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Two-Terminal Dielectric Measurements Up to 6×10^8 Hz

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A two-terminal dielectric specimen holder has been constructed and used to make dielectric constant and loss measurements on a single disk specimen at room temperature over a frequency range from 10^{-2} to 6×10^8 Hz. The measurement procedures are outlined and a detailed analysis of the working equations and measurement errors is presented.

1. Introduction

It is often important for the study of dielectric properties that measurements be made on a single specimen using a minimum amount of the material over as wide a range of temperatures and frequencies as possible. Low-frequency measurements can conveniently be made on a single disk-shaped specimen, but at higher frequencies conventional equipment requires the use of cylindrical or rectangular specimens of various sizes to fit different microwave guides and cavities. If the dielectric properties of a material are sensitive to differences in specimen preparation, then the use of different specimens to cover the desired frequency range could lead to results which seriously misrepresent the frequency dependence of the dielectric constant and loss even though the individual measurements may themselves be quite accurate.

The two-terminal parallel-plate dielectric specimen holder described in this paper was designed to make measurements on single disk-shaped specimens (about 1 to 2 g of material) over as wide a range of frequencies as possible. Minimum specimen and holder size together with simplicity of construction and operation have been emphasized in order that the cell could easily be fitted with an insulating jacket and operated at temperatures from −200 to +200 °C. The variable temperature operation of the holder will be reported in a later paper.

In this paper, we will discuss the construction of the holder currently in operation at room temperature, and develop the equations for its use with the low-frequency Scheiber bridge [1],[1] the General Radio Models 716C Schering Bridge and 1615A Transformer Bridge, and the Boonton Radio Corporation Models 260A, 280A, and 190A Q-meters. Dielectric constant and loss data obtained with the holder for a disk of commercial poly(methyl methacrylate) over a frequency range of 10^{-2} to 6×10^8 Hz will be presented to illustrate the use of the holder. The experimental errors will be analyzed and discussed in detail. Use of the holder to measure liquid specimens is described in a separate paper [2]. Some of the measurement techniques used here are related to techniques described in a previous publication from this laboratory [3].

[1] Figures in brackets indicate the literature references at the end of this paper.

2. Design and Construction of Specimen Holder

A schematic diagram of the specimen holder is shown in figure 1. The case, cap, and electrodes are made of brass, gold-plated to reduce surface losses at high frequency. The lower electrode, which is operated at high potential, is supported and insulated from the case by a fused silica ring, which in turn is supported by the bottom rim on the holder case. The lower electrode and silica ring are held fixed by small beads of epoxy cement. The upper electrode is connected securely to a brass pin, finely threaded on the upper end to accommodate the spacing nut.

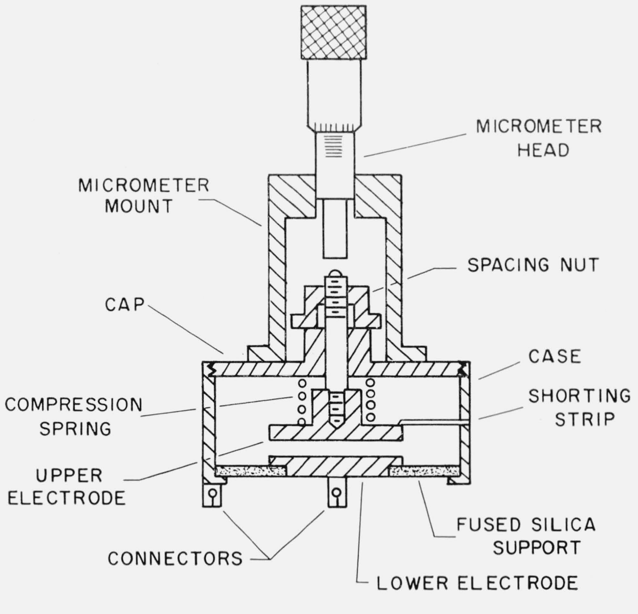

FIGURE 1. *A schematic diagram of the two-terminal dielectric specimen holder.*

The compression spring holds the spacing nut against the shoulder on the holder cap and rotation of the spacing nut raises or lowers the electrode and hence regulates the spacing. The spacing between the electrodes is determined by measuring the position of a steel ball, mounted at the top of the upper electrode assembly, with a micrometer barrel mounted rigidly to the holder cap. Electrical connection to the upper (ground potential) electrode is made by soldering four flexible gold foil strips between the upper electrode and case. Such an arrangement was preferred to conventional metal bellows mounting because of the former's shorter electrical path. Electrical connection to the holder is made through two connector caps soldered to the case and lower electrode, which fit snugly over pins suitably attached to the measuring equipment. The holder electrodes are 2.54 cm in diameter. The micrometer measures 10 μ per division and readings are estimated to 1 μ ($1\ \mu=10^{-6}$ m).

