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REFINEMENTS IN PRECISION KILOVOLT PULSE MEASUREMENTS*

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ABSTRACT

This paper describes techniques for reducing errors encountered in measuring the amplitude of 100-300 kV pulses which are a few microseconds in length. The accuracy to which such measurements can be made depends, for the most part, on how precisely the behavior of the voltage dividing network is known. Problems due to stray reactances, temperature, voltage effects, dielectric and dimensional instabilities, losses, improper terminations and external circuitry are dealt with, with particular emphasis on capacitive voltage dividers. Also described briefly are an ultra stable laboratory standard divider, calibration techniques, and measuring instrumentation.

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I. INTRODUCTION

The fundamental techniques used today for measuring high voltage pulses to be delivered to high power radio frequency tubes were established two decades ago. In general, these techniques have served adequately, although those who have worked with high power pulse modulators have had to overcome certain problems in order to improve the accuracy of the pulse measurement. We find, however, that remarkably little progress has been made in some areas and some special programs now demand accuracies exceeding the present state of the art. This paper presents some of the needed refinements in the accurate measurement of pulses from 100-300 kV, which are a few microseconds in length and are delivered by line type pulse modulators. The techniques described may apply as well to hard tube pulsers with similar pulse specifications.

Stimulated by the discussions last year at the High Pulse Voltage Seminar at the National Bureau of Standards in Washington, D. C., and by our laboratory's needs, these measurement problems have been investigated in an effort to extend the accuracy to ± 0.1 percent or better. The purpose of this paper is to examine known sources of error in the measurement of high pulse voltages to determine more precisely what the kilovolt really is. The sources of error will be examined quantitatively where possible.

The standard approach to the high voltage pulse measurement problem is to reduce the amplitude of the pulse while still retaining the initial character so that it can be measured accurately by conventional low voltage instruments of reasonably well known precision. This reduction is accomplished with a voltage-dividing network which usually has negligible loading effect of the pulse modulator output; it should be carefully designed taking into consideration voltage and temperature effects, stability with time and transient response over a wide range of frequencies.

The present state of the art allows the measurement of short, high voltage pulses to accuracies of from 1 to 3 percent.[1] Most of the uncertainty in these measurements lies in the inability to predict the exact response of the dividing network to the high voltage pulse.

II. TYPES OF HIGH VOLTAGE PULSE DIVIDING NETWORKS

The basic pulse voltage dividing network is the RC divider shown in Fig. 1a. Special cases of this general form are the pure resistive and the pure capacitive dividers. The former is used primarily for measuring dc and low frequency and the latter for microsecond pulse measurements.

The resistive divider high frequency response is usually quite poor due to the distributed capacity within the resistor, stray capacity to other parts of the circuit, inherent inductance and the shunting effect of the viewing cable; all resulting in pulse waveform distortion significantly affecting a precision measurement. The effects of distributed and stray capacity are less significant in the low resistance divider but it is limited to applications where the additional power absorbed does not adversely affect the pulse modulator performance. Care must be taken here to insure that the temperature coefficients of the resistors are uniform and the power handling ratings are conservative. Another advantage is that the bottomside resistor can be made equal to the characteristic impedance of the viewing cable where termination problems and distortion are minimum.

Many of the above disadvantages are minimized in the RC divider. In theory, the time constants R_1C_1 and R_2C_2 of each section, when equal, result in a uniform response at all frequencies. In practice, however, the distributed capacity in the topside resistor is difficult to compensate for so that the result is equivalent to a section which has the same RC time constant as the bottom section. As in other dividers, the topside resistor must be well shielded to eliminate stray capacity to other parts of the circuit.

The pure capacity divider is probably the most popular for short, high voltage pulses[1] and the emphasis of this paper will be on this device. The schematic is shown in Fig. 1b. The problems of distributed capacity, stray capacity and power absorption are virtually eliminated. The capacity divider can be designed to work at very high voltages, occupy relatively little space, and can be made stable and relatively unaffected by its environment, making it a suitable choice for a laboratory standard for dividing down pulse voltages.

The main disadvantages are that its very low frequency response drops off and that since it is virtually a pure reactance, it tends to form a resonant circuit with the inductance of connecting leads which can result in high frequency ringing on the leading portion of fast rising voltage pulses.[2]

The voltage division ratio of the pure capacity voltage divider may be anywhere from 500:1 to 10,000:1. This demands that the topside capacitor withstand virtually the full voltage of the pulse and be small enough to have negligible reactive loading effect on the pulse modulator output. Vacuum capacitors are suitable for voltages up to 100 kV or so, but problems with field emission have been experienced and to eliminate them would perhaps necessitate an unusually large design physically. High voltage capacitors using solid insulators such as epoxy as a dielectric material have also been used. Perhaps most common for use above 100 kV is the divider which uses transformer oil for the dielectric in the topside capacitor which typically has a value of 1 to 10 picofarads. Usually the bottom or low voltage capacitor is a silver-mica type having a value of 0.001 to 0.05 microfarads.

The earlier designs of this type were quite arbitrary in the choice of electrode geometry for the topside capacitor emphasizing primarily voltage breakdown strength and careful shielding from stray capacity. The coaxial capacity divider using guard rings as shown in Fig. 2 was suggested by Dedrick[3] and perhaps others a decade or so ago. This divider has two main advantages over the earlier type. First, its topside capacity could be calculated and built easily to within 1 percent.[4] Second, its capacitance value is virtually insensitive to geometrical misalignment or electrode deformation.[3,4,5]

The most common problem with the oil type capacity dividers is the sensitivity of the topside capacitor value to changes in oil temperature which may typically be 0.06 percent per degree centigrade, making them normally unsatisfactory for precision high voltage measurements where the oil temperature is likely to vary.

III. DEVELOPMENT OF A LABORATORY STANDARD

An ultra-stable capacity divider usable to 300 kV was designed and built at Stanford by Brady and Dedrick[3,4,5] in 1960. Extreme care was taken in its design to make the dividion ratio essentially independent of voltage, temperature, position and time over a wide band of frequencies. The temperature independence was achieved by providing uniform properties in both capacitors. Dow-Corning 200 silicone electrical grade oil is used for the dielectric in both C_1 and C_2 and brass electrodes are used throughout. A cutaway section is shown in Fig. 3.

The division ratio remains unchanged due to changes in physical dimensions caused by temperature variations, since each electrode capacity change with temperature goes as the expansion coefficient of brass. Since the same oil is used for the dielectric in each capacitor, the division ratio remains unchanged with changes in dielectric constant due to temperature. The topside capacitor is a coaxial configuration which has a radii ratio of e which, it can be shown, provides the maximum ratio of capacity to electric field. This obviously allows the optimum package size for a given capacity and voltage rating. The bottomside capacitor is made up of 50 annular rings separated by 0.050 inch. The capacities of the top and bottom side capacitors are nominally 8 pf and 8000 pf respectively, giving a division ratio of approximately 1000:1. Errors due to ellipticity or lack of concentricity of circular electrodes are second order, but nevertheless considered and known.[5]

IV. PRECISION CALIBRATION

After the high pulse voltage seminar mentioned earlier it was decided to use the Brady capacitive divider as a laboratory standard at SLAC after a suitable calibration was performed. It was of course desirable to calibrate this standard as well as other dividers under pulsed high voltage conditions.

At present the best certified high voltage pulse standard calibration service is at the National Bureau of Standards. The Brady divider was calibrated at NBS at 20, 60 and 100 kV using a 12.5 microsecond pulse. The uncertainty in this calibration is $\pm$ 1.0%.

It was decided to make a precision calibration under low voltage conditions and then carefully examine the possible deviations which might exist when translating to the high voltage pulse conditions. One calibration method used was to carefully measure the ratio of capacity divider impedances at 1000 Hz on a precision bridge using a cascaded pair of ratio transformers; each having a division accuracy of 1 part in 10^6. The bridge circuit is shown in Fig. 4. The outer shield and one terminal of the bottomside capacitor of the divider being calibrated are common and normally at ground potential. It can be seen from the bridge circuit schematic that this ground is incompatible with the bridge ground thereby making it necessary to "float" the divider being calibrated inside a shielded cubicle.

