Another Calibration Method for Microcalorimeters

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Abstract-In the paper is proposed an original calibration method for microcalorimeter measurements on thermistor power sensors and certain preliminary experiments are illustrated, which are necessary for the effective application of the method. This new method avoids some critical steps in the usual microcalorimeter calibration, saves the time and does not require vectorial measurements for the feeding path sections.

I. Introduction

Microcalorimeter is a primary power standard used for measuring the effective efficiency of transfer standard, as bolometers [1]. Bolometers require a *direct current* (*dc*) bias power, which allows maintaining a quite constant power development inside of the microcalorimeter. Feeding lines have a negative contribution on power equilibrium and therefore their perturbation must be determined through microcalorimeter calibration. *S*-parameter measurements for feeding lines losses computation [1] and dummy loads, as ideal sensors, are normally used for performing the microcalorimeter calibration. An uncertainties of 2% for effective efficiency, η_e, can be obtained also at frequencies above 18 GHz. A part of this uncertainty comes from *S*-parameter measurements because these quantities, although small, have a great weight in the propagation. The proposed calibration technique should improve the mentioned uncertainty.

II. Method Description

The method we propose is based on the measurement set-up schematized in Fig. 1. Basically, this is a digitally controlled current-loop whose function is to maintain constant the power dissipation inside the microcalorimeter while measurement data are acquired through a data acquisition board-DAQ and a dedicated nanovoltmeter for the output of the microcalorimeter thermopile.

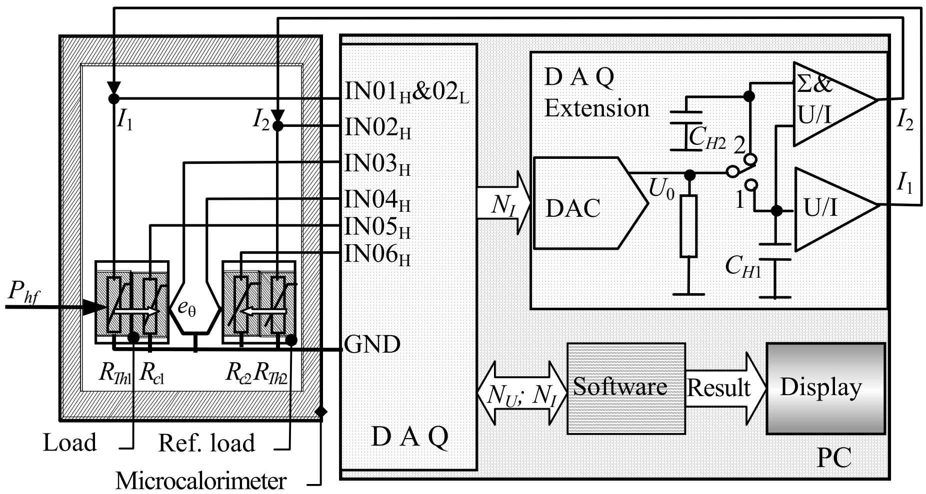


Figure 1. Instrumentation set-up for microcalorimeter measurements in a full-digital loop

Assuming an ideal behavior, all the thermal contributions of the feeding paths appear like common-mode quantities and therefore they are rejected. This however happens only if the feeding lines have similar insertion losses, the thermopile is a perfect differential detector and the same power separation occurs on the two channels. In practice we are more or less far from these conditions.

The used DAQ has multiple differential analogue input channels and a digital output bus. For our purpose, a DAQ extension card is anyway necessary with at least two current outputs and maximum output current large enough to satisfy the balance in power for different bolometer power sensor types. The main role of the DAQ board is the multi-channels A/D conversion. The measurement result is the effective efficiency of the bolometer power sensor, roughly determined from thermopile data measured with nanovoltmeter and from the bolometer bias power measured with the 12 bits DAQ.

