

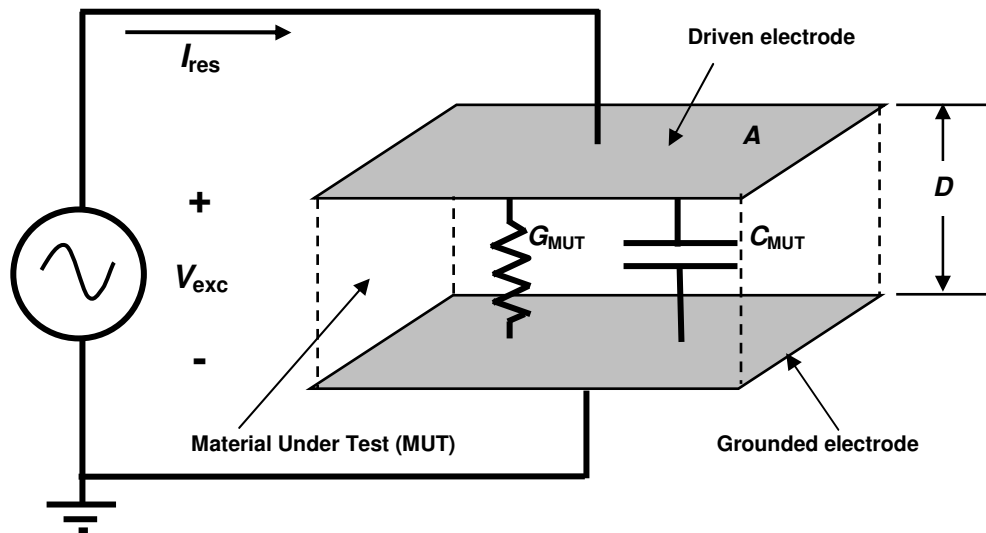


## Insight — Application Note 2.13

### Dielectric Measurement Techniques

#### Introduction

Dielectric instrumentation measures the conductance  $G$  (or resistance  $R$ ) and capacitance  $C$  between a pair of electrodes at a given frequency. The Material Under Test (MUT) between a pair of electrodes can be modeled as a conductance in parallel with a capacitance, as shown in Figure 13-1.

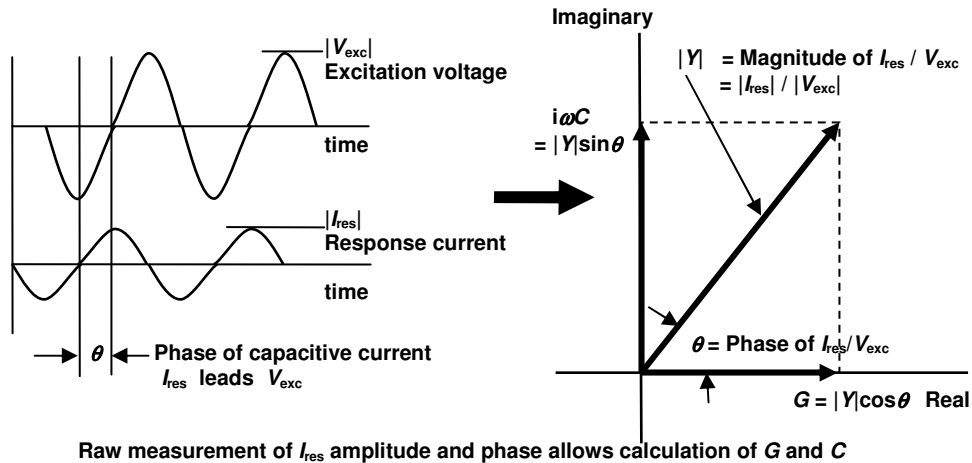


**Figure 13-1**  
**Electrical model of dielectric Material Under Test**  
**(Grounded electrode configuration)**

#### Grounded electrode configuration

Figure 13-1 also happens to depict a grounded electrode configuration for measuring the dielectric response. Note that one of the electrodes shares the ground of  $V_{exc}$ . An AC excitation voltage  $V_{exc}$  is applied between a pair of parallel plate electrodes, and the response current  $I_{res}$  is measured. Several methods exist for measuring current, but they all involve passing the current through a component such as a resistor, hall effect device, inductor or capacitor. The voltage induced across the component is measured and current is calculated from the component transfer function.

The amplitude of this current, and its phase shift relative to the excitation signal, are determined by the capacitive and conductive currents. The amplitude of the current and phase relationship between  $V_{exc}$  and  $I_{res}$  provide the information to calculate admittance  $Y$ , as shown in Figure 13-2.