3. Theory of Measurements

Each dielectric constant and loss determination involves a specimen in and specimen out measurement of the holder impedance at the measurement frequency, and two calibration measurements of the empty holder capacitance at some convenient audio frequency. The calibration measurements can either be done separately and compiled in the form of a curve of measured capacitance versus electrode spacing which is then applicable to all measurement frequencies, or they can be performed at the time of each determination with increased accuracy.

Figure 2 shows a schematic diagram of the two-terminal holder for the two measurement frequency conditions, together with the assumed equivalent circuit and the effective series resistance and capacitance of the cell. The symbols used in figure 2 are explained in table 1. The two audio-frequency calibration measurements can be represented in a way similar to figure 2b except that the effects of lead inductance can be ignored. When the measurement frequency is itself in the audio range then the general measurement procedure outlined here can be simplified as is described in the following section.

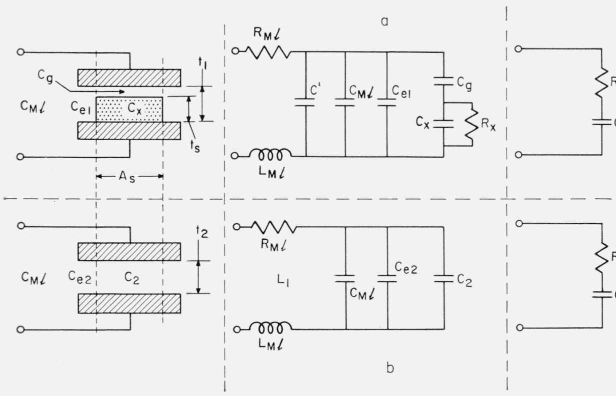

FIGURE 2. *A schematic representation of the two-terminal holder with (a) and without (b) a specimen and the corresponding assumed equivalent circuits.*

The symbols are defined in table 1.

TABLE 1. *Description of symbols used in text*

$C_x=\epsilon'\epsilon_0\dfrac{A_s}{t_s}$=equivalent parallel capacitance of the specimen.

$C_v=\epsilon_0\dfrac{A_s}{t_s}$=equivalent parallel vacuum capacitance of the specimen.

R_x=equivalent parallel resistance of the specimen.

$C_g=\epsilon_0\dfrac{A_s}{t_1-t_s}$=capacitance of the airgap between the specimen and electrode.

$C_2=\epsilon_0\dfrac{A_s}{t_2}$=capacitance berween adjacent areas A_s of the electrodes. C_1 corresponds to t_1.

C_c=the capacitance reading of the standard capacitor in a bridge when the cell is connected. C_{c1} corresponds to t_1 and C_{c_2} to t_2.

C_e=capacitance between the remainder of the electrodes including that portion outside the area A_s and the edges. C_{e_1} corresponds to spacing t_1 and C_{e_2} to t_2.

C_{Ml}=capacitance between the leads when the holder is in the measurement circuit.

C'=error capacitance due to distortion of the electric field (see discussion of errors).

R_{Ml}, L_{Ml}=equivalent resistance and inductance of holder leads when the holder is in the measurement circuit.

R_{Mi}, C_{Mi}=effective series resistance and capacitance of holder with specimen in.

C_{lc}=capacitance between the leads when the holder is in the calibration circuit.

Δf=the difference in frequency between the upper and lower half-power points of an L–C–R series resonant circuit.

R_{Mo}, C_{Mo}=effective series resistance and capacitance of holder with specimen out.

t_s, A_s=thickness and area of specimen.

t_1, t_2=separation of electrodes at settings 1 and 2.

ϵ'=dielectric constant.

ϵ_0=permittivity of free space.

ϵ''=dielectric loss.