Thermistor power sensors are working by equivalence between thermal effects in *dc* and *high frequency* (*hf*). By keeping its dynamic resistance R_{Th0} forced at a fixed value, any changes in the input power P_{inS} will lead to changes in the bias *dc* power from P_{dc1} to P_{dc2}. Starting from electro-thermal equilibrium equation for an ideal microcalorimeter and from usual effective efficiency definition we obtain [1]:

$$\eta_e = \frac{\Delta P_{dc}}{\Delta P_{dc} + P_{xS}} = \frac{1}{1 + (e_{on} - e_{off}) / \left(G_1 \cdot \Delta P_{dc}\right)}, \quad \Delta P_{dc} = P_{dc1} - P_{dc2} \tag{1}$$

where e_{off} and e_{on} are output voltages of the thermopile with and without *hf* power applied, G_1 is its sensitivity in volts/watt and P_{xS} denotes high frequencies sensor losses. Using thermo-voltage ratios e_R and power ratios p_R, defined in the following, it results:

$$\eta_e = \frac{1 - p_R}{e_R - p_R}, \quad p_R = \frac{P_{dc2}}{P_{dc1}} \text{ and } e_R = \frac{e_{on}}{e_{off}} = 1 + \frac{P_{xS}}{P_{dc1}} \tag{2}$$

An important error affects the result because (2) does not take in account feed line losses contribution [3]. Following [1], the power separation coefficients determined in microcalorimeter calibration step [2], would correct it satisfactory [3]. The proposed new method avoids this step by using a thermal feedback loop. We connect two identical thermistor mounts; one of them has to work as thermal reference, inside a twin coaxial microcalorimeter. In commercial power sensors there is normally a secondary thermistor, R_c, used for temperature compensation [1]. In microcalorimeter technique, this thermistor is unused, but now we consider it as an auxiliary thermometer. Though thermistors are nonlinear thermal sensors [4] and therefore not very suitable for direct temperature measurements, through a control loop we can obtain $R_{c1} = R_{c2}$, that is, the same temperature for both power sensors. The whole power dissipated in the load, that is *dc* plus *hf* power, may be determined from *dc* power dissipated in the reference load. Assuming the equivalence of the thermal effects, a simplified equation for effective efficiency results:

$$\eta_e = \frac{P_{dc11} - P_{dc12}}{P_{dc22} - P_{dc12}} \cong \frac{\Delta P_{dc1}}{\Delta P_{dc1} + \Delta P_{dc2}} \approx 1 - \frac{\Delta P_{dc2}}{\Delta P_{dc1}} = 1 + \frac{\left(N_{U2} N_{I2}\right)_1 - \left(N_{U2} N_{I2}\right)_2}{\left(N_{U1} N_{I1}\right)_1 - \left(N_{U1} N_{I1}\right)_2} \tag{3}$$

The second finger in the indexes corresponds to measurements on the unknown power sensor, without, and with *hf* power applied to it. The thermopile seems to be unnecessary in these conditions, but it confirms obtaining of the balance. If there are measurable differences, the reference load, connected to a short-circuit terminated feeding line, must be supplied with about half the power that feeds the line of the load. The equivalence of unknown and reference power sensor is necessary only in *dc* bias. The power contribution of the temperature sensors, R_c, must be near zero.

III. Preliminary experiments

Considering that the proposed balancing method is based on a digital loop, the power sensor needs a specific characterization before implementing the control algorithm. We had begun our research from the determination of the mount parameters in a working configuration as proposed for absolute power measurement. In this case, only a single power sensor was supplied with *dc* power and the effect was measured by recording the resistance variation. Both detection and compensation thermistors are *dc* biased, though the power dissipated in the compensation element is very small. This means a low value for I_2, which is used as test signal for resistance measurement based on the Ohm law. The current I_1 must have a step variation: high values for developing the test power in the heating state and small values for measurements of the thermistor resistance in the cooling state. The virtual instrument of Fig. 2, implemented with Visual Basic package, allows setting the data record length and the adjustment of the power pulse duty factor. This last factor was 50% from the recording time T_m of 60 seconds. The current I_1, generating the power pulse, was selected between 5 mA and 20 mA in steps of 5 mA. After 30 seconds, this current drops to the same I_2 level of 1 mA.