**Figure 13-2**  
**Signal relationships for grounded electrode configuration**

$Y$ , and therefore the conductance  $G$  and capacitance  $C$ , of the MUT is calculated by equation 13-1:

$$(eq. 13-1) \quad Y_{MUT} = G_{MUT} + i\omega C_{MUT} = I_{res} / V_{exc}$$

where:

- $I_{res}$  = AC current through MUT (a complex number, amps)
- $V_{exc}$  = AC voltage across MUT (a complex number, volts)
- $C_{MUT}$  = Capacitance of MUT (a real number, Farads)
- $G_{MUT}$  = Conductance of MUT (a real number, ohms<sup>-1</sup>)
- $f$  = Excitation frequency (Hz)
- $\omega$  =  $2\pi f$  (angular frequency, radians/sec)

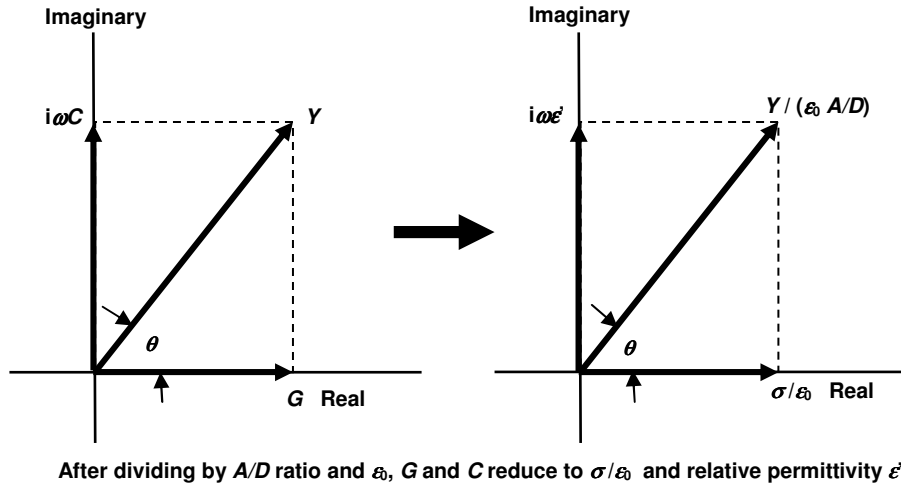
If the electrode  $A/D$  ratio is known, then the dimensionless material properties of  $\sigma / \epsilon_0$  and relative permittivity  $\epsilon'$  can be calculated from equations 13-2 and 13-3, as shown in Figure 13-3.

$$(eq. 13-2) \quad \sigma / \epsilon_0 = G / (\epsilon_0 A/D)$$

(eq. 13-3)  $\epsilon' = C / (\epsilon_0 A/D)$  (relative permittivity)

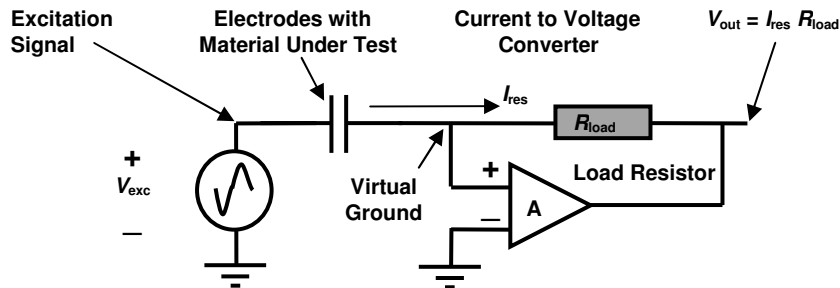
where:

$\epsilon_0 = 8.85 \times 10^{-14}$  F/cm (permittivity of free space)



**Figure 13-3**

In most implementations, the current through the MUT passes into the virtual ground of a current-to-voltage converter, as shown in Figure 13-4. An instrument can measure the voltage output,  $V_{out}$ , and current  $I_{res}$  can be calculated from the known value of  $R_{load}$ . Finally, from the amplitude of  $I_{res}$  and the phase of  $I_{res}$  relative to  $V_{exc}$ , the capacitance and conductance of the material between the electrodes can be calculated from equation 13-1.