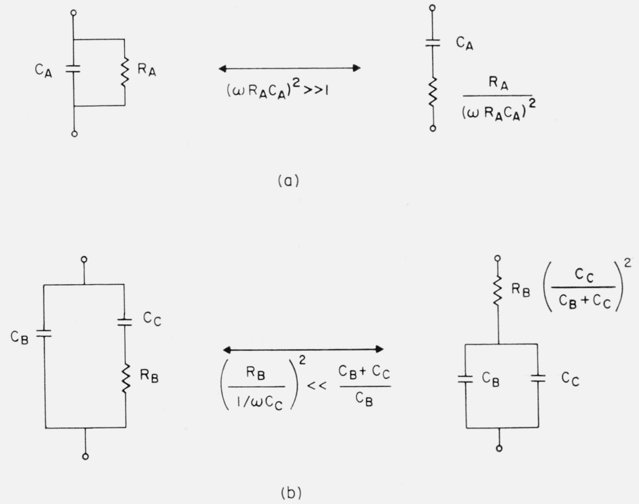

FIGURE 3. *Two simple transformations used to derive the working equations.*

In order to relate the equivalent circuits in figure 2 to their effective measured circuits, the approximate transformations shown in figure 3 were used. The approximations will be suitable for most low loss materials such as many polymers, glasses, and organic compounds. The validity criteria are shown in figure 3, and in general if $\epsilon''/\epsilon' < 0.1$ the transformation approximations will introduce less than 1 percent error in the results.

At the measurement frequency, the effective measured series capacitance and resistance of the holder with specimen in, C_{Mi} and R_{Mi}, and with specimen out, C_{Mo} and R_{Mo}, can be expressed as follows:

$$C_{Mi}=\frac{C_{Hi}}{1-\omega^2 L_{Ml} C_{Hi}} \tag{1}$$

$$R_{Ml}=R_{Ml}+\frac{R_x}{(\omega C_x R_x)^2}\left[\frac{C_x C_g}{C_{Hi}(C_x+C_g)}\right]^2 \tag{2}$$

$$C_{Mo}=\frac{C_{Ho}}{1-\omega^2 L_{Ml} C_{Ho}} \tag{3}$$

$$R_{Mo}=R_{Ml} \tag{4}$$

where

$$C_{Hi}=C'+C_{Ml}+C_{e_1}+\frac{C_x C_g}{C_x+C_g} \tag{5}$$

and

$$C_{Ho}=C_{Ml}+C_{e_2}+C_2. \tag{6}$$

One can write an expression for C_x which is independent of frequency or the type of electrical equipment used. If C_{Mo} is experimentally set equal to C_{Mi}, eqs (1) and (3) can be equated with the result,

$$C_x=\frac{C_g[C_2+(C_{e_2}-C_{e_1})-C']}{C_g-[C_2+(C_{e_2}-C_{e_1})-C']}. \tag{7}$$

From the calibration measurements (circuits correspond to that of figure 2b ignoring the lead inductance),

$$C_{c_1}=C_{lc}+C_{e_1}+C_1$$

$$C_{c_2}=C_{lc}+C_{e_2}+C_2, \tag{8}$$

one finds

$$C_2+(C_{e_2}-C_{e_1})=(C_{c_2}-C_{c_1}+C_1), \tag{9}$$

which can be substituted into eq (7) to give

$$C_x=\frac{C_g[(C_{c_2}-C_{c_1})+C_1-C']}{C_g-[(C_{c_2}-C_{c_1})+C_1-C']}. \tag{10}$$

Writing C_g and C_1 in the numerator in terms of measured quantities and applying the definition of ϵ' one finds

$$\epsilon'=\frac{C_x}{C_v}=\frac{1}{C_v}\left[(C_{c_2}-C_{c_1})+\frac{\epsilon_0 A_s}{t_1}-C'\right] \times\left[1-\frac{t_1-t_s}{t_1}-\frac{(C_{c_2}-C_{c_1})}{C_g}+\frac{C'}{C_g}\right]^{-1}. \tag{11}$$

The last term in brackets is the airgap correction term whose value is close to unity. It should be noted that this airgap term will be used in this form throughout the text. The effect of the error capacitance C' is also small (see discussion of errors) and hence the value of ϵ' is primarily dependent on the difference between two empty holder capacitances measured at audio frequencies and on the sample dimensions and initial holder spacing.

Since the holder is a part of the measurement circuit when the specimen is both in and out of the holder, the lead losses R_{Ml} can be eliminated by combining eqs (2) and (4), giving,

$$\Delta R_M=R_{Mi}-R_{Mo}=\frac{C_x D_x}{\omega}\left[\frac{C_g}{C_{Hi}(C_x+C_g)}\right]^2 \tag{12}$$

or

$$\epsilon''=\frac{C_x}{C_v}D_x=\frac{\omega}{C_v}\,\Delta R_M\left[\frac{C_{Hi}(C_x+C_g)}{C_g}\right]^2. \tag{13}$$

Unfortunately, there is not a convenient way to make resistance substitutions for the specimen loss so that the calculation of R_x will depend on the particular methods used at the measurement frequency.

