHEMATIC.

shaft soldered to the printed circuit board. A replaceable tip of any desired size may be attached to the threaded stud and comprises the measuring plate of the gage. The size and shape of this plate affects the gage sensitivity and gaging range (as indicated in the section on Variable Oscillator Capacitance Gages, Page 7). Gaging tips from 3/16 inch in diameter to 1 inch square have been used with excellent results, but the more common sizes are 3/8, 1/2, and 3/4 inch in diameter.

Originally, the electronics for these probes were mounted on epoxy-fiberglass printed circuit boards with cable attached, then potted into the outer shield tubes. A wire lead was brought from the circuit ground and soldered to the outer tube to ensure shielding. This method of fabrication proved to be acceptable for operation in the temperature range from 10 to 45° C (50 to 113° F). However, large unacceptable temperature errors occurring in the 55 to 70° C region were attributed to variation in the dielectric constant or resistivity of the epoxy potting compound. In order to overcome this problem, the oscillator circuit was left unpotted (only the amplifier or electronic switch part of the board was potted in

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Figure 7. CAPACITANCE PROBE PHYSICAL CONSTRUCTION.

epoxy to hold the board into the tube) and a Teflon dust seal was attached to the probe end. The latter method of fabrication was satisfactory for probes operating in the 10 to 85° C (50 to 185° F) region, which is expected to be the limit for ultrastable operation of the semiconductors in the circuit. It is presumed that these units will also operate satisfactorily below 0° C, but the limit of the tests conducted when cooling below room temperature was 10° C.

Compensation, Stabilization, and Calibration

Temperature compensation of these probes is a complex problem since none of the circuits is exactly linear. The input leakage of the unijunction of the FET transistor can be expected to double for every 10° C increase in temperature so that obviously these devices must be selected for their as-low-as-possible initial leakage at 25° C. Eventually, however, this leakage becomes significant compared to the current through the charging resistor (R_1), and an increase in operating frequency occurs.

The expected maximum operating temperature (for which the frequency increase is 0.05% or less) for the 2N4853 unijunction with a 1.5-megohm charging resistor and 25-volt DC supply is about + 50° C. As noted before, the maximum operating temperature of the FET probe can be increased by adjusting the calibrate voltage to maximum, and a frequency increase of 0.05 percent would occur near + 80° C.

The temperature response of the FET probe can also be affected by changes occurring in the charging resistor. To offset these changes, a 10K thermistor is used in series with the 1/8-watt composition resistor. The DC temperature coefficient of the FET V_{gs} voltage can also affect the stability. The value of the source resistor is adjusted to control this problem. A value of 4.7K for the source resistor (R_2) gives a slightly decreasing temperature coefficient near room temperature, and balances the curve for minimum temperature error in the + 10 to + 85° C region, while a value of 6.8K provides for a flat characteristic near room temperature. A typical temperature characteristic curve for the unijunction probe is presented in Figure 8; typical curves for the FET probe are seen in Figure 9.

If the probe is to be operated at the highest permissible temperature, the F_{∞} value should be adjusted to maximum in the calibration procedure. Obviously, Curve b (Figure 9) is more useful for laboratory or inspection measurements for extreme precision and accuracy while Curve a provides the highest operating temperature range for use near welders, furnaces, or in other high-ambient-temperature locations.

The selected calibration method used was a 3 by 3-inch-square target mounted on a micrometer-actuated slide with the probe mounted in a vise (fixed position) positioned so that it was near the center of the target and the probe face parallel to the target face. Motion of the target away from the probe (d) was read from the micrometer. The conformity and stability of the probe was checked and typical

Figure 8. UNIJUNCTION PROBE TEMPERATURE CURVE.

fine structure curves compared to the theoretical equation (Equation 24). Agreement within ± 15 microinches was obtained for two FET probes fabricated to Curve b at the normal laboratory ambient temperature range (68 to 75° F) for measurements taken on three successive days.