Without going into the mathematics of the bridge balancing equations, it should be pointed out that corrections must be made for stray capacities within the bridge and for lead inductances where critical. Precision phase balance in the bridge is accomplished with the decade resistance box in series with the divider. With care and appropriate corrections this calibration can be made to better than 50 ppm. However, this calibration applies necessarily at only the voltage, temperature and frequency at which it is made.

A second calibration is made using the bridge circuit in Fig. 5. This method lacks the precision of the first method but does have two advantages. First, it is less unwieldy and can be used to calibrate dividers while they are in place in the pulse transformer tank since the shielded cubicle is not required. Secondly, that the calibration may be performed at any frequency up to 100 kHz while the ratio transformer bridge accuracy drops off above 1 kHz. This circuit is similar to a Schering bridge except that all arms of this bridge are primarily capacities. Two General Radio precision standard capacitors, one fixed and one variable, are balanced with the capacity divider undergoing calibration. Care must be taken to minimize the inductance of the leads connecting the bridge components where coax is not used. The inductance inherent in the components must be known and corrected for. Proper phase balance is achieved by adjusting R_n, the diode resistance box, along with the variable capacitor to achieve a good null. The topside precision standard capacitor is a GR 1422-CD precision variable capacitor which has two sections which can be set from 0.05 to 1.10 picofarads and from 0.5 to 11.0 picofarads, respectively. The bottomside standard capacitor is one of the GR-1402 series. Three different values have been used, .001, .005 or .01 $\mu\mu$f depending on the division ratio of the divider being calibrated. The value is chosen which allows the variable topside capacitor to operate near its full-scale setting for minimum uncertainty. For greater accuracy, it is planned to have both standard capacitors fixed and one of them "trimmed" to balance the bridge with a precision variable capacitor. At present, the largest single source of error with the present capacitor bridge is the uncertainty in the variable capacitor setting which is about 0.1%.[6]

At first this bridge was used for calibrations without the quadrature balancing resistor R_n and the null was quite broad and difficult to set, thereby limiting the repeatability to about 0.15%. This is due primarily to the finite resistance to each input of the differential amplifier rather than to the dissipation factor loss in any of the capacitors. The exception of course occurs when a symmetrical situation exists when $R_2C_2 = R_4C_4$; eliminating the need for R_n.

V. BEHAVIOR UNDER HIGH VOLTAGE PULSE CONDITIONS

A capacity divider can be calibrated at a particular voltage, temperature and frequency. The division ratio under these conditions can be known to within $\pm$ 0.01%. Let us now examine many of the possible deviations from the measured division ratio which may exist under pulsed high voltage conditions.

A. Temperature and Dimensional Stability

Small changes in the geometry of a capacity voltage divider will normally occur with environmental temperature changes due to thermal expansion of the electrodes, insulator supports, etc. This discussion will be confined to capacity dividers which have a coaxial configuration for the topside capacitor since most of the capacity voltage dividers at SLAC are of this type. For a divider which has guard electrodes the capacity is given by[4]

$$C_1 = \frac{2\pi \epsilon_r \epsilon_o \ell}{\ln(b/a)} \tag{1}$$

where ϵ_r is the relative dielectric constant between the inner and outer electrodes; ϵ_o is the permittivity of free space; a and b are the radii of the inner and outer electrodes respectively and ℓ is the effective length of the outer electrode which is sometimes called the signal ring. This length includes the effect of the fringing fields in the vicinity of the gap next to each guard electrode. If the electrodes are made of the same material, then the radii ratio will remain constant with thermal expansion and the capacity can be expressed as

$$C_1(T) = C_{1_o} (1 + \rho\Delta T) (1 + \alpha\Delta T) \tag{2}$$

where α is the thermal coefficient of expansion of the electrode material and ρ is the temperature coefficient of the dielectric medium. The change in capacity with temperature is then given approximately by

$$\frac{\Delta C_1(T)}{C_{1_o}} = (\rho + \alpha) \Delta T + 2\rho \alpha\Delta T \tag{3}$$

For a vacuum coaxial capacitor $\rho = 0$ and the capacity is affected only by expansion in the logitudinal dimension. A capacitor with transformer oil between the electrodes is primarily governed by ρ rather than α . The ρ term varies with type and manufacturer but is typically -3×10^{-4} to -1×10^{-3} per degree centigrade. α is typically 2×10^{-5} per degree centigrade for brass or aluminum.

The changes in the bottomside capacitor due to temperature changes may not be so easy to predict. If the bottomside capacitor is the silver-mica type, the capacity temperature coefficient may be as high as $\pm 3 \times 10^{-4}$ per degree centigrade but is usually unspecified. This would govern the ratio of a divider with a vacuum capacitor but would be relatively insignificant compared with the oil temperature coefficient in an oil dielectric capacitor. This points out one obvious advantage of the vacuum capacity divider over the oil capacity divider.

In the stable laboratory standard (Fig. 3) the capacity of the bottomside capacitor is approximately

$$C_2 = \frac{N \epsilon_r \epsilon_o A}{d} \tag{4}$$

where A is the effective area of the plates, d is the spacing per gap and N is the number of gaps. The area of course goes with temperature as[5]

$$A(T) = A_o (1 + \alpha \Delta T)^2 \tag{5}$$

and the gap goes as

$$d(T) = d_o (1 + \alpha \Delta T) \tag{6}$$

Therefore, the capacity of the bottomside capacitor goes with temperature as

$$C_2(T) = C_{2_o} (1 + \rho \Delta T) (1 + \alpha \Delta T) \tag{7}$$

and

$$\frac{\Delta C_2(T)}{C_{2_o}} = (\rho + \alpha) \Delta T + 2\rho\alpha \Delta T \tag{8}$$

making the ratio of capacities in the Brady divider independent of the oil and electrode temperature.[5]

The effect of unequal temperature coefficients is most evident in the voltage divider which uses transformer oil for the dielectric C_1 and a small paper or mica capacitor for C_2. This can cause errors in the pulse measurement of 2 to 4% since 40 degree oil temperature variations are not uncommon.

The effect of temperature on a commercial capacity divider of this type was measured using a variation of the precision capacitor bridge described earlier and shown in Fig. 6. The divider was placed in transformer oil which was circulated and carefully temperature controlled from 20 to 80 degrees centigrade while the division ratio was measured as a function of temperature. The silver-mica bottomside capacitor was then removed and replaced by a General Radio precision standard capacitor of approximately the same value, but which was placed in a constant temperature environment while the test was repeated, varying only the temperature of the topside capacitor. The results indicated that the topside capacitor (oil dielectric) accounts for most of the division ratio change with temperature. The entire divider has a temperature coefficient of 556 ppm/$^{\circ}$C, while the topside capacitor has a coefficient of 614 ppm/$^{\circ}$C indicating that the temperature coefficient for C_2 was much smaller and opposite in sign to that for C_1.

The accuracy of high voltage pulse measurements, using the commercial divider which has an oil-coaxial topside capacitor and a silver-mica bottomside capacitor under varying temperature conditions, can be improved considerably by a temperature compensating technique. Capacitors with unusually large temperature coefficients are available and can be used to "trim" the bottomside capacitor in such a way as to allow nearly identical capacity changes with temperature, thereby keeping a fairly uniform division ratio. To compensate the divider in this manner, the trimming capacitor must be

$$C_t \approx \frac{(\alpha_2 - \alpha_1)}{(\alpha_1 - \alpha_t)} C_2 \tag{9}$$

where α_n is the temperature coefficient of C_n. The division ratio will be altered, of course, and the new division ratio becomes

$$K = \frac{C_1}{C_1 + C_2'} \tag{10}$$

where $C_2' = C_2 + C_t$. It should be mentioned at this point also that this compensation relationship does not hold where the viewing cable capacity is significant. As discussed later, the quasi steady state division ratio is

$$K = \frac{C_1}{C_1 + C_2 + C_c} \tag{11}$$

where C_c is the capacity of the viewing cable. If the cable temperature is assumed to remain relatively constant the temperature term α_2 in Eq. (9) must be replaced by α_2' which is given by

$$\alpha_2' = \frac{C_2}{C_2 + C_c} \alpha_2 \tag{12}$$

and C_2 in Eq. (11) must be replaced by C_2' to express the division ratio of the temperature compensated divider correctly.