4. Measurement Procedure

4.1. Q-Meter Procedure of Measurement

For Q-meter measurements in the frequency range 10^5 Hz to 2.5×10^8 Hz, two Q-meters were used, the Boonton Radio 260A Q-meter (100 kHz to 20 MHz) and the Boonton Radio 190A Q-meter (20 MHz to 250 MHz). The measurement procedure used is as follows: The holder is mounted on top of the Q-meter and connected in parallel with the Q-meter capacitor, C_s, with the shortest possible leads. With the specimen in the holder, the circuit is brought to resonance by tuning the Q-meter capacitor, C_s, to C_s'. The values of C_s', the electrode spacing t_1, and the Q of the circuit with the specimen in the holder, Q_i, are recorded. The specimen is then removed from the holder and the electrode spacing is decreased until resonance is achieved while holding C_s fixed at C_s'. The electrode spacing t_2 and the Q of the circuit with the specimen out, Q_o, are recorded.

The two resonant situations, with the specimen in and with the specimen out, occur at the same frequency and the total capacitance in the resonant circuits must be the same in both cases. The value of the total capacitance, C_T, is taken as the value of C_s when the Q-meter is tuned to resonance with the holder removed.

For the Q-meter measurement, the circuits of figure 2 may be viewed in parallel with the Q-meter capacitance C'_s. Considering figure 2a, one finds that if C_{Mi} and R_{Mi} are placed in parallel with C'_s then the transformation of figure 3b can be used to reduce the circuit to a simple L–C–R series circuit, for which the total series capacitance, the total series resistance and the Q of the resonant circuit with the specimen in the holder, C_T, R_{si}, and Q_i, are given as

$$C_T = C_{Mi} + C'_s, \tag{14}$$

$$R_{si} = R_{Mi}\left(\frac{C_{Mi}}{C_T}\right)^2, \tag{15}$$

and

$$1/Q_i = \omega R_{si} C_T \tag{16}$$

where C_{Mi} and R_{Mi} are defined by eqs (1) and (2).

Likewise, with the specimen out and with the holder at spacing t_2, the resonant circuit can be represented by figure 2b in parallel with C'_s. The total capacitance, the total series resistance and the Q of the resonant circuit with the specimen out of the holder, C_T, R_{so}, and Q_o, are given as

$$C_T = C_{Mo} + C'_s, \tag{17}$$

$$R_{so} = R_{Mo}\left(\frac{C_{Mo}}{C_T}\right)^2 \tag{18}$$

and

$$1/Q_o = \omega R_{so} C_T, \tag{19}$$

where C_{Mo} and R_{Mo} are given by eqs (3) and (4).

Equating the expression for C_T in eqs (14) and (17) will yield eq (11) of the previous section for the measured value of ϵ'.

An expression for ϵ'' is obtained from the difference, $Q_i^{-1} - Q_o^{-1}$, using eqs (16) and (19). Thus,

$$\Delta\left(\frac{1}{Q}\right) = Q_i^{-1} - Q_o^{-1} = \frac{\omega R_x}{(\omega R_x C_x)^2}\left[\frac{C_x C_g}{C_{Hi}(C_x + C_g)}\right]^2 \left(\frac{C_{Mo}}{C_T}\right)^2 C_T. \tag{20}$$

Rearranging, dividing through by C_v, and using $C_{Hi} = C_{Ho}$, one obtains,

$$\epsilon'' = \frac{1}{\omega R_x C_v} = \left[\frac{C_T}{C_v}\Delta\left(\frac{1}{Q}\right)\right]\left(\frac{C_{Ho}}{C_{Mo}}\right)^2 \left(\frac{C_x + C_g}{C_g}\right)^2. \tag{21}$$

In the above equation, the term in brackets is the dominant term and it alone would be present if there were no airgap or lead inductance. The term $(C_{Ho}/C_{Mo})^2$ results from the effect of lead inductance, which becomes significant above 30 MHz (see eq (3)). The third term, $[(C_x + C_g)/C_g]^2$, is the airgap correction.

Equation (21) can be reduced to following working equation:

$$\epsilon'' = \frac{C_T}{\epsilon_o A_s / t_s}\Delta\left(\frac{1}{Q}\right)\left(\frac{C_{Ho}}{C_T - C'_s}\right)^2 \left[1 - \frac{t_1 - t_s}{t_1} - \frac{C_{c_2} - C_{c_1}}{C_g} + \frac{C'}{C_g}\right]^{-2}. \tag{22}$$

In the above equation C_{Mo} has been replaced by $C_T - C'_s$ in accordance with eq (17). The capacitances, C_T and C'_s, are obtained from the calibrated Q-meter capacitor. The capacitance C_{Ho} is obtained from the audio frequency calibration. Omitted from the above equations are the high-frequency corrections which must be applied to the values of Q and C_s as read from the 190A Q-meter. The manner in which these corrections are carried out is explicitly stated in the 190A Q-meter manual.

