High-Frequency Response of Micropotentiometers

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Abstract

The RF-dc differences of a micropotentiometer(μpot) are analyzed, calculated, and measured. Significant improvements have been made to reduce the RF-dc differences. Observations show good stability over a long period, which makes μpots suitable as primary RF and audio standards in the microvolt and millivolt ranges.

Introduction

Significant progress has been made in the last decade to extend the frequency range of the RF μpot down to 10 Hz[1,2,3] and up to 1.2 GHz. The measurement accuracies have also been improved. Since μpots are used as primary standards for low voltages, it is necessary to investigate all possible error sources when determining the RF-dc differences. Equations will be described for calculating the RF-dc differences of μpots, determining measurement errors, and what precautions to take when performing the measurements in order to achieve the best accuracies.

Figure 1 shows a diagram of the RF μpot assembly. When excited with an external RF source, it is designed to provide a precisely determined voltage, V_{RF}, at its output terminal. The input current, I_h, flows through a UHF-type thermocouple to a disk resistor, R_o; and the voltage drop across R_o is a low-impedance source of rf voltage. The output voltage is nominally the product of the heater current and the resistance of the disk resistor. To provide additional voltage ranges, the disk resistor is replaced with another of smaller or larger resistance value, or may be inserted in another housing of a different current rating.

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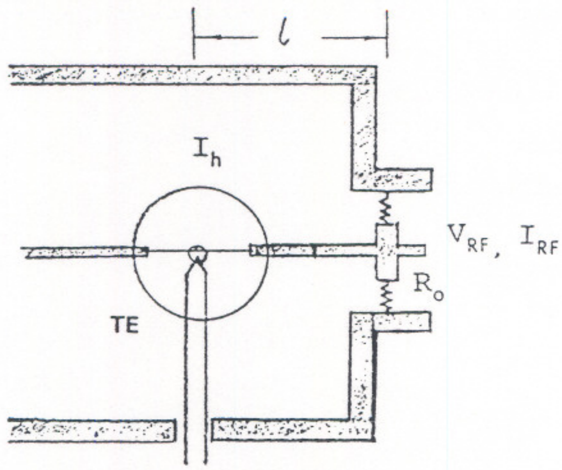

Fig. 1 Diagram of An RF Micropotentiometer

RF-DC Differences of a Micropotentiometer

The μpot is usually characterized by its RF-dc difference, d, which is defined by:

$$d= \frac{V_{RF}}{V_{dc}} - 1 \qquad (1)$$

where V_{RF} is the RF output voltage of the μpot and V_{dc} is the average of the positive and negative dc voltage outputs for the same EMF output of the thermoelement.

The design considerations and characteristics of the μpot which affect its RF-dc differences are discussed below.

(1) Current errors caused by transmission line effect

A vacuum thermoelement is usually used to measure the RF current through the μpot. Any standing wave which exists will cause errors in measuring the RF currents, the current at the center of thermoelement, I_{hRF}, is less than the current at the disk resistor I_{RF}.

According to the transmission line equation:

$$I_{hRF} = V_{RF}\sinh(ZY)^{0.5}/(Z/Y)^{0.5} + I_{RF}\cosh(ZY)^{0.5} \qquad (2)$$

where:
V_{RF} phasor output voltage of μpot
I_{RF} phasor current at the center of disk resistor
I_{hRF} phasor current at the center of heater of thermoelement
I_{dc} dc current through the heater of thermoelement
l distance from the center of heater to the center of disk resistor, see Fig. 1

L distributed inductance of heater and lead with length l
R distributed resistance of heater and lead with length l
C distributed capacitance of heater and lead with length l to the shield
Z $= R + j\omega L$
Y $= j\omega C$

If the disk annular resistor, R_o, is a thin-film resistor constructed using good state-of-the-art techniques, the distributed parameters can be neglected. Thus $V_{RF}= I_{RF}R_o$,

$$I_{hRF} = I_{RF} [R_o(Y/Z)^{0.5} \sinh(ZY)^{0.5} + \cosh(ZY)^{0.5}]$$

and the RF-dc difference caused by the current measurement, d_c, is simply:

$$\begin{aligned} d_c &= (V_{RF} - V_{dc})/V_{dc} \\ &= (I_{RF} - I_{dc})/I_{dc} \end{aligned} \qquad (3)$$