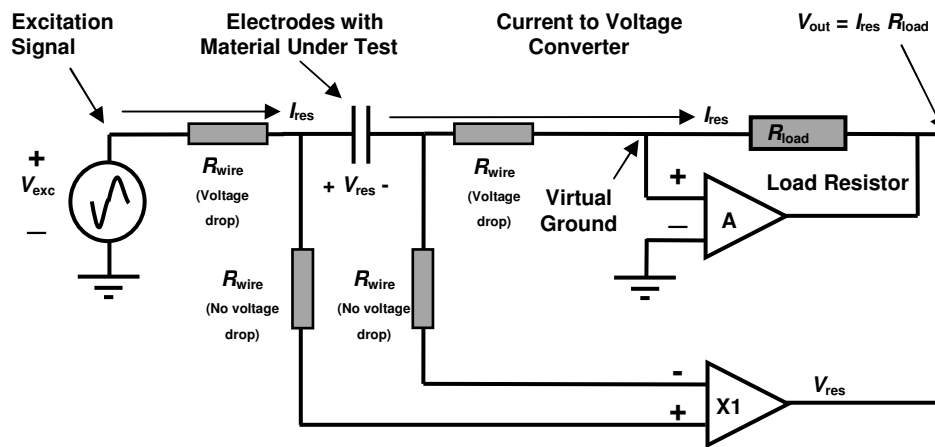
**Figure 13-4**  
Current-to-voltage conversion of  $I_{res}$

While simple, the circuit configuration of 13-4 has the disadvantage of low capacitive signal at low frequencies, and low conductive signal for highly insulating materials. These conditions can and do occur in thermosets at the end of cure, resulting in increased noise and reduced measurement accuracy during that period.

### Kelvin (four-wire) connection

The circuit of Figure 13-4 can be modified to exclude the effect of wire resistance by using a Kelvin connection, as shown in Figure 13-5. As in Figure 13-4, a current-to-voltage converter outputs a signal for measurement of the current through the MUT. By Ohm's law, current  $I_{res}$  through wire resistances produces voltage drops that reduce the response voltage  $V_{res}$  across the electrodes.

In a Kelvin connection, a differential amplifier with high input impedance measures the voltage across these electrodes, using leads that pass no current and therefore produce no voltage drop from the wire resistance. With  $V_{res}$  and  $I_{res}$ , it is possible to calculate the capacitance and conductance of the material between the electrodes using equation 13-1.



**Figure 13-5**  
**Kelvin (4-wire) connection for dielectric measurements**

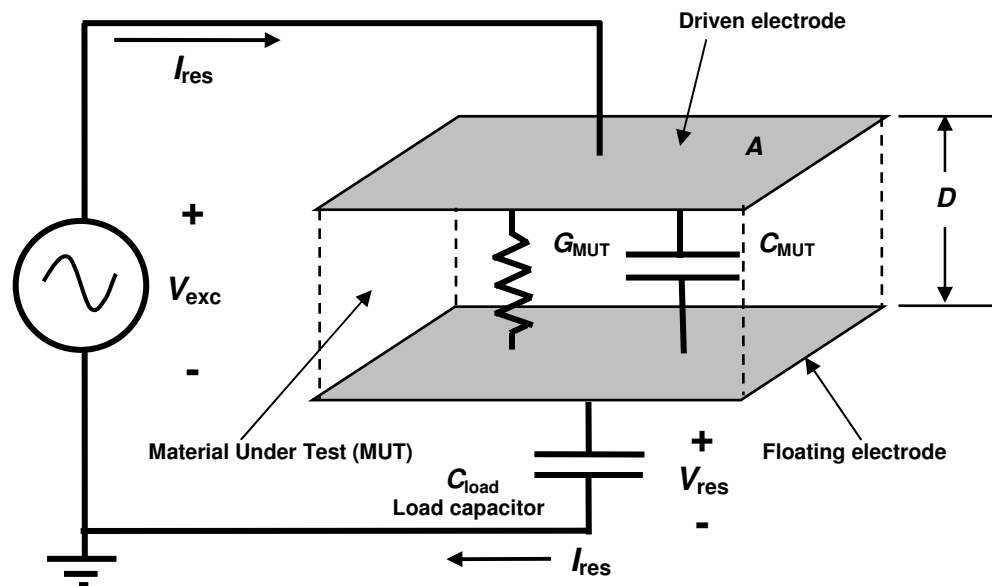
A Kelvin connection requires four wires and an extra amplifier, and is more complex than the configuration of Figure 13-4. It is most useful when the resistance being measured is small and comparable to the wire resistances. For example, 28 AWG wire has a resistance of 64.9 m $\Omega$ /ft (212.9 m $\Omega$ /m). A 20-foot

cable has 40 feet of wire (20 feet to the electrode and 20 feet from the electrode), for a total resistance of  $2.6 \Omega$ . For a measurement error of 1%, the MUT must have a resistance of about  $260 \Omega$  across the electrodes.