A more detailed description of Q-meter dielectric measurements may be obtained from a paper by Hazen [4] in which he shows how the Q-meter circuit may be adapted for the measurement of very low losses.

4.2. Two-Terminal Measurements at Audio Frequencies

The General Radio (GR) 716C Schering bridge was used for audio frequencies from 10^2 Hz to 10^5 Hz, and the GR 1615A transformer ratio arm bridge was used from 5×10^2 Hz to 2×10^4 Hz.

For audio frequencies the effects of the lead inductance can be ignored, and the circuits of figure 2 without the inductance are considered here. With these bridges one measures the dissipation factor and the effective series capacitance of the unknown. However, when $\epsilon''/\epsilon' < 0.1$ the series capacitance and the parallel capacitance of the unknown are equal (within 1 percent) and this case is treated here. For $\epsilon''/\epsilon' > 0.1$ the GR bridge manuals should be consulted.

If the holder and specimen are connected as a two-terminal device in an arm of the bridge and if the procedure of measurement which was used with the Q-meter is also used with the bridge, then the result for ϵ' is the same as eq (11) and

$$\epsilon'' = \frac{C_T}{C_v}(D_i - D_o)\left(\frac{C_x + C_g}{C_g}\right)^2, \tag{23}$$

where D_i and D_o are the dissipation factors with the specimen in and out of the holder, and C_T is the total capacitance in the arm of the bridge which contains the holder. Just as with the Q-meter measurement, C_T must be the same for the D_i and D_o measurements.

A variation on the above experimental procedure which utilizes the standard capacitor in the bridge is preferable when using a bridge circuit and null detector for a two-terminal measurement. First, a bridge balance is obtained with the specimen and holder in the bridge circuit. Then t_1, D_i, and the standard bridge capacitor reading C_{si} are recorded.

The sample is taken out of the holder. Keeping the spacing fixed at t_1, the capacitance of the standard capacitor, C_s, is varied to rebalance the bridge. The specimen-out values, C_{so} and D_o are recorded.

If a bridge measurement is made by changing the standard capacitor while holding the electrode spacing at t_1, then a distinction must be made between using the substitution method of measurement and using the bridge as a direct reading instrument. In the substitution method, the unknown is connected in parallel with the standard capacitor, whereas for the direct method the unknown is connected in the bridge arm electrically opposite the standard capacitor. The GR 1615A is most conveniently used to measure C and D directly. The substitution method is used with the GR 716C bridge by connecting a precision air ballast capacitor to the "unknown direct" terminals.

When the electrode spacing is t_1 for both "in" and "out" measurements and a direct measurement is made, the total capacitance in the unknown bridge arm will be different for the "in" and "out" balances. Let C_{Ti} and C_{To} be the total capacitance in the unknown bridge arm with the sample in and out of the holder respectively. For a direct measurement we have,

$$C_{Hi}=C_{Ml}+C_{e_1}+C'+\frac{C_xC_g}{C_x+C_g}, \tag{24}$$

$$C_{Ho}=C_{Ml}+C_{e_1}+C_1 \tag{25}$$

from which ϵ' can be calculated in terms of measurable quantities. Using $C_{Hi}-C_{Ho}=C_{si}-C_{so}$ we have for a direct measurement

$$\epsilon'=\frac{1}{\epsilon_0A_s/t_s}\left(C_{si}-C_{so}+\frac{\epsilon_0A_s}{t_1}-C'\right)\times\left[1-\frac{t_1-t_s}{t_1}-\frac{C_{si}-C_{so}}{C_g}+\frac{C'}{C_g}\right]^{-1} \tag{26}$$

and

$$\epsilon''=\frac{C_{Ti}}{\epsilon_0A_s/t_s}\left(D_i-\frac{C_{To}}{C_{Ti}}D_o\right)\left[1-\frac{t_1-t_s}{t_1}-\frac{C_{si}-C_{so}}{C_g}+\frac{C'}{C_g}\right]^{-2}. \tag{27}$$

For a direct measurement with the GR 1615A transformer bridge it is conveniently true that $C_{To}=C_{so}$ and $C_{Ti}=C_{si}$.