Figure (6) shows the effect of compensation on the divider on which the temperature coefficient measurements were made. Before compensation the division ratio changed approximately 3% over a 60° C temperature range. After compensation the division ratio deviation was about $\pm$ 0.1% over the same temperature interval. The trimming capacitor C_t had an average temperature coefficient of approximately $-5 \times 10^{-3} / ^{\circ}C$. There is an error in the capacity divider which has been temperature compensated without allowing for the cable, or in the divider which has equal temperature coefficients for C_1 and C_2 such as in the Brady Standard. The error due to temperature with the cable present is given by

$$\frac{\Delta K}{K_o}(T) \approx - \frac{\alpha_1 C_c \Delta T}{C_1 + C_2 + C_c} . \tag{13}$$

This effect obviously can be minimized by keeping the ratio C_c/C_2 as small as possible. For a 20 foot length of RG58A/U coax and a change in oil temperature of $20^{\circ}C$, the change in division ratio of the Brady Standard is $\sim 0.1\%$.

One unique advantage of the coaxial capacitor, which has guard rings in addition to the signal ring, is that its capacity is relatively insensitive to misalignment of the two electrodes compared with the parallel plate geometry.[4] The most likely misalignment problem with a coaxial capacitor would be skewed conductor axis. This problem, being too difficult to solve analytically, can be solved intuitively by examining the case where the two electrodes are not concentric. Referring to Fig. (7a) it can be shown[4] that small deviations from concentricity the change in capacity is approximately

$$\frac{\Delta C}{C} \approx \left(\frac{\delta}{a}\right)^2 \left[\left(\sigma^2 - 1\right) \ln\sigma\right]^{-1} \tag{14}$$

where $\sigma = b/a$. For a coaxial capacitor which has $\sigma = e$ a 5% misalignment results in a capacity change of only 0.04%.

Another dimensional change which might occur is warping of the outer conductor or signal electrode. Referring to Fig. 7b, it can be shown[4] that the change in capacity of a coaxial condenser, whose outer electrode has been distorted into an ellipse is given approximately by

$$\frac{\Delta C}{C} \approx \frac{(h/b)^2}{\ell n \sigma} \left[\frac{5}{4} + \frac{2}{\sigma 4} \right] \tag{15}$$

where h is the deformation, and $\sigma = b/a$. For a deformation of 1% where $\sigma = e$, the change in capacity is 0.012%.

B. Frequency Response

The data supplied by transformer oil manufacturers indicates that the dielectric constant of most oils is relatively independent of frequency between[7] 20 and 100°C although exact figures over the bandwidth of interest are not readily available. Bridge measurements in our laboratory at room temperature indicate that the change in capacity from 1 KHz to 100 KHz is less than 1 part in 10^3.

Because the capacity divider is a pure reactance it tends to form a resonant circuit with the length of wire connecting the topside capacitor to the high voltage terminal.[2] With a moderate amount of care this does not present a problem unless the pulse to be measured has an extremely short rise time. A 10 inch length of large diameter wire, say 1/8 inch diameter copper tubing has a self inductance of about 0.3 μh, having a reactance at 1 MHz of about 2 ohms compared with a total reactance of a capacity divider with a 5 pf topside capacitor of about 30 kilohms at the same frequency.

The frequency response of three different types of coaxial dividers was measured in two ways. First, the division ratio was carefully measured on the capacitance bridge as a function of frequency from 1 KHz to 100 KHz. At the higher frequencies the residual inductances of the precision standard capacitors and the dielectric changes have a small effect that is known and is corrected for. Second, the response of each divider was then measured from 50 KHz to 50 MHz using a constant amplitude sine wave generator and two rf voltmeters; one to insure that the input voltage to the high voltage terminal remained constant and the other to monitor the output of the divider. All three dividers were flat to within 0.2% from 1 KHz to 100 KHz. Above 1 MHz each exhibited similar resonances as shown in the response curves of two of these dividers in Fig. 8. Some resonances were observed which turned out to be caused by harmonics in the signal generator and were disregarded and are not shown on the response curves. The dip occurs at about 13 megahertz on the Brady Divider and at about 23 megahertz on a commercial divider. Because of the relatively large values of C_2, the bottomside capacitors in each case, it would only take a few hundredths of a microhenry to form a series resonant circuit to cause the dip at these frequencies. The peak which occurs at the higher frequency can have high Q's but are not serious for high voltage pulses with normal rise times.

At this writing it is not clear exactly what the equivalent circuit inductance values are or how they are distributed although it is generally believed that the first dip is due to the series "self resonance" in the bottomside capacitor. It is not apparent why the dip occurs at such a low frequency in the Brady divider since C_2 has very low residual inductance by virtue of its design.

C. Voltage Effects

A 1000 ohm non-inductive resistive divider/dummy load combination was built; one purpose being to investigate possible voltage effects on the capacity divider pulse voltage measurement. It was hoped that a resistance divider of this type could be built with uniform voltage and temperature coefficients and suitable frequency response which could handle 50 MW peak power, 12 kW average power with adequate oil cooling. Unfortunately, the inductance of the 0.2 ohm bottomside resistor affects the frequency response to the point where a precision measurement cannot be made for comparison with the capacitive divider measurement. The resistive divider at NBS in Washington, D.C., is probably the best suited for this investigation. At this writing, the authors are attempting to improve the response so this comparison can be made. It has been assumed that the voltage coefficient of the divider oil is linear and the ratio of capacities in the Brady Standard is independent of voltage.

D. The Viewing Cable

Quantitative information on the effect of the viewing cable on the precision high voltage pulse measurement is for the most part nonexistant except for the effect of the added capacity presented to the system by the cable. This has the obvious effect of increasing the effective division ratio of the divider and it is a simple matter to include the cable as part of the divider when making bridge measurements of its ratio and to include its known capacity in division ratio calculations.

There are other effects that should be mentioned. It is generally known that when the divider is purely capacitive, the output end of the viewing cable cannot be terminated in its own characteristic impedance without differentiating the pulse beyond recognition. The universally accepted method around this problem is to insert a series termination $R_m = Z_o$ between the input end of the cable and the capacity divider output as shown in Fig. 1b. It is assumed that the input impedance to the oscilloscope is essentially an open circuit. At the higher frequencies the reactance of the bottomside capacitor in the divider is sufficiently low so that most of the pulse that is reflected from the open circuit end of the transmission line is absorbed by R_m. It has been found that the common 1/2 watt carbon resistor makes an adequately non-inductive termination.

There is a problem of transient cable loading, which is generally overlooked and accurs when the length of the transmission line is long enough so that the transit time of the signal is a significant fraction of the pulse width or the capacity of the cable is an appreciable portion of the bottomside capacitor in the capacity voltage dividing network. When either or both of these conditions exist, waveform distortion occurs which cannot be over looked if an accurate measurement is desired.

When the high voltage pulse is applied to the capacity voltage divider, C_1 and C_2 assume their appropriate voltages. However, the impedance seen across C_2 initially is $R_m + Z_o$. As the divided voltage pulse enters the cable, it is further divided by $Z_o/(R_m + Z_o)$ and as the signal propagates along the cable, the voltage across C_2 drops with a time constant $(R_m + Z_o)(C_1 + C_2)$. Assuming that the load at the viewing end of the cable has essentially infinite impedance, the voltage doubles at the scope and the wave is reflected essentially unchanged back toward the capacity voltage divider arriving at such time that the voltage across C_2 has dropped to

$$V_2 = \frac{C_1}{C_1 + C_2} \quad V_o e^{-2\ell/v_p(R_m + Z_o)(C_1 + C_2)} , \tag{16}$$

where v_p is the velocity of propagation and ℓ is the cable length. It is assumed the $2\ell/v_p$ is less than the pulse width T_o, that $R_m = Z_o$ and that the transmission line is ideal. Therefore

$$C_c \equiv \frac{\ell}{v_p Z_o} = \frac{T_1}{v_p Z_o} \tag{17}$$

where C_c is the capacity of the cable and T_1 is the one-way cable transit time. If $C_c/C_2 \ll 1$ then after one down and back transit or cable filling time, $2T_1$,

$$V_2 \approx V_3 \approx \frac{C_1}{C_1 + C_2 + C_c} V_o \quad . \tag{18}$$

This is the familiar expression for the division ratio of a capacity divider which has significant cable capacity added to it. It is however only an approximation during the transient filling time of the cable and exactly expresses the divided down voltage value after several reflections have occurred within the cable. It is important that the measuring point be made far enough from the leading edge so that a quasi-steady state level is reached when the cable is uniformly charged to the same potential as the final voltage across the bottomside capacitor C_2.