The normal calibration procedure involves the following steps:

1. Measure F_{∞} (probe in air with no target closer than 6 inches), then move the target in until the frequency counter just stops counting (which will occur just before contact between the probe and target, and creates a small zero error, Δd).
2. Record the micrometer reading. This value is zero d_i . All other values of d_i are determined by difference from this reading.
3. Move the target until a separation distance of 20 to 30 mils is indicated by the micrometer. Record d_i and F for this distance.
4. Move the micrometer for a separation distance of 50 to 60 mils, and again record d_i and F .

Values of a and Δd can then be obtained by solving the two simultaneous equations of the form:

$$d_i + \Delta d = \frac{aF_{\infty}^2}{F_{\infty} - F_i} - aF_{\infty} \quad (26)$$

Since d_i , F_{∞} , and F_i were measured, only a and Δd are unknown.

Figure 9. FET PROBE TEMPERATURE CURVES. (Dotted Line Shows Epoxy Potted Probe)

The zero error (Δd) caused by the gage arc out at close proximity to the target is usually very small (being in the order of 0.7 to 2.0 mils, depending upon the probe diameter, atmospheric humidity and pressure, target and probe surface roughness, and probe alignment). If excessively large zero errors are encountered, then probe normality is suspect and should be examined. A warped target could also contribute to this error.

Solution of the two simultaneous equations (Equation 26) for "a", along with a measurement of F_{∞} , completes the probe calibration.

APPLICATIONS

Capacitance probes are readily applicable to the noncontact measurement of part dimensions (such as length and thickness) for dimensional inspection and product certification. Noncontact capacitance measurements are desirable for parts with a high finish or delicate surface which are easily deformed by contact pressure, or to obtain an average measurement for a defined area. The only requirement for these measurements is that the part being measured must be a conducting or semiconducting material so that the effective series resistance is low compared to the 1.5-megohm charging resistance, and that the part be connected to a common circuit. The extremely high input impedance of these probes allows considerably more part resistance than existing commercial probes before appreciable errors are introduced.

As with all measuring applications, the system performance may be degraded by unstable mounting and fixturing of the probe or part. In addition, when making

capacitance measurements, care must be taken to shield the probe from any moving objects (since the effects of changing the stray capacitance can cause measurement errors). Personnel, too, can be a problem if they locate their hands or bodies within 10 inches or so of the probe while measurements are being made.

In addition to dimensional inspection, the probes can be used for dielectric constant or dielectric homogeneity measurements of such materials as plastic, fiberglass/resin composites, and paint films, when a separate thickness measurement can be made by other gaging methods and a conductive backing is provided. In this case, the dielectric constant is the ratio of the actual dielectric thickness to the (air calibrated) indicated thickness from the capacitance probe.

Use of the gage for direct computer processing is obvious. It may also be used with direct linearized digital processing, as outlined in Figure 10. The probe or F_{∞} reference oscillator of Figure 10 is adjusted so that the frequencies are matched at infinite d . The output of the counter is then $(F_{\infty} - F)/10,000$ and the period meter measures the inverse of this value. The time base of the period meter could be adjusted so that the readout is calibrated to the desired engineering units and an analog linear record obtained from the DAC and Recorder. This readout has a zero offset but is linear and can have an accuracy approaching 0.01 percent in the lower values of d .

Figure 10. LINEAR DIGITAL READOUT FOR CAPACITOR PROBES.

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APPENDIX

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ITEM 1 - TIP

SCALE: 2"=1"

MAT. - YELLOW BRASS

ITEM 2 - SLEEVE

SCALE: 2"=1"

MAT. - YELLOW BRASS OR FREE MACHINE BRONZE
TOL. - FRACTIONS $\pm \frac{1}{64}$ "
DECIMALS AS INDICATED

ITEM 3 - TIP COUPLER

SCALE: 4"=1"

MAT. - YELLOW BRASS

ITEM 4 - DUST SEAL

SCALE: 2"=1"
MAT. - TEFLON

ASSEMBLY

SCALE: 4"=1"

REF. DWGS.

E-I-72334 CAPACITANCE DISTANCE GAGE-P.C. BD.
E-I-72350 FLAT PLATE DENSITY GAGE
CAPACITANCE PROBE SCHEMATIC

CAPACITANCE GAGE. (Probe Assembly)

CAPACITANCE GAGE. (Schematic)