Then combined equation (3) and above equation:

$$d_c = \{1/[R_o(Y/Z)^{0.5}\sinh(ZY)^{0.5}+\cosh(ZY)^{0.5}]\}-1 \qquad (4)$$

Generally, R_o ranges from 1 milliohm to 22 ohms, and the characteristic impedance $(Z/Y)^{0.5}$ is about 200 ohms, thus $R_o(Y/Z)^{0.5} \ll 1$:

$$d_c \approx [1/\cosh(ZY)^{0.5}] -1$$

Since $\omega CR \ll 1$, and if $\omega L/R \ll 1$:

$$\begin{aligned} d_c &\approx \omega^2(LC-C^2R^2/6)/2 \\ &\approx \omega^2 LC/2 \end{aligned}$$

For a coaxial line, $L/l=2\times10^{-7}\ln(b/a)$ in henries per meter, $C/l=1/1.8\times10^{10}\ln(b/a)$ in farads per meter, a is the outer diameter of inner conductor, b is the inner diameter of outer conductor, then

$$d_c \approx 2.19\times10^{-20} l^2 f^2 \qquad (5)$$

where the units are l: cm, f: Hz. In order to reduce the inductance of the thermoelement, it is necessary to reduce the size of the housing, and the length l as much as practical. An experiment has shown that reducing the length about 2 mm in a commercial unit the RF-dc difference was decreased 1.5% at 900 MHz.

The thermoelement in the μpot must sometimes be replaced due to some failure. The RF-dc differences are usually not appreciably affected if the replacement is the same type thermoelement and inserted at the same location.

(2)Ground impedance

Ground impedance such as RF resistance and equivalent inductance, which are present at high frequencies, increases the equivalent length, l, resulting in differences between the calculated RF-dc differences according to equation (5) and measurement results. Furthermore, the RF-dc differences will be different for various materials of the μpot housing. The calculated RF-dc differences for Ballantine Micropotentiometer Model 440* caused by the current transmission line effect and ground impedance according to equation (5) are shown in Table 1.

Table 1. Estimated RF-dc differences caused by current transmission line effect and ground impedance

Estimated equivalent length(cm)	d_c	RF-DC Differences (%)			
		10MHz	50MHz	100MHz	500MHz
2.3	$11.6X10^{-20}f^2$	0.0012	0.029	0.12	2.90
3.0	$19.7X10^{-20}f^2$	0.0020	0.049	0.20	4.93

(3)Skin effect of the heater and its inner leads

Due to Skin effect at high frequencies, the RF resistances of the heater and its inner leads are larger than their dc resistances. Many commercial vacuum thermoelements are made with copper-coated magnetic leads, so the skin effect is larger than for non-magnetic leads. There is also a skin effect in the platinum leads. Therefore, the RF current is less than the dc value for the same EMF output causing negative RF-dc differences. The larger the current range, the bigger the diameter of the heater, the smaller the heater resistance, and the larger the skin effect. The RF-dc differences of μpots caused only by the skin effect in the thermoelement, d_s, is derived as follows:

For the same EMF output, $I_{dc}^2 R_{hdc} = I_{hRF}^2 R_{hRF}$

Then $d_s = I_{RF}/I_{dc} - 1$
$\approx I_{hRF}/I_{dc} - 1 \approx (R_{hdc}/R_{hRF})^{0.5} - 1$

Let $\Delta = R_{hRF}/R_{hdc} - 1$

Using Taylor series expansion:

$$d_s \approx -\Delta/2 \approx Af^{0.5} \qquad (6)$$

*Certain commercial equipment, instruments, or materials are identified in this paper in order to adequately specify the experimental procedure. In no case does such identification imply recommendation or endorsement by National Institute of Standards and Technology, nor does it imply that the material or equipment identified is necessarily the best available for the purpose.

where:
R_{hdc} dc resistance of heater and its internal leads
R_{hRF} RF resistance of heater and its internal leads
A coefficient determined by individual thermoelement

When frequency is higher than 1 MHz, Δ is proportional to the square root of frequency. This error can compensate for the current error caused by the transmission line effect.