Figure 9 shows the measured effect of transient cable loading on a 225 kV pulse which was divided nominally 5000:1 with a divider having $C_1 = 1.2$ pf and $C_2 = 6000$ pf. The pulse has a rise time of 0.5 microseconds. It is seen that initially the pulse is divided down by $C_1/(C_1 + C_2)$. Each reflection will contribute to the charge adjustment along the cable; the n^{th} reflection having the form

$$V_3 \Big|_{n^{th}\Gamma} \sim \frac{(t - 2nT_1)^n}{\tau^n n!} \quad e^{-(t - n2T_1)/\tau} \quad . \tag{19}$$

As the number of reflections approaches infinity, Eq. (18) becomes exact. The cable lengths are abnormally long merely to illustrate this effect. It is important to remember, however, that the early part of the pulse will always be distorted by the transient loading effect. The extent depends on the cable length and for a precision measurement a selected point at least 10 transit times from the leading edge of the pulse should be sufficient. The distortion caused by cable high frequency dispersion or dielectric loss is difficult to predict. For frequencies which primarily contribute to the flat portion of the pulse the effect is considered negligible.

Another form of distortion is caused by the finite resistance shunting the viewing end of the cable resulting in pulse droop. This causes the voltage across C_2 and C_c to droop with a time constant $R_i(C_2 + C_c + C_i)$, where R_i and C_i are the input resistance and capacity to the scope. The error is greater the further the measurement is made from the leading edge. For the other reasons discussed it is generally still desirable to choose a point on the pulse for the measurement well away from the leading edge since the high frequency errors are not so well known. The error due to the droop is given approximately by

$$\frac{\Delta V}{V} \approx - \frac{T_s}{R_i\,(C_2 + C_c + C_i)} \tag{20}$$

where T_s is the time from leading edge selected for the measurement. For this approximation an ideal step function is assumed. For the capacity divider mentioned above and an oscilloscope input resistance of 1 megohm, this error is about 0.017% per microsecond referred to the leading edge.

VI. MEASUREMENT OF THE REDUCED VOLTAGE PULSE

Once the high voltage pulse has been reduced and passed by the dividing network the uncertainty in the measurement is typically an order of magnitude better than that of the divider itself. Divided pulse voltages may be measured with pulse voltmeters of either the slideback or direct peak reading type. For precision measurements however there are intrinsic drawbacks to these pulse voltmeters, and oscilloscope voltage comparator methods are preferred. Here the divided down pulse amplitude is compared with a dc voltage and fed into the high gain differential amplifier and the difference displayed on an oscilloscope. Such a circuit for oscilloscope presentation is provided in the Tektronix types "Z" or "W" oscilloscope preamplifiers. Care must be taken to insure that the equipment has been carefully calibrated so that the comparator voltage is accurately known. This depends on the comparator voltage power supply stability and the uncertainty in the comparator voltage dividing potentiometer. Also contributing to the uncertainty are the limited resolution of the video display and the common mode rejection capability of the differential amplifier. When using a good voltage comparator differential amplifier oscilloscope the resolution uncertainty is typically 0.01% for measuring a pulse 50 volts in amplitude. The common mode rejection capability in the Tektronix unit is about 3 parts in 10^5. This particular unit is known to experience severe instability when the rate of rise of the voltage pulse exceeds 6 volts per nanosecond. Care should be taken to insure a true response of the amplifier to the waveform to be measured, and to insure that its leading edge does not have a rise time approaching the critical specification. There is some uncertainty in the input attenuator to the differential amplifier. Where possible, the reduced pulse should be fed directly into the comparator circuit, i.e., when the input attenuator is in the zero attenuation position thereby eliminating any uncertainty in the attenuator divider ratio.

For more precise measurements the "Z" or "W" preamplifier is used only as a differential comparator whereby an external dc comparator voltage monitored by a precision dc voltmeter is compared with the pulse voltage. Similar precautions must be taken with this method also, but the uncertainty in the comparator voltage is limited only to the uncertainty in the external precision voltmeter. The overall uncertainty in this measurement, apart from that in the dividing network, is determined by the resolution of the video display, the common mode rejection capability of the differential amplifier and the accuracy to which the external comparator voltage can be measured with the precision dc meter. This can be made typically to 0.05% using a Fluke differential voltmeter.

When circumstances require rapid, routing pulse voltage measurements, oscilloscope methods may be impractical. At this facility where data must be taken on 246 klystron modulator systems which are operating simultaneously, the pulse voltmeter has the advantage that it can be made portable or that it may provide a suitable analog signal for feeding operating conditions into a computer; but at the expense of at least two orders of magnitude in accuracy.

Peak reading pulse voltmeters, in addition to normal meter errors, are subject to errors introduced when the duty cycle is short. For the simplified peak-above-zero circuit in Fig. 10 the error due to the short duty cycle is given approximately by

$$\frac{\Delta V}{V} \approx \left(1 + \frac{DR_m}{R_d}\right)^{-1} \quad , \tag{21}$$

where R_d is the equivalent resistance of the peak reading diode, R_m is the resistance of the metering circuit and D is the duty cycle. It is therefore desirable to have as high impedance metering circuit as possible. It is interesting to note that the value of the peak reading capacitor, C_p, does not affect the division ratio when loading a capacity divider once a steady state condition is reached.

In addition to the intrinsic duty cycle dependent error, there is an error due to the shunt capacity of the peak reading diode and is given approximately by

$$\frac{\Delta V}{V} \approx C_d/C_p \tag{22}$$

and also by irregularities in the pulse waveform which may exceed the portion of the pulse which is of primary interest. An example is a small amount of overshoot on the leading edge of an otherwise flat pulse waveform.

In view of these problems the pulse voltmeters should not warrant consideration for precision pulse measurements but rather should be limited to applications where quick voltage measurements of the order of 5-10% are adequate.

VII. CONCLUSION

Most of the error in a high voltage pulse measurement is due to the uncertainty in the voltage dividing network. Using the described capacitive divider standard, which has been calibrated on a precision ratio transformer bridge and the oscilloscope voltage comparator method, the overall uncertainty is about $\pm 0.2\%$.

This paper has been limited primarily to improvements in capacity divider techniques in order to accommodate our laboratory's needs. It is hoped that this work will encourage others to improve these methods and perhaps supply new techniques in the field of high voltage pulse measurements. X-ray hardness and electron momentum methods perhaps could be applied to precision high voltage measurements. Electrical breakdown in gases has also been suggested. Low voltage bridge techniques are presently adequate for their purpose but the need exists for a high voltage bridge for pulse divider standardization; extended perhaps to 1 MV.