For typical dielectric cure monitoring applications, the resistance of material on a sensor is on the order of  $100,000 \Omega$  or greater. Without a Kelvin connection, the error from a 20-foot cable to the sensor is  $2.6 \Omega / 100,000 \Omega = 2.6 \times 10^{-5} = 0.0026\%$ . Thus, the extra complexity and cost of the Kelvin connection is difficult to justify for dielectric cure monitoring.

### Floating electrode configuration

Dielectric measurements can also be made by driving one electrode with an AC excitation voltage without grounding the second electrode, also known as the floating electrode. This floating electrode configuration is shown in Figure 13-6. Note the load capacitor  $C_{load}$  between the floating electrode and ground.

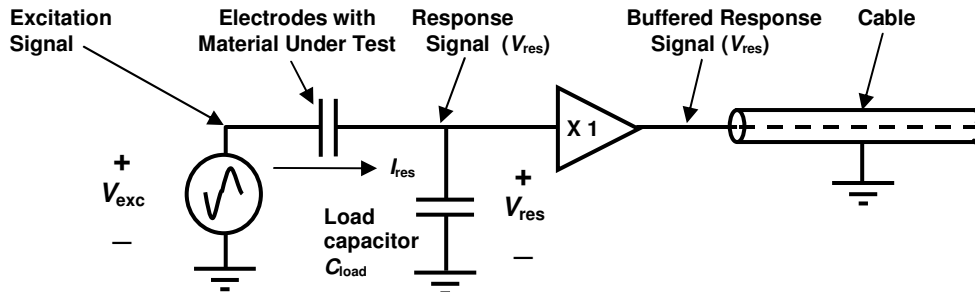


**Figure 13-6**  
**Electrical model of Material Under Test**  
**(Floating electrode configuration)**

The response voltage,  $V_{res}$ , across  $C_{load}$  depends on the load capacitance as well as the conductance and capacitance of the MUT. This load capacitance integrates current  $I_{res}$  to produce a large signal at low frequencies while also reducing noise. In effect, the current  $I_{res}$  is measured by the voltage it creates

across the load capacitance, similar to the way the current-to-voltage converter of Figure 13-4 creates a voltage across a load resistance.

In practical implementations, a x1 amplifier buffers the voltage of the floating electrode, as shown in Figure 13-7. This buffer prevents cable capacitances from affecting the floating electrode and the response signal. The buffer also drives a reproduction of the response signal through cabling to the instrumentation.



**Figure 13-7**  
**Floating electrode configuration**

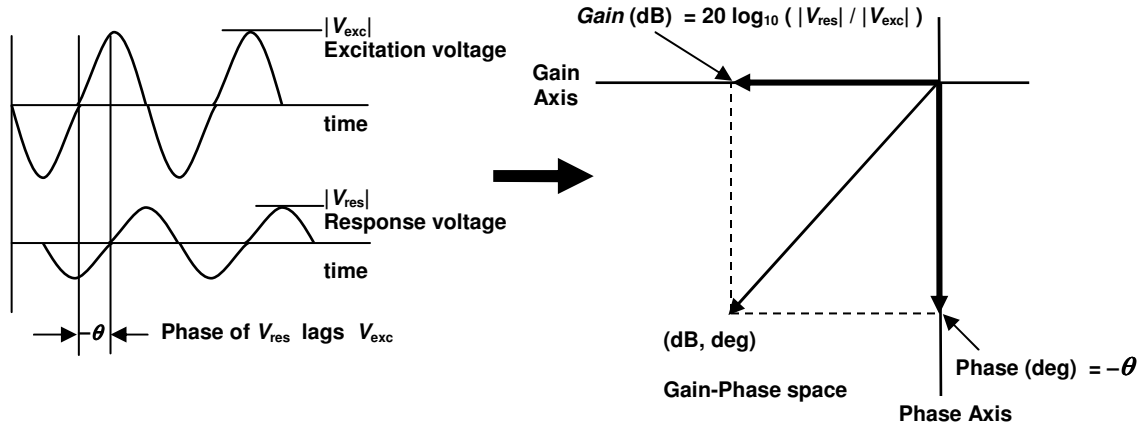
The output of the buffer is  $V_{res}$ , which has an amplitude that is attenuated and a phase that is shifted relative to  $V_{exc}$ . With a known load capacitance, the response amplitude and phase can be used to calculate the capacitance and conductance of the material between the electrodes. The load capacitance collects and integrates the current  $I_{res}$  through the MUT. As a result, the floating electrode configuration produces a larger signal with less noise compared to current to voltage conversion with a resistor. The advantages of the floating electrode configuration become especially apparent when making measurements in highly insulating materials at the end of cure.