On the other hand, for the substitution measurement, $C_{To}=C_{Ti}=C_T$ and $C_{si}<C_{so}$. For ϵ' and ϵ'' we have,

$$\epsilon'=\frac{1}{\epsilon_0A_s/t_s}\left(C_{so}-C_{si}+\frac{\epsilon_0A_s}{t_1}-C'\right)\times\left[1-\frac{t_1-t_s}{t_1}-\frac{C_{so}-C_{si}}{C_g}+\frac{C'}{C_g}\right]^{-1} \tag{28}$$

and

$$\epsilon''=\frac{C_T}{\epsilon_0A_s/t_s}(D_i-D_o)\left[1-\frac{t_1-t_s}{t_1}-\frac{C_{so}-C_{si}}{C_g}+\frac{C'}{C_g}\right]^{-2} \tag{29}$$

where the bracketed term in eqs (26) to (29) is the airgap correction term. To obtain the total capacitance, C_T, in the unknown arm of the GR 716C Schering bridge a separate bridge measurement is necessary. During the "in" and "out" measurement the ballast capacitance, C_B, remains fixed at C_T. After the specimen holder has been disconnected, C_B is measured by switching the bridge to "direct" and balancing the bridge by changing the standard capacitor. The dial reading of the standard capacitor is calibrated by measuring a known C_B. The calibration should also include the 1 pF capacitance of the bridge terminals.

4.3. Measurements at Ultra-Low Frequencies

The Scheiber ultra-low-frequency bridge was used in the frequency range 10^{-2}Hz to 2×10^2Hz [1]. Although the bridge was designed for use with a three-terminal dielectric cell, we have found that it may be easily adapted for measurements with a two-terminal cell. Ordinarily the leads from the bridge to the electrodes are shielded at ground potential and the shield encompasses the entire cell. For the two-terminal holder the electrode leads are shielded but the shield is abandoned at the base of the holder where the electrodes connect. A shield around the entire cell was improvised by lowering an aluminum box over the holder and connecting the the box to ground.

The substitution method of measurement with the electrode spacing fixed at t_1 for the "in" and "out" bridge balance is used. Equations (22) and (23) of Scheiber's paper [1] in conjunction with the airgap correction term of this paper were used to calculate the capacitance and conductance of the unknown.

For measurements below 10 Hz the surface conductivity of the sample can become a problem. It can be eliminated by wiping and cleaning the edges of the sample and thereafter handling the sample with tweezers and by introducing dry nitrogen through one of the ports of the cell.

4.4. Self-Resonant Measurements

The two-terminal holder can be used up to 600 MHz by shorting the terminals with an appropriate length conductor and exciting the resulting circuit at its self-resonant frequency. The Boonton model 280A UHF Q-meter was used as a combined oscillator and detector. The oscillator was coupled to the holder through a variable attenuator by means of a single-loop inductive probe. The signal was picked up with a capactive probe and rectified to give a d-c voltage which was applied to a voltemeter in the model 280A. The voltmeter is calibrated in full and half-power levels which can be conveniently checked by switching 3 dB into the inductive probe. The oscillator frequency was measured with a Hewlett Packard model 524C frequency meter and 540A transfer oscillator.

Measurements were made by placing the specimen in the holder and tuning the circuit by adjusting the oscillator frequency. The frequency corresponding to the half-power points is then determined and recorded along with the electrode spacing. The specimen is then removed and the circuit retuned by adjusting the holder electrode spacing. The frequencies corresponding to the half-power points are again measured and recorded. One or more additional measurements of electrode spacing and corresponding resonance frequency must be made in order to determine the change in capacitance with resonance frequency.

Calculations are based on eqs (1) through (6), where the terminals in figures 2a and 2b are shorted. Thus we treat the problem as a simple series $L_TC_TR_T$ circuit for which the dissipation factor D_T and total capacitance C_T are

$$D_T=\omega C_T R_T, \tag{30}$$

$$C_T=\frac{1}{\omega^2 L_T}. \tag{31}$$

One may obtain ϵ' from eq (11) and ϵ'' from eq (13) remembering that here $C_{Hi}=C_T$ and

$$\Delta D_M=D_{Mi}-D_{Mo}=\omega C_T(R_{Mi}-R_{Mo}). \tag{32}$$

Substituting (32) into eq (13) gives

$$\epsilon''=\frac{C_T}{C_v}\Delta D_M\left(\frac{C_x+C_g}{C_g}\right)^2. \tag{33}$$

The ratio of the frequency separation of the half-power points to the resonance frequency is measured to obtain the dissipation factor. The total capacitance which cannot be measured directly can be obtained by differentiating eq (31) to obtain

$$\frac{dC_T}{df_M}=-\frac{2}{f_M}C_T. \tag{34}$$

Hence we can write

$$\epsilon''=-\frac{dC_T}{df_M}\cdot\frac{\Delta f_i-\Delta f_0}{2C_v}\left(\frac{C_x+C_g}{C_g}\right)^2. \tag{35}$$