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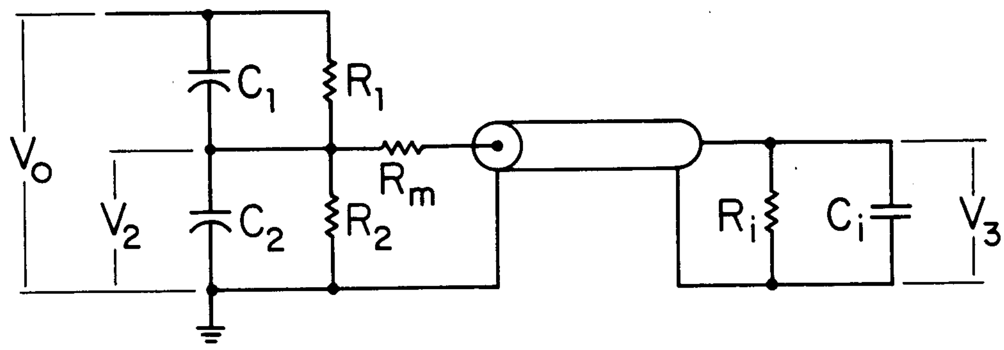
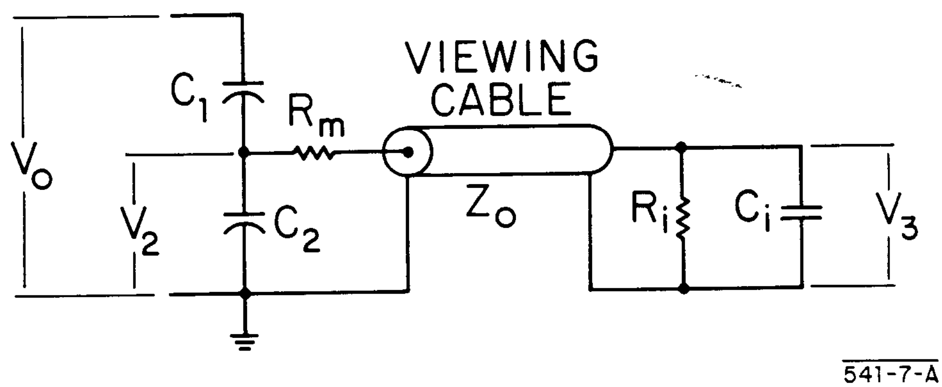
FIGURE CAPTIONS

1. Typical high voltage pulse divider networks.
2. Coaxial capacity divider with guard electrodes.
3. Brady capacity divider standard.
4. Ratio transformer bridge for 0.005% calibration.
5. Precision capacitor bridge for 0.2% calibration.
6. Temperature behavior of a commercial divider using oil and mica dielectrics for C_1 and C_2 respectively and the results of temperature compensation with a "trimming" capacitor.
7. Coaxial geometry irregularities.
8. Frequency response of capacitive dividers.
9. Transient cable loading effect when cable lengths are excessive. Γ_2 and Γ_3 are the reflection coefficients at each end of the viewing cable.
10. Typical peak reading circuit.