(4)Errors of annular resistor

When the thickness of the resistance film is much smaller than its skin depth, the inductance and skin effect of a well made annular resistor can generally be neglected. However, the RF-dc differences may be very large if the annulus is damaged.

When the resistance value is very small, e.g., several milliohms for the lower microvolt ranges, the thickness of the resistance film is relatively large and is comparable to its skin depth. The film annulus can be considered as a section of coaxial line with a solid conductor as the propagation medium. Using the same method which
M. Selby[4] used, the transfer impedance, Z_m, that is the output voltage of annular resistor versus the input current of the same resistor, is:

$$Z_m = R_o[(1+j)t/\delta]/\mathrm{Sinh}[(1+j)t/\delta]$$

the RF-dc difference, d_R, caused by the annulus is:

$$d_R = |Z_m|/R_o - 1 \qquad (7)$$

Then, using Taylor series expansion:

$$d_R = 1/|1 + jt^2/3\delta^2 - t^4/30\delta^4 - jt^6/630\delta^6 + \cdots| - 1 \qquad (8)$$

where t is the thickness and δ is the skin depth of the annulus. The calculated RF-dc differences caused by the annulus are negative and large, when the resistance value is low, then the resistor film is thicker, and this is why the RF-dc differences of microvolt range μpot's are usually negative.

(5)Transmission line effect of output connector

Transmission line error is the voltage difference between the reference planes of the test instrument and the standard. At frequencies above a few MHz the distance between reference planes is not negligible compared to a wavelength. Therefore the transmission line effects needs to be considered, especially when the impedance of the interconnections and the input impedance of the test device are not matched. The RF voltage will usually rise from the center of the film annular resistor to the reference plane of the test device, and continue to rise to the input load of the test device. Although the transmission line effect can be calculated[4,5,6], the measurements of the load impedance and the total length of the actual transmission line are difficult to obtain practically. When the input impedance of the test device is much higher than the characteristic impedance of the interconnection line, the calculation formula is as simple as the following[5]:

$$d_T \approx 2.19\times10^{-20} l_T^2 f^2 \tag{9}$$

where l_T is the distance between the film annular resistor and the front face of the input load of the test device. When performing high-frequency voltage calibrations, all interconnecting devices, adapters and leads between the μpot and instrument under test should be as short as possible in order to reduce transmission line errors.

(6) Empirical equation

From the above analysis, we can obtain the empirical equation for the RF-dc difference of a μpot:

$$\begin{aligned} d &= d_o + d_c + d_s + d_R + d_T \\ &\approx d_o + Af^{1/2} + Bf^2 \end{aligned} \tag{10}$$

where d_0 is the ac-dc difference at audio frequency which can be measured directly, A is related to the skin effect, and B is related to the distributed inductance and capacitance of the μpot, and others. A and B can be derived by a non-linear fit from the RF-dc differences of the μpot to be measured. The calculated RF-dc differences from equation (10) agree with the measurement data. At higher frequencies, higher-order frequency terms, such as f^4 can not be neglected, so they appear in the equation also.

(7) The improvements in the new design

Based on the theory described above, a new commercial μpot housing has been designed and constructed recently. The RF-dc differences of μpots with these housings are reduced significantly; the measurement results are shown in Fig. 2.

Measurement Method and Results

(1) Calibration method and uncertainties

The primary standard μpots are calibrated against a bolometer at NIST, Boulder. The bolometer is connected at the output of the μpot. The RF output voltage of the μpot is derived from the power and impedance measurements for the bolometer.

The calibration of a μpot may be performed by comparison with a calibrated μpot[7] or an RF thermal voltage converter. In both cases, a transfer or comparison device such as an RF receiver is used to transfer the RF output voltage of the standard to the μpot under test. The block diagram of a typical measurement system is shown in Fig. 3.