Since the only part of C_T which changes with spacing, t, is the calibration capacitance C_c of the holder one can write

$$\epsilon''=-\frac{dt}{df_M}\cdot\frac{dC_c}{dt}\cdot\frac{\Delta f_i-\Delta f_0}{2C_v}\left(\frac{C_x+C_g}{C_g}\right)^2. \tag{36}$$

Finally

$$\epsilon''=-\frac{dt}{df_M}\cdot\frac{dC_c}{dt}\cdot\frac{\Delta f_i-\Delta f_0}{2\epsilon_0 A_s/t_s}\times\left[1-\frac{t_1-t_s}{t_1}-\frac{C_{c_2}-C_{c_1}}{C_g}+\frac{C'}{C_g}\right]^{-2}. \tag{37}$$

5. Sample Data

In order to illustrate the results which can be obtained with the dielectric specimen holder described in this paper, room temperature measurements were made on two specimens machined from the same 2.54-cm-diam poly(methyl methacrylate) rod into disks of average thicknesses 0.3150 cm and 0.1577 cm. The disk faces were made flat and parallel to better than 10 μ and no contact electrodes were applied. Both specimens were kept at about 23 °C at a relative humidity of 50 percent for several months before being measured. The measuring techniques used are those described in detail in the previous sections, and the measurements on the two specimens over the whole frequency range took about 10 hr. Calculations of ϵ' and ϵ'' were done by hand. The results of the measurements are shown in figure 4 which is a plot of the values of ϵ' and ϵ'' versus log frequency for the 0.3150-cm-thick specimen of poly(methyl methacrylate).

The data for the two specimens measured were found to agree to within 1 percent for ϵ' and 5 percent for ϵ'' at all frequencies, and with just a few exceptions, the agreement was within 0.3 percent for ϵ' and 2 percent for ϵ'' over the frequency range covered. In addition, the highest frequency values, which are the most crucial as far as the magnitude of the circuit correction is concerned, agree with each other and with measurements made using a reentrant cavity [3] within 0.3 percent in ϵ' and 2 percent in ϵ''. Additional measurements made on specimens of polychlorotrifluoroethylene and fused silica gave results which were also accurate to within the limits claimed in this paper except for the loss in fused silica which could not be detected at high frequency.

6. Discussion of Measurement Errors

The two-terminal method described in this paper was intended to provide values of ϵ' and ϵ'' over the entire range of applicable frequencies without imposing rigid requirements on specimen preparation and cell construction. In addition to uncertainties in the quantities explicitly appearing in the working equations of the previous section, attention must be given to the assumptions and approximations upon which these equations are based, the errors due to which will appear in an error capacitance term C'.

The error capacitance (in fig. 2 and eq (5)) includes corrections to the ideal assumed capacitance

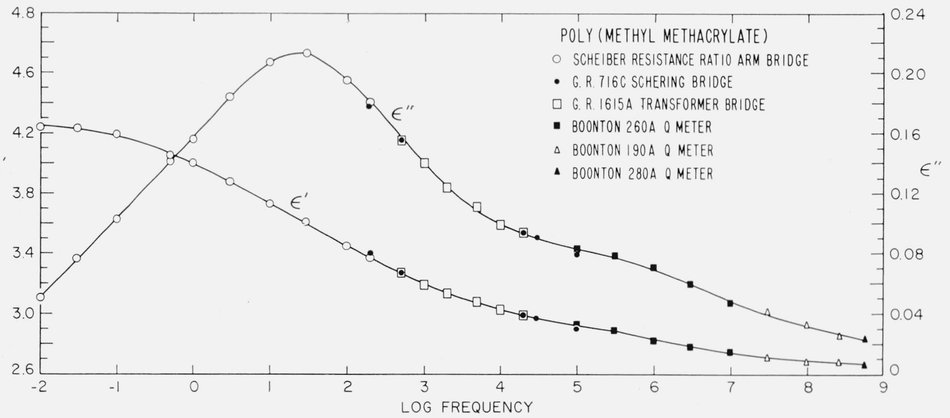

FIGURE 4. *The dielectric constant (ϵ') and loss index (ϵ'') of a 0.3150-cm-thick disk of commercial poly(methyl methacrylate) as a function of frequency as measured with the two-terminal dielectric specimen holder using various measuring equipment.*

resulting from distortions of the electric field between the measuring electrodes because of (a) tilted, misalined or nonflat electrode surfaces, (b) the presence of an airgap, (c) departure of the specimen geometry from an ideal right circular cylinder, (d) electrode edge effects, and (e) the presence of air between the electrodes.