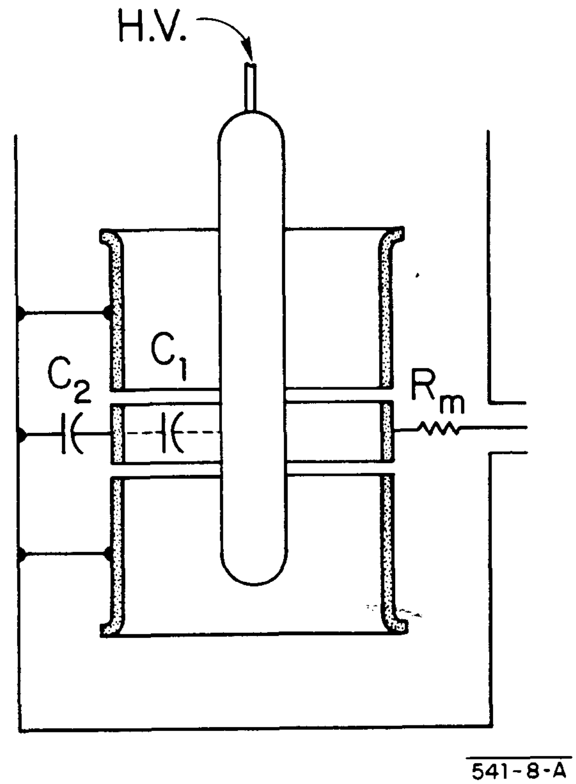


a. COMPENSATED RC DIVIDER



b. PURE CAPACITY DIVIDER

FIG. 1



COAXIAL CAPACITY DIVIDER

FIG. 2

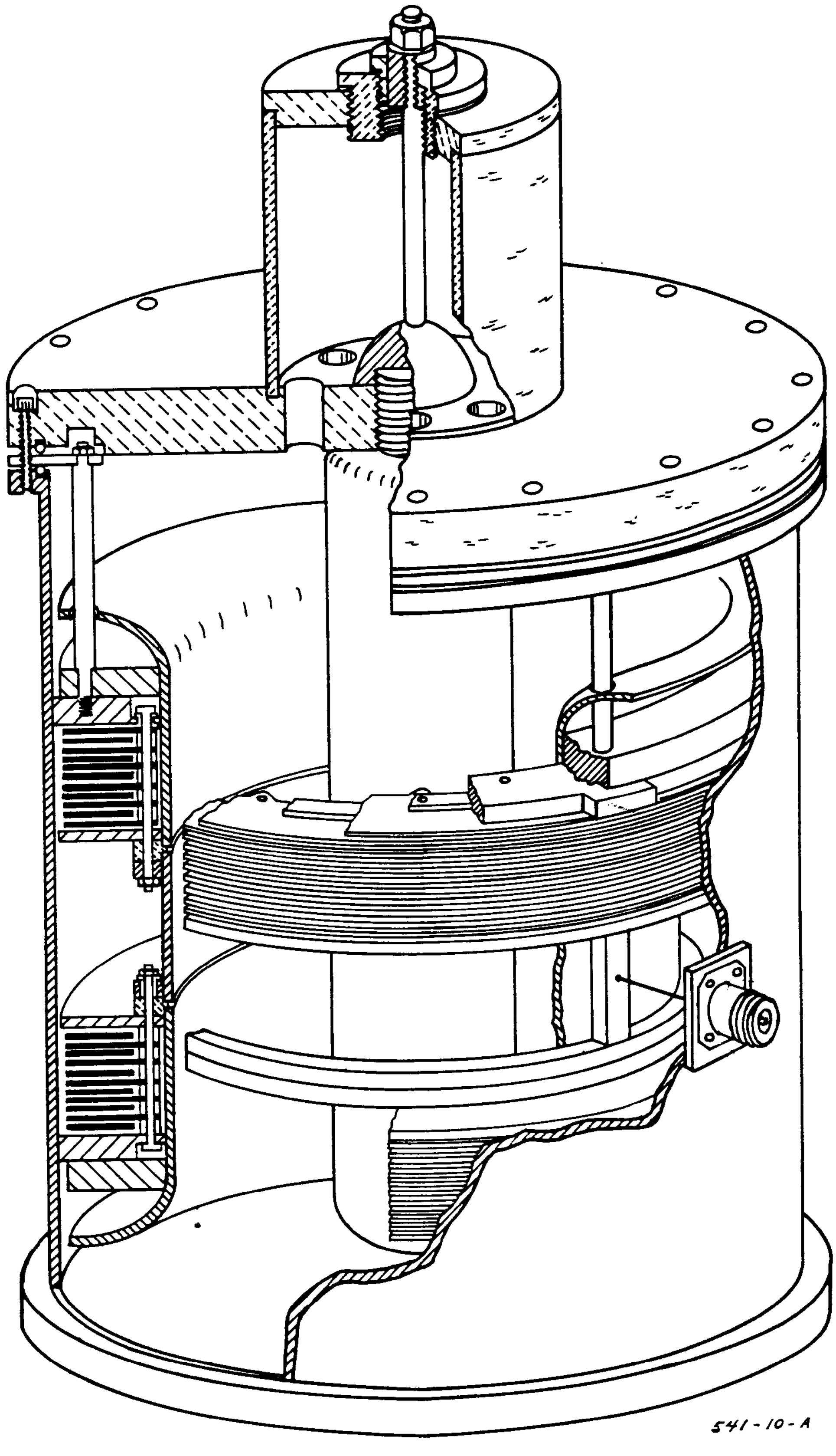


Fig. 3

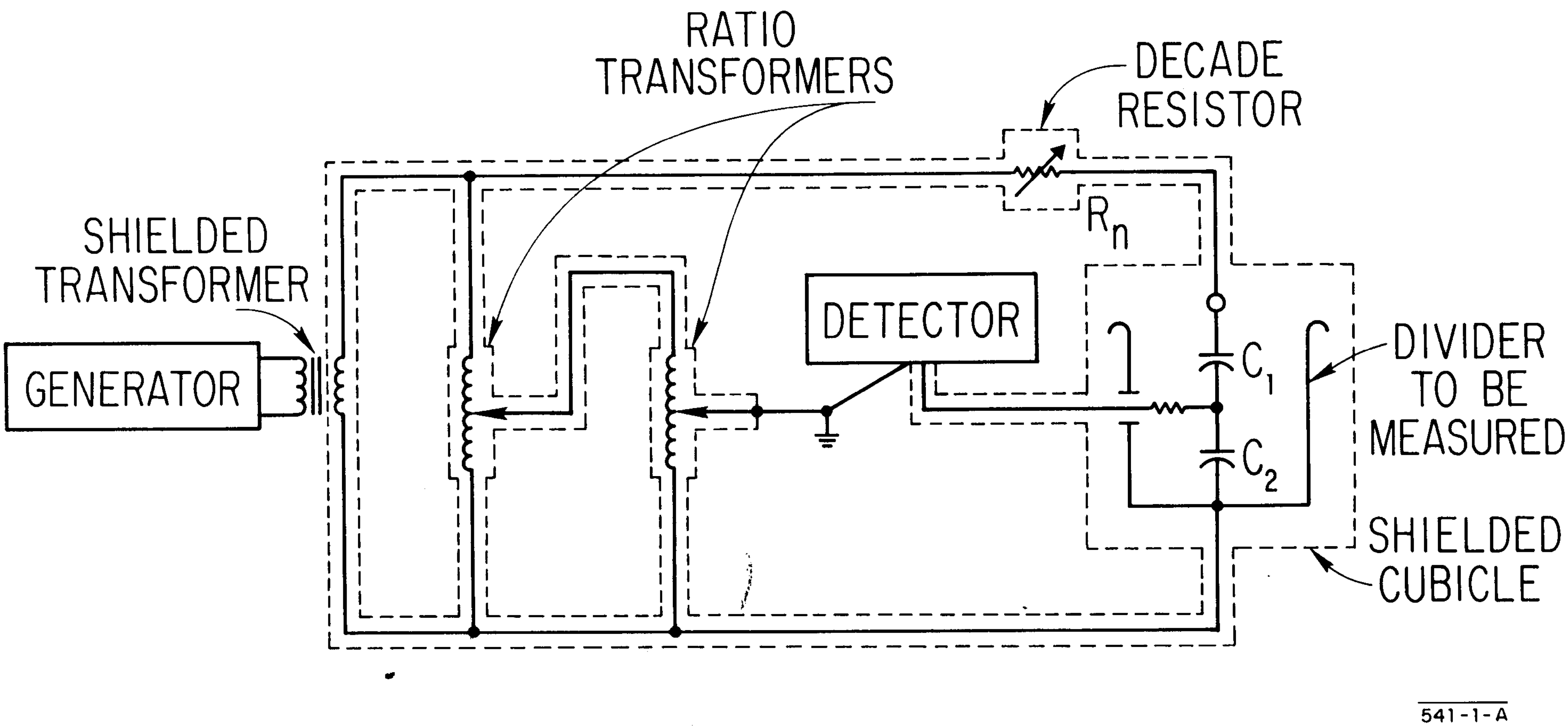


FIG. 4 - RATIO TRANSFORMER BRIDGE

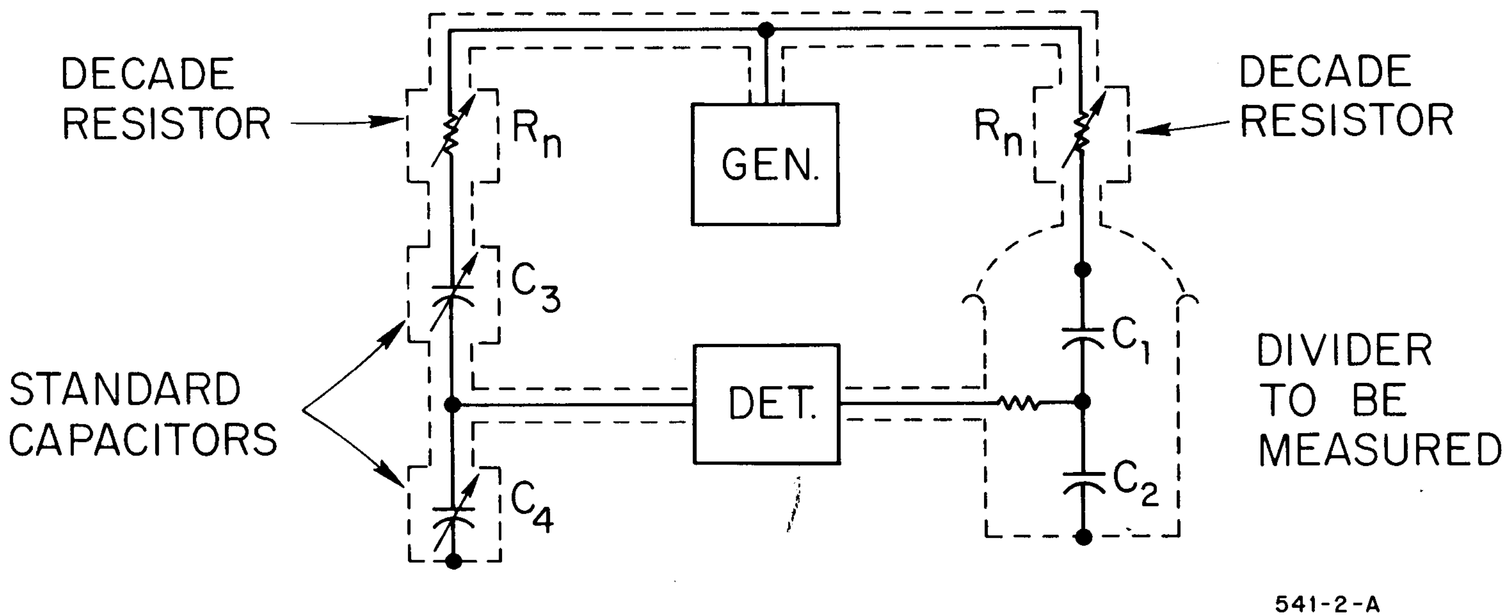
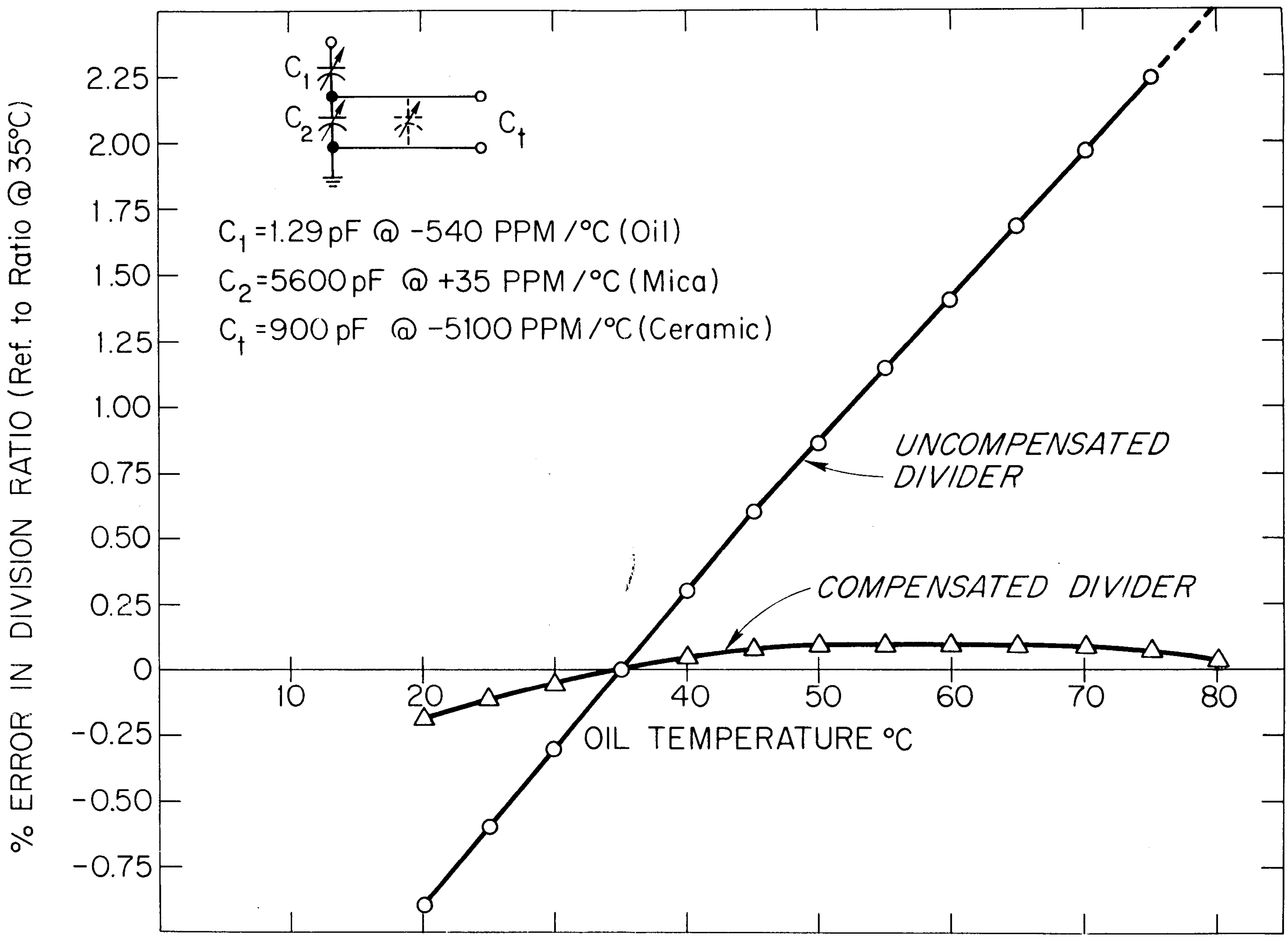