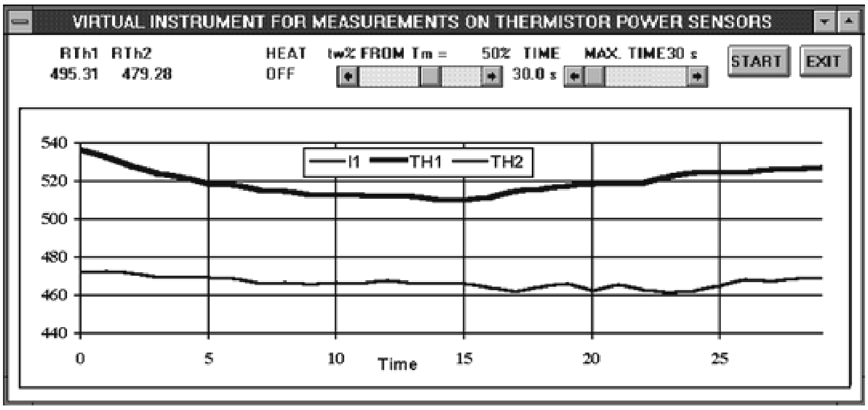


Figure 2. Panel status of the virtual instrument after a recording time T_m = 30 s; I_I = 5 mA, t_W = 15 s

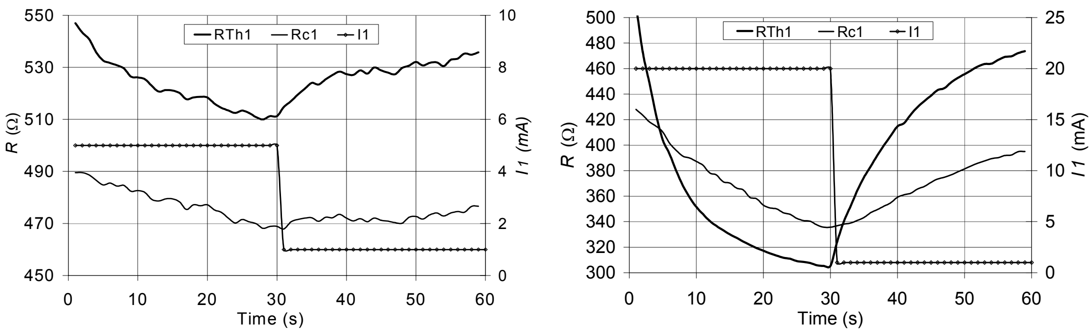
The resistance is computed from voltage values measured with DAQ input channels, while the values of the currents are known as output values:

$$R_{Th1} = R_{SCAL}\frac{N_{U1}}{N_{I1}} \quad \text{a)}; \quad R_{c1} = R_{SCAL}\frac{N_{U2}}{N_{I2}} \quad \text{b)} \tag{4}$$

where R_{SCAL} is the ratio between the range of the measuring voltage channel and the nominal value of the output current, 200 Ω in this case. The resistances, R_{Th1}, R_{c1}, the current I_I and the time, are stored and graphically represented in the same virtual instrument panel of Fig. 2. In this representation, R_{Th2} is in fact R_{c1}. In this manner, appropriated experiments have been done and representative data sets were obtained.

IV. Computation and results

The resistance of the compensation thermistor is strongly dependent of the power dissipated in the detection element. In the following graphs, two different cases are shown, differing for the injected *dc* power. In the first case weak power was used, about 5 mW at the equilibrium, and in the other, high power was applied, about 80 mW at the equilibrium.



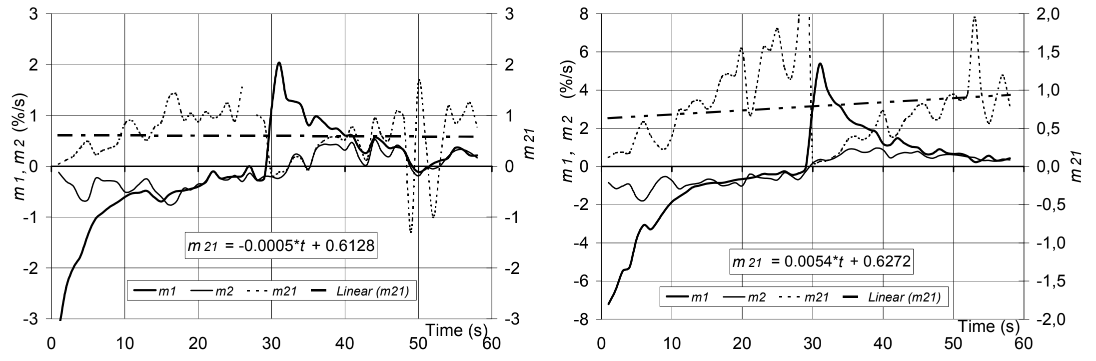
a) b)