(a) The errors due to nonideal electrodes are all second order effects with the actual capacitance given by

$$C=\epsilon'\epsilon_0\frac{A}{t}\left[1+0\left(\frac{\Delta}{t}\right)^2\right]$$

where A and t are the electrode area and separation and Δ represents the variation in t in the case of tilted or uneven electrodes or the distance between axes of the electrodes in the case of misalined electrodes. It is not difficult to keep mechanical errors to within 25 μ (25 μ=0.001 in.). With typical separations, t, of 1000 μ these errors amount to roughly <0.1 percent. In addition, the tilt error increases the capacitance whereas the misalinement error decreases it. Also since each determination of C involves a difference between two measurements, the effects tend to cancel. Thus the mechanical construction need not require special attention.

(b) The airgap causes some distortion of the field at the edge of the specimen (which increases with gap width and dielectric constant) and makes the actual capacitance higher than its uncorrected value. This in turn leads to a high value for the specimen capacitance C_x. An upper limit to this error ($\epsilon'=\infty$) for a typical measurement was calculated to be 0.2 percent [6], and one can reasonably conclude that in most cases the error from this source is <0.1 percent.

(c) Geometric imperfections in the specimen also lead to distortions. Rounded corners would cause an error similar to the airgap error. Uneven or nonparallel specimen surfaces lead to an actual capacitance greater than that assumed. The airgap capacitance is particularly affected by nonideal specimen surfaces. The specimen surface may well be tilted with respect to the adjacent electrode surface so that the airgap is a wedge of average thickness t=15 μ and variation Δ=10 μ. In this case the assumed airgap capacitance may be in error by 20 percent. (This calculation is based on the worst case—the capacitance of two concentric but mutually tilted circular disks in the limit of minimum separation.) Fortunately, the airgap capacitance enters the calculations in a very insensitive way so that an error of 20 percent in C_g would result in an error of perhaps 0.5 percent in C_x, and the error would be such as to increase the value of C_x above its true value. The geometry of the specimen does not appear to be particularly critical and special molding and machining techniques are not required.

(d) Electrode edge effects have been studied extensively [5]. With the specimen in place the edge capacitance is higher than one assumes and hence the calculated value of C_x is too high. In general, this error increases with ϵ' and also with the proximity of the specimen to the edge of the electrode. If the distance between the specimen and the electrode edges is twice the separation of the electrodes then this error is <0.1 percent [5]. Thus it is advisable to design the holder accordingly keeping in mind that measurement sensitivity decreases as the ratio of electrode area to specimen area increases, i.e., as C_x becomes a smaller part of the total capacitance.

(e) Since ϵ' for air is about 0.06 percent higher than assumed, then the actual capacitance with specimen removed is higher than calculated, thus lowering ϵ' by <0.1 percent. The effect on the airgap is negligible.

A final approximation of the accuracy of the measurements can be obtained from an estimate of the accuracy of the measured quantities in the working equations. Generally, the calibration capacitances can be measured to the nearest 0.1 percent, the linear dimensions to the nearest micron and the area of the sample using a traveling microscope to 0.02 percent. Considering eq (11) for ϵ' the factors and their percent uncertainty are C_v, ± 0.3 percent; $[(C_{c_2} - C_{v_1} + \epsilon_0 A_s/t_s - C']$, ± 0.3 percent; the airgap correction term, ± 0.5 percent. In eqs (22), (23), (27), and (29) for ϵ'', the factors and their percent uncertainty are C_T, ± 0.2 percent; $\epsilon_0 A_s/t_s$, ± 0.3 percent; $\Delta\left(\frac{1}{Q}\right)$, ± 3 percent; $C_{Ho}/(C_T - C_s')$, ± 0.3 percent; $(D_i - D_o)$, $\pm(0.5\% + 0.00005)$ for GR 1615A; $(D_i - D_o)$, $\pm(2\% + 0.0005)$ for GR 716C; and the airgap correction ± 1 percent. For eq (37) the percent uncertainty of the terms are dt/df_M, ± 0.5 percent; dC_c/dt, ± 1 percent; and Δf_i and Δf_o, $\pm(0.2\% + 10\text{kHz})$.

The total percent uncertainty for ϵ' and ϵ'' is obtained from the square root of the sum of the squares of the uncertainties in the working equations. For ϵ' the percent uncertainty is ± 0.7 percent over the entire frequency range 10^{-2} to 6×10^8 Hz. For ϵ'' we have $\pm(1.5\% + 0.0005)$ for the ultra low frequency bridge, $\pm(1.8\% + 0.0005)$ for the GR 716C bridge, $\pm(0.6\% + 0.0001)$ for the GR 1615A bridge, $\pm(4\% + 0.0005)$ for the 260A and 190A Q-meters, and $\pm(1.6\% + 0.0005)$ for the self-resonant measurements.

7. References

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