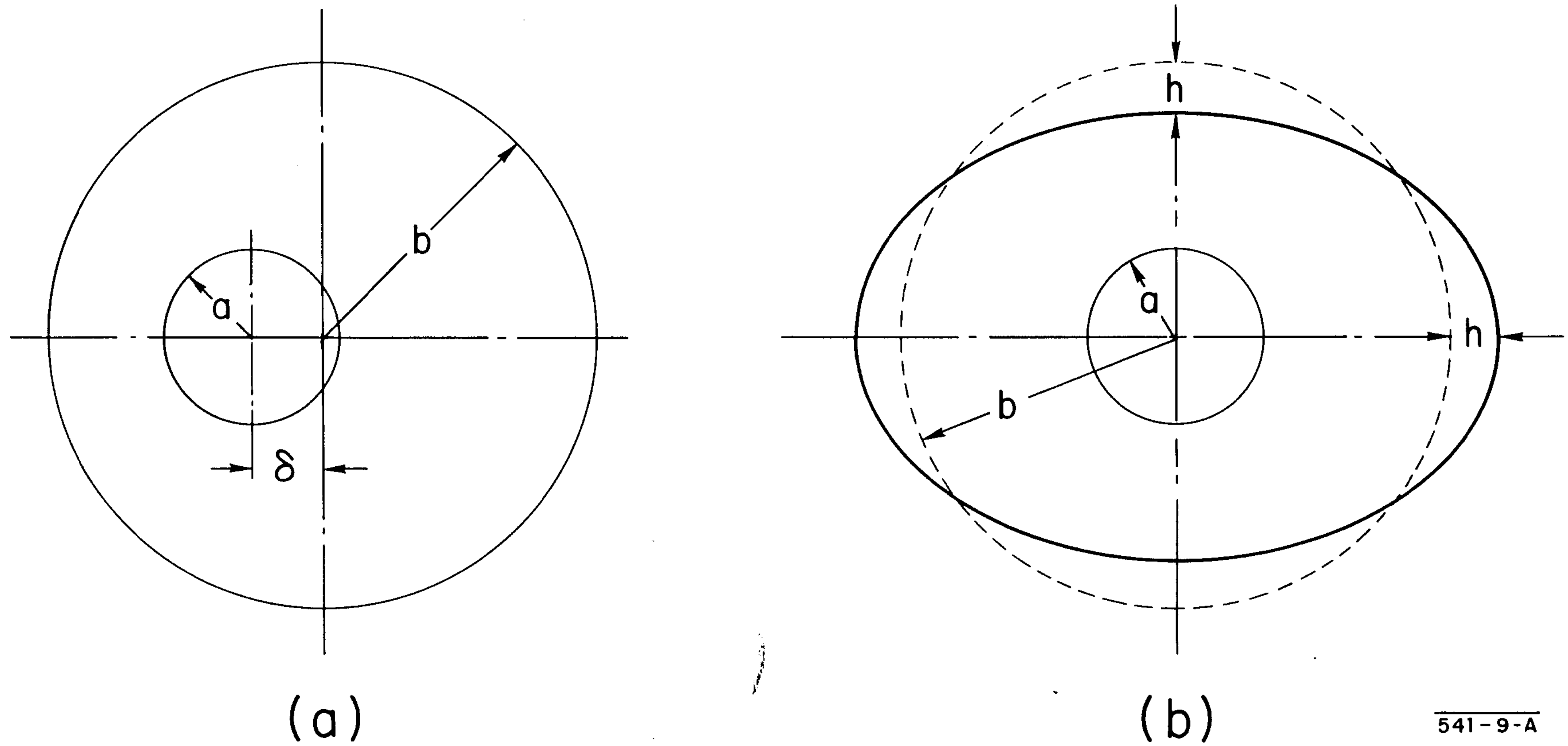
FIG. 5 - CAPACITY BRIDGE



C_1
C_2
C_t
C_1 =1.29 pF @ −540 PPM / °C (Oil)
C_2 =5600 pF @ +35 PPM / °C (Mica)
C_t =900 pF @ −5100 PPM / °C (Ceramic)
UNCOMPENSATED DIVIDER
COMPENSATED DIVIDER
% ERROR IN DIVISION RATIO (Ref. to Ratio @ 35°C)
OIL TEMPERATURE °C
2.25
2.00
1.75
1.50
1.25
1.00
0.75
0.50
0.25
0
−0.25
−0.50
−0.75
10
20
40
50
60
70
80