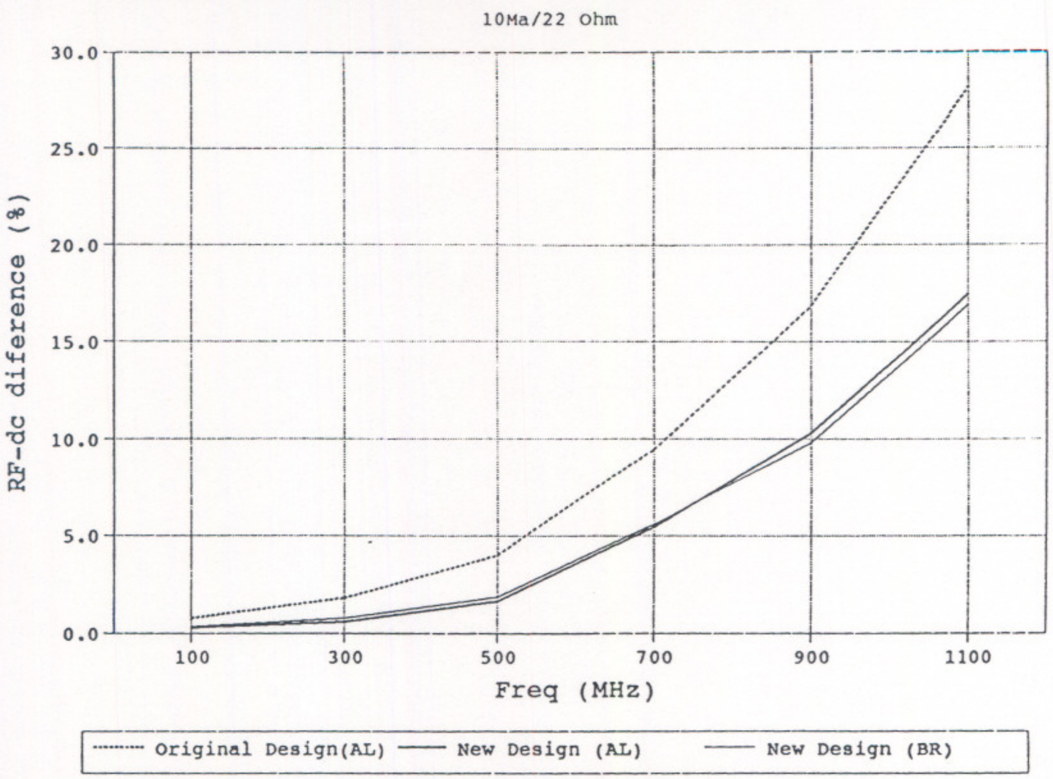

Fig. 2 The RF-dc differences of different designs of Model 440 μpots (AL: Aluminum, BR: Brass)

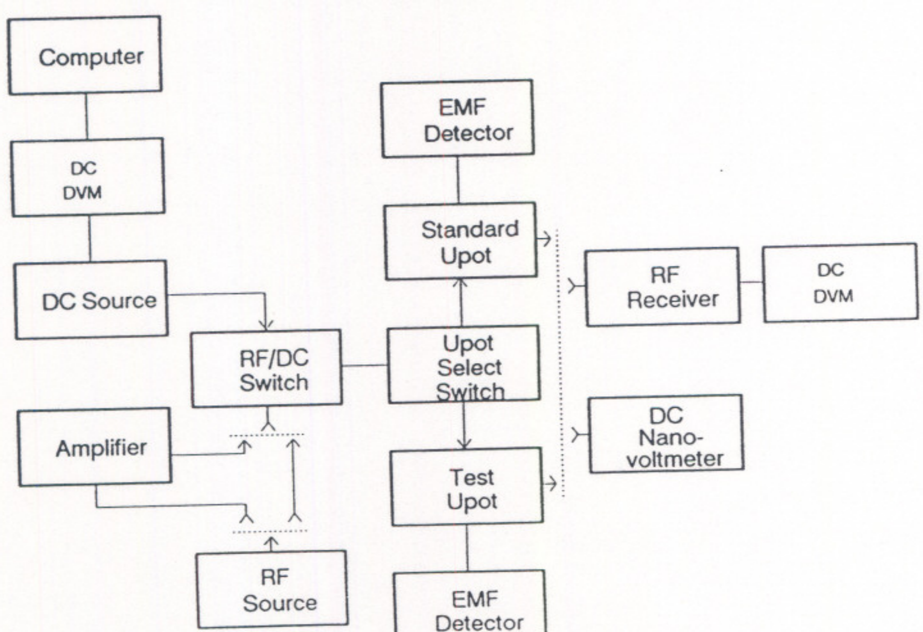

Fig.3 The block diagram of a μpot measurement system