Due to the added load capacitor, the equations to obtain conductance and capacitance for the floating electrode configuration are more complex than for the grounded electrode configuration. To span a wide dynamic range, the attenuation of the response signal is often expressed as “gain” on a logarithmic scale, as given by equation 13-4.

$$(eq. 13-4) \quad Gain \text{ (dB)} = 20 \log_{10}(|V_{res}| / |V_{exc}|)$$

The gain and phase of the response can be plotted in Gain-Phase space as shown in Figure 13-8. The use of Gain-Phase space is convenient for charting the

changing response for curing thermosets, and provides insight into the state of cure.



**Figure 13-8**  
**Signal relationships for floating electrode configuration**

Given *gain* (dB), phase  $\theta$  (deg), frequency  $f$  (Hz) and  $C_{load}$  (farad), conductance and capacitance for the floating electrode configuration can be calculated as follows:

$$(eq. 13-5) \quad |V_{res} / V_{exc}| = 10^{(Gain/20)}$$

$$(eq. 13-6) \quad A = \cos(-\theta) / |V_{res} / V_{exc}|$$

$$(eq. 13-7) \quad B = \sin(-\theta) / |V_{res} / V_{exc}|$$

$$(eq. 13-8) \quad G_{tot} = G_{MUT} = (2\pi f C_{load} B) / ((A - 1)^2 + B^2)$$

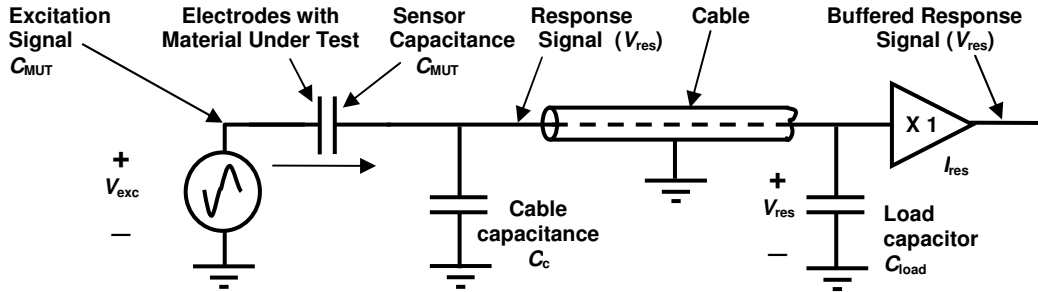
$$(eq. 13-9) \quad C_{tot} = (C_{load} (A - 1)) / ((A - 1)^2 + B^2)$$

$G_{tot}$  ( $G_{MUT}$ ) is the total conductance between the electrodes.  $C_{tot}$  is the total capacitance between the electrodes, including any stray, or base, capacitance contributed by various sources. *Note that this base capacitance must be subtracted from  $C_{tot}$  before further calculation to obtain conductivity and permittivity.*

### Subtraction and non-subtraction mode interfaces

Figure 13-9 shows a dielectric sensor in the floating electrode configuration. When the response voltage  $V_{res}$  is directly buffered by a x1 amplifier, the interface operates in *non-subtraction* mode. In this arrangement, load capacitance  $C_{load}$  is grounded and any cable capacitance on the response signal is in parallel with  $C_{load}$ . If the load capacitance is very large compared to

sensor capacitance  $C_{MUT}$ , then cable capacitance has relatively little effect on the response signal.



**Figure 13-9**  
**Non-subtraction mode interface**

From basic circuit theory, the response with capacitances only is:

$$(eq. 13-10) \quad V_{res} / V_{exc} = C_{MUT} / (C_{load} + C_{MUT} + C_c)$$

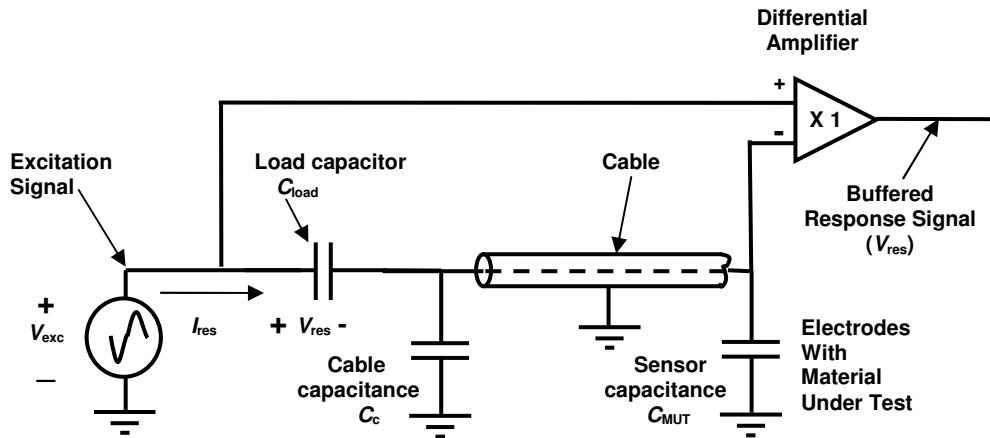
For example, if  $C_{load}$  is 2000 pF and  $C_{MUT}$  is 20 pF, for various cable capacitances the response for the interface of Figure 13-9 is shown in Table 13-1. Even for relatively large cable capacitances, the effect on the response is small.