FIG. 6

541-3-A

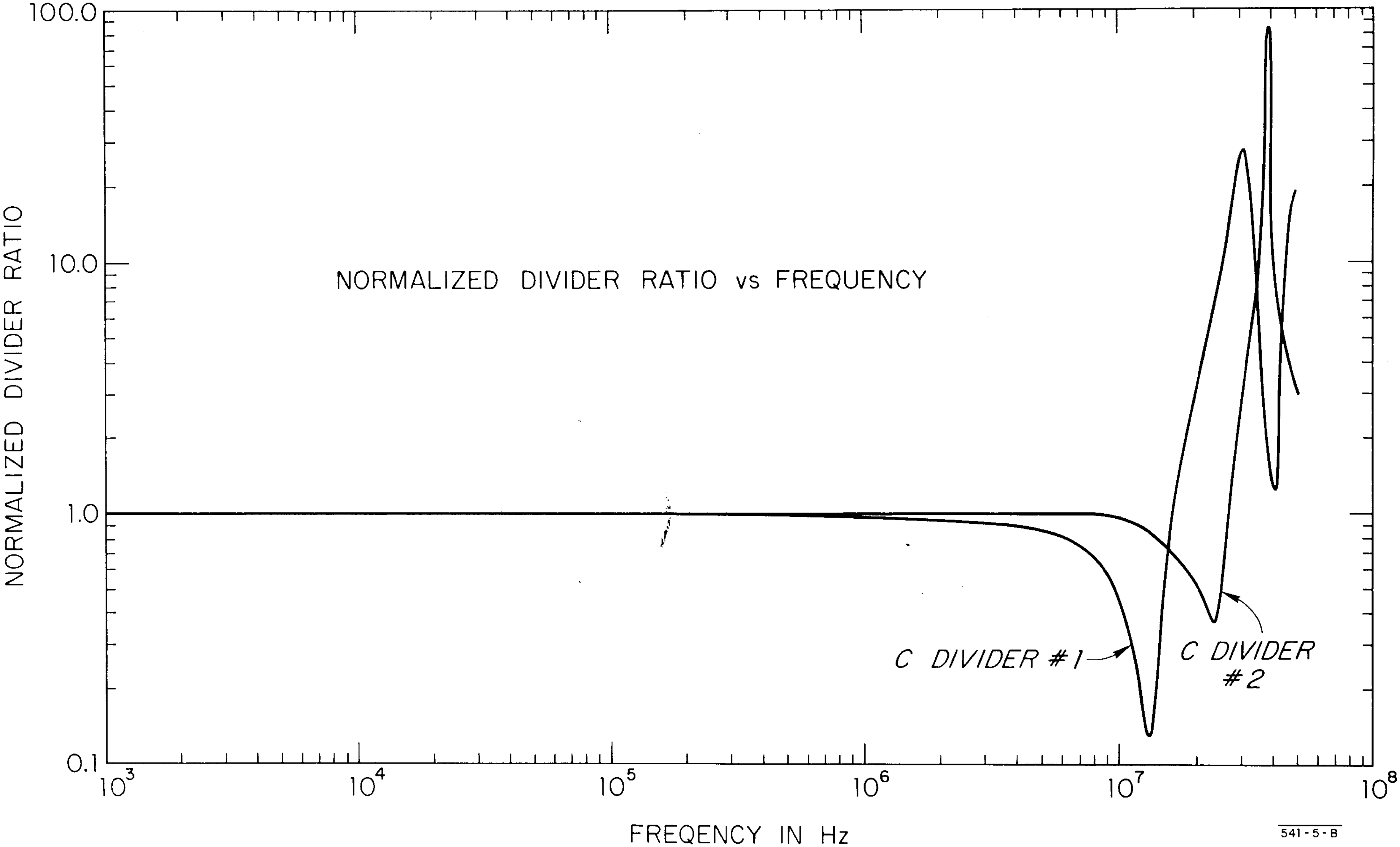


(a) (b)

541-9-A

COAXIAL GEOMETRY IRREGULARITIES

FIG. 7



NORMALIZED DIVIDER RATIO vs FREQUENCY
NORMALIZED DIVIDER RATIO
100.0
10.0
1.0
0.1
10^3
10^4
10^5
10^6
10^7
10^8
FREQENCY IN Hz
C DIVIDER #1
C DIVIDER #2
541-5-B

FIG. 8

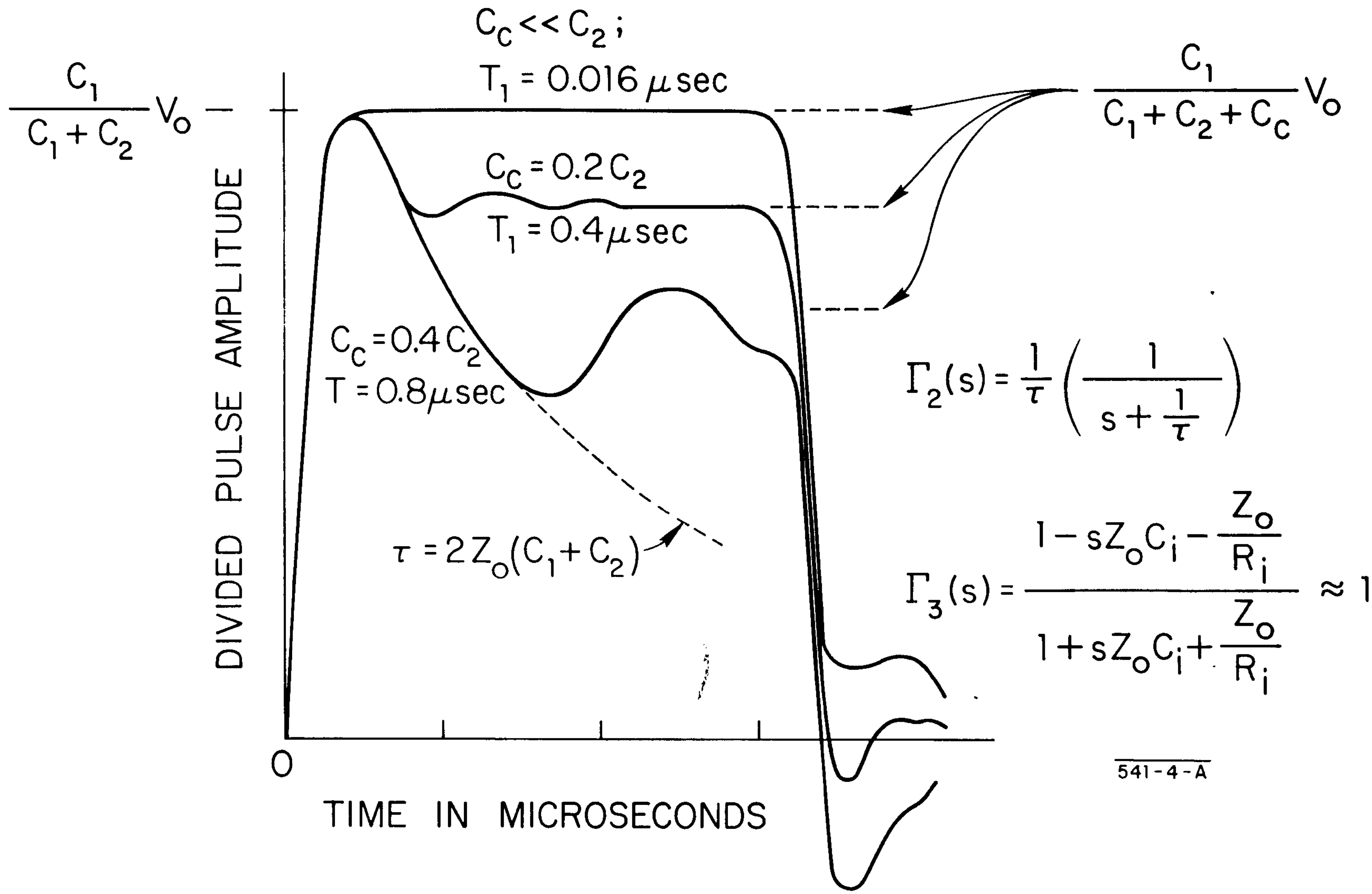


FIG. 9 – EFFECT OF TRANSIENT CABLE LOADING

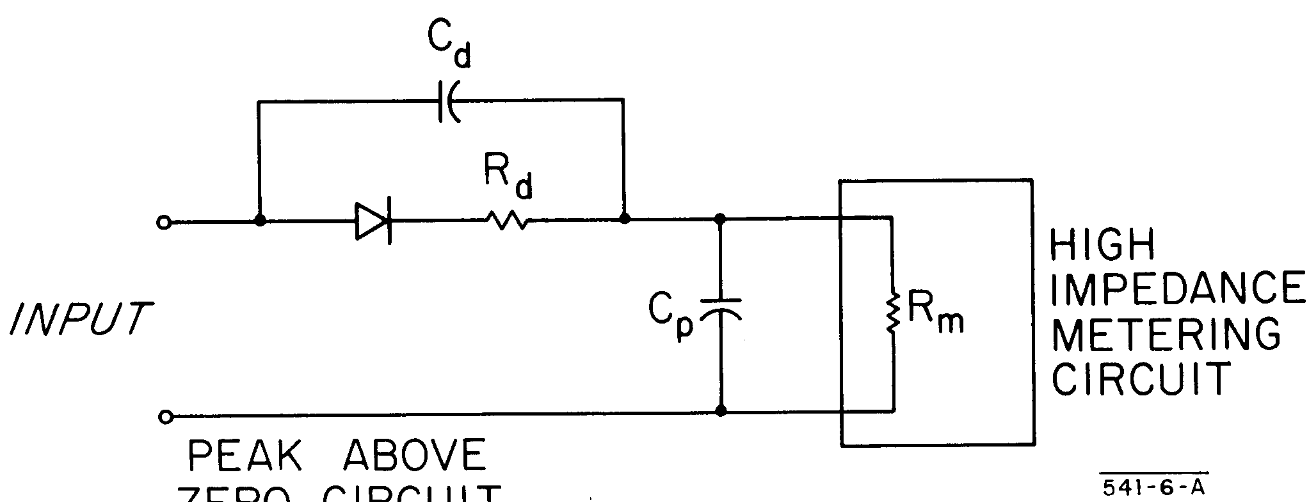


FIG. 10 - PEAK READING VOLTMETER CIRCUIT