Figure 3. Recorded data for weak heating power, a) and for high heating power, b)

In these graphs, the resistance variations and their time delay can be observed. For the detection thermistor, an exponential resistance variation is visible, while the compensation thermistor seems to have a linear variation for the recording time of 60 s. By using a fitting algorithm with an exponential function, the thermal time constant τ_{Th} for the detection thermistor results in a mean value of 9 seconds. Other relevant and important parameters as the relative time variation of the thermistor resistance, m_1 and m_2, as defined by the next equations, were obtained by data processing:

$$m_1 = \frac{100}{R_{Th1}}\frac{dR_{Th1}}{dt} \quad [\%/s] \quad \text{a)}; \quad m_2 = \frac{100}{R_{c1}}\frac{dR_{c1}}{dt} \quad [\%/s] \quad \text{b)}; \quad m_{21} = \frac{m_2}{m_1} \quad \text{c)} \tag{5}$$

Fig. 4 shows the possibility to measure the sensing delay for the compensation thermistor and the possibility to have a constant m_{21} for relative variation transfer coefficient. The m_1 and m_2 curves cross the time axis in different two points around of 30 seconds. This means that a sensing delay t_d, resulted less then 2 seconds, exists. The trend for relative transfer coefficient is quite independent from the time, *t*, and its constant value is around 60 percents at all applied test *dc* powers.



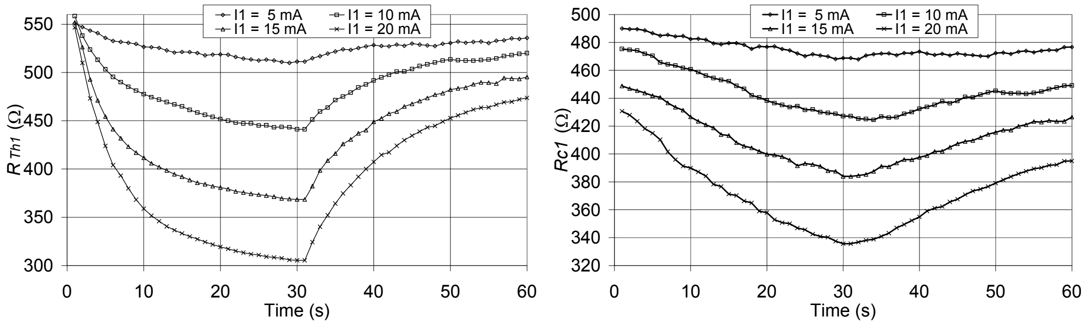
a) b)

Figure 4. Slope of the rezistances variation *vs.* time for I_1 = 10 mA, a), and I_1 = 20 mA, b)

The relative power sensitivity coefficients, S_{PTh1} and S_{PRc1}, defined by the equations:

$$S_{PTh1} = \frac{100}{R_{Th1}} \frac{dR_{Th1}}{dP} \quad [\%/\mathrm{mW}] \quad \text{a)}; \quad S_{PRc1} = \frac{100}{R_{c1}} \frac{dR_{c1}}{dP} \quad [\%/\mathrm{mW}] \quad \text{b)} \tag{6}$$

are obtained by other data analysis, as that represented in the next two graphs.



a) b)

Figure 5. Resistance *vs.* time for thermistor power sensor a), and for the compensation thermistor, b)

Mean values of - 0.00314 Ω/Ω/mW for S_{PTh1}, - 0.00186 Ω/Ω/mW for S_{PRc1} and the same ratio of 0.6 between these relative sensitivity were obtained. At the balance, the expected cross-sensitivity of the compensation thermistor will be – 0.37 Ω per mW dissipated in the detection thermistor.

V. Conclusions

Characterization of both thermistor elements for a commercial power sensor has been done with the same instrumentation set-up normally used for absolute power measurements. This characterization is necessary for implementing the digital balancing loop on which our new calibration method is based. The full implementation of the new calibration goes further on the INRiM coaxial microcalorimeter with the aim to simplify the primary power standard realization. The proposed arrangement will allow microcalorimeter measurements and calibration in the same time and will be an alternative solution to that one based on extensive data corrections [3].

References

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