**Table 13-1**  
**Response of Non-subtraction Mode Interface**

$C_{load}$ (pF)	$C_{MUT}$ (pF)	$C_c$ (pF)	$V_{res} / V_{exc}$	Gain (dB)	$\frac{V_{res} / V_{exc}}{V_{res} / V_{exc} \text{ for } C_c = 0 \text{ pF}}$
2000	20	0	0.0099	-40.09	1.00
2000	20	20	0.0098	-40.17	0.99
2000	20	40	0.0097	-40.26	0.98
2000	20	60	0.0096	-40.34	0.97
2000	20	100	0.0094	-40.51	0.95

In contrast, Figure 13-10 shows a dielectric sensor interface operating in *subtraction mode*. Here the sensor capacitance  $C_{MUT}$ , not the load capacitance, is grounded. A differential amplifier subtracts the sensor signal from the excitation signal, resulting in a response that ideally is the same as for non-subtraction mode.





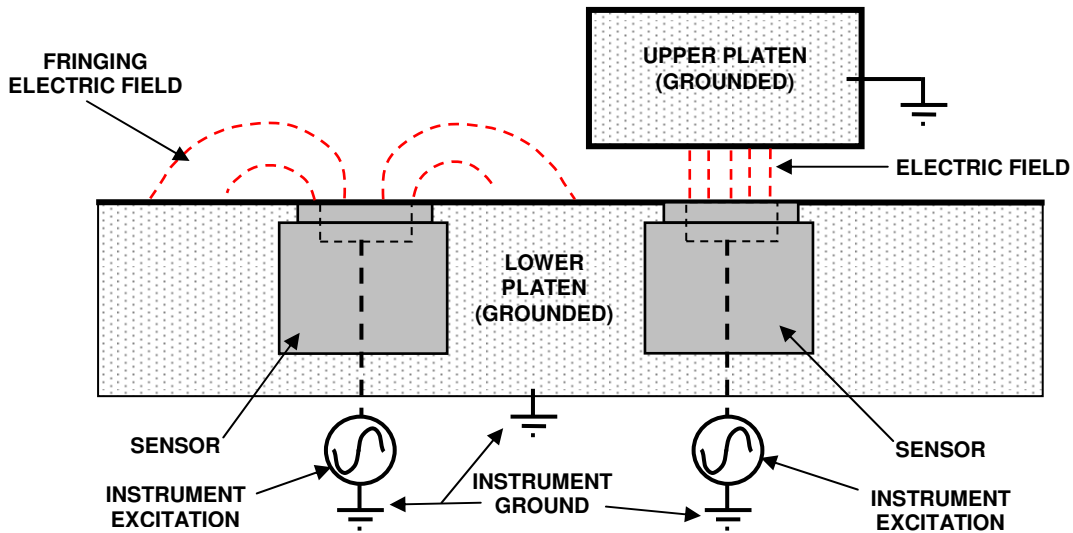
**Figure 13-10**  
**"Subtraction" mode interface**

Both subtraction mode and non-subtraction mode circuits use equations 13-4 through 13-9 to convert the response signal to conductance and capacitance. Subtraction mode has the advantage of allowing a dielectric sensor to operate with a single electrode, such as the sensor of Figure 13-11.



**Figure 13-11**  
**Dielectric sensor with single electrode**

A sensor with one electrode uses a nearby grounded surface as an effective second electrode. This surface may be a mold or the platen of a press, either surrounding the sensor or the opposite it as shown in Figure 13-12. Thus a single electrode sensor may be smaller than one that requires two electrodes.



**Figure 13-12**  
**Single electrode sensor using nearby grounded surface as second electrode**

Non-subtraction mode requires a sensor with two electrodes, while a subtraction mode interface may use sensors with either one or two electrodes. A subtraction mode interface, however, is more influenced by cable capacitance than a non-subtraction mode interface. As apparent in Figure 13-10, for a subtraction mode interface the cable capacitance is in parallel with the sensor capacitance.

From circuit theory, the response with capacitances only is:

$$(eq. 13-11) \quad V_{RES} / V_{EXC} = (C_{MUT} + C_C) / (C_{LOAD} + C_{MUT} + C_C)$$

Again, for example, if  $C_{load}$  is 2000 pF and  $C_{MUT}$  is 20 pF, for various cable capacitances the response for the interface of Figure 13-10 is shown in Table 13-2.

**Table 13-2**  
**Response of Subtraction Mode Interface**

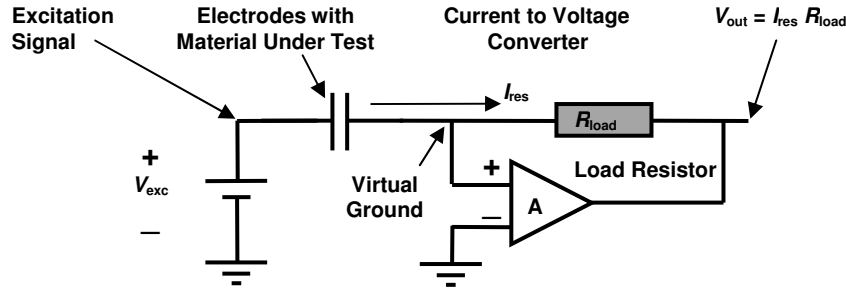
$C_{load}$ (pF)	$C_{MUT}$ (pF)	$C_c$ (pF)	$V_{res} / V_{exc}$	Gain (dB)	$\frac{V_{res} / V_{exc}}{V_{res} / V_{exc} \text{ for } C_c = 0 \text{ pF}}$
2000	20	0	0.0099	-40.09	1.00
2000	20	20	0.0196	-34.15	1.98
2000	20	40	0.0291	-30.71	2.94
2000	20	60	0.0385	-28.30	3.89
2000	20	100	0.0566	-24.94	5.72

Even for relatively small cable capacitances, the influence on the response is large. The effect of cable capacitance may be subtracted from the response, but uncertainty in the value of cable capacitance will correspondingly affect the accuracy of sensor capacitance measurements.

In subtraction mode, conductance measurement is not affected by cable capacitance. For dielectric cure monitoring, when conductance is the property of interest, subtraction mode offers flexibility in the choice of sensors. For measurements when capacitance is also important, then non-subtraction mode should be used.

### DC measurements

Instruments for measuring only the conductance of the material under test have the advantage of simplicity compared to those measuring both conductance and capacitance. These conductance measurement instruments (essentially highly sensitive ohmmeters) use a DC voltage source to drive current through the MUT between the sensor electrodes. This current typically passes into the virtual ground of a current-to-voltage converter, as shown in Figure 13-13. A data acquisition system can measure the DC voltage output,  $V_{out}$ , and DC current  $I_{res}$  can be calculated from the known value of  $R_{load}$ . From the value of  $I_{res}$ , the frequency independent (DC) conductance of the material between the electrodes can be calculated from Ohm's law.



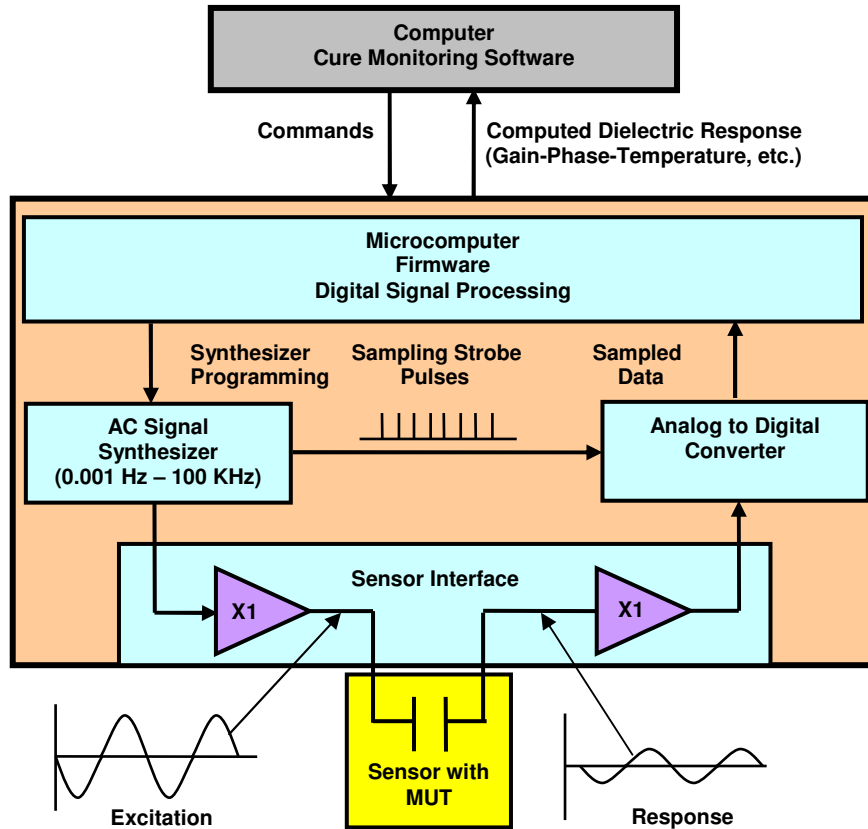
**Figure 13-13**  
**DC measurement of dielectric properties (conductance only)**

The simplicity of the circuit of 13-13, however, comes at the expense of flexibility and possibly accuracy. DC methods can only measure the DC conductance; measurement of capacitance requires an AC excitation. DC measurements may be confounded by offset voltage drifts, thermal drifts and leakage currents in the interface electronics, which cannot be distinguished from the true DC signal.

Finally, DC measurements are not possible with release layers, which are very thin insulating sheets sometimes used to prevent material from adhering to a mold or platen. Release layers block DC current and therefore prevent measurements of DC conductance.

### Digital sampled data system

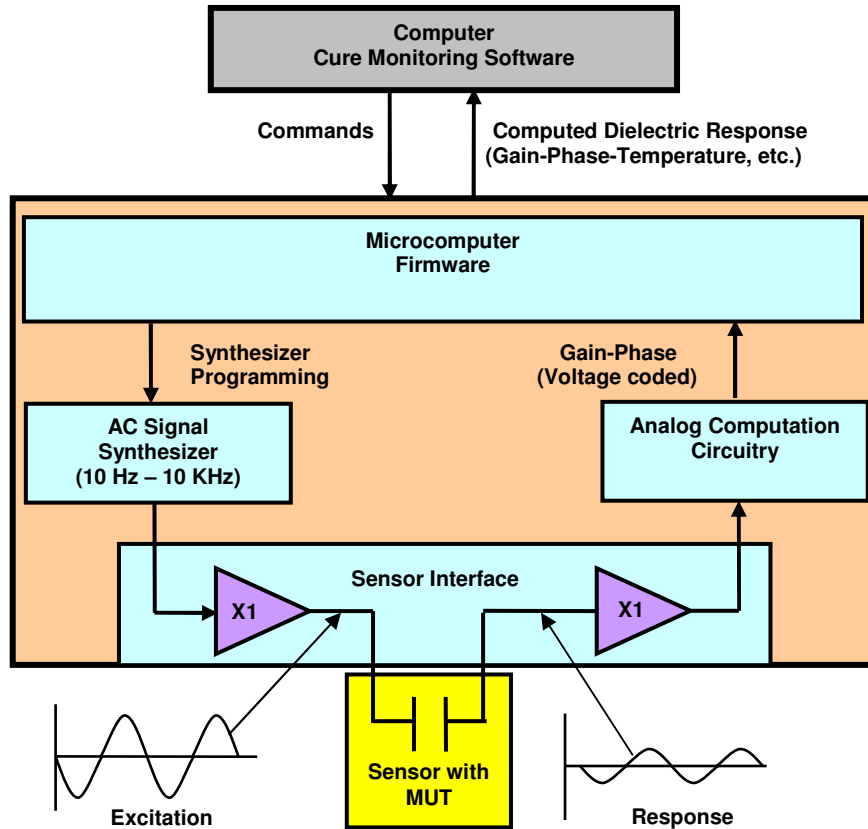
Some dielectric measurement instruments use sampled data acquisition, as shown in the block diagram of Figure 13-14. A sampled data acquisition system analyzes the sensor response using digital signal processing (DSP), which can extract information from a very noisy signal. This ability results in more accurate, precise and repeatable data than from an instrument using analog computation, but with greater complexity and cost.



**Figure 13-14**  
**Block diagram of a sampled data acquisition system**

## Analog computation system

Analog measurement techniques are less expensive and much faster, but more noise sensitive and less accurate, than digital sampled data systems. However, analog instruments can perform well in applications where rapid data acquisition is a desired tradeoff for accuracy and precision. The block diagram of Figure 13-15 shows how an instrument can use analog computation to measure dielectric properties.



**Figure 13-15**  
**Block diagram of analog computation system**



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