Low noise temperature control: application to an active cavity radiometer

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Abstract

We have designed low noise temperature sensing and control units with the objective of using them for the fabrication of far infrared active cavity radiometers. The sensing unit, first characterized at 300 K using industrial platinum resistance thermometers, has a noise level of \( \sim 25 - 30 \mu K_{\text{eff}} \) for a 3 hours measuring time and in a 1 Hz bandwidth. Using YBCO superconducting thermometers, the noise level goes down to \( 2.5 \mu K_{\text{eff}} \), which is strongly limited by excess 1/f noise in the YBCO film at the superconducting transition. The sample holder used in the 90 K experiments is built with an auxiliary heating resistor, which enables an easy and accurate identification of the electrothermal model, even in the closed loop operation. Based on a design previously published by NIST, we estimate from these experimental results that the overall noise limitations of radiometers could be lowered by one order of magnitude.

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1 INTRODUCTION

The performances of electrical-substitution radiometers (ESR) depend mainly on low noise temperature control. Such instruments have been developed for various radiometric measurement purposes, including, for example, the measurement of the Stefan-Boltzmann constant\textsuperscript{1,2}, and the long-term monitoring of the solar irradiance\textsuperscript{3}, and are used as primary standards of optical power at national standards laboratories\textsuperscript{4,5}. In these experiments, an optical absorber-receiver is thermally isolated from a temperature regulated heat sink by a weak thermal link and is alternately heated by the unknown radiant power and by the resistive heater (Joule heating). The incident radiant power entering the aperture of the absorber-receiver can be deduced from the measurement of the equivalent electrical power needed to ensure the same temperature rise for the two heating effects. The resolution of the ESR was so far mainly limited by the temperature stability of the receiver. Thus, two temperature controllers, one for the heat sink and the other for the receiver, have been proposed to improve the ESR\textsuperscript{6}. Our work is inscribed in this framework, and we report in this paper on the progress we made on the temperature control of the heat sink using a low noise switched electronics and an optimized analog PID. Furthermore a heat perturbation method is presented to clearly specify the long time limitation of the system.

This paper is arranged as follows. Section 2 describes the experimental setup and preliminary measurements at room temperature. The third section relates a set of improvements and measurements at liquid-nitrogen temperature. In section 4, the perspectives for the ESR performance improvements are presented.
2 READ-OUT ELECTRONICS AND EXPERIMENTAL SETUP AT ROOM TEMPERATURE

2.1 Read-out electronics for a standard platinium resistor

The fundamental Johnson noise limitation of a metallic thermistor such as an Industrial Platinium Resistance Thermometer (IPRT), 100 Ω thermistor with $\alpha = 1/R \times dR/dT = 3.85 \times 10^{-3}$ K$^{-1}$, leads to an rms value of the noise equivalent temperature (NET) of $\sigma = 4.2 \mu K_{\text{eff}}$ within a 1 Hz bandwidth (first order) at 1 mA bias current, or equivalently, 26 $\mu K_{\text{pp}}$ around an ideally stable thermostat at room temperature. The required resolution for the read-out electronics is in the $10^{-7}$ range (i.e. $7^{1/2}$ digits or 25 bits resolution). The home made read-out electronics presented here was designed to be within a factor 2 of such resolution at low cost. The read-out electronic system is presented in Fig. 1.

An elementary Wheatstone DC-bridge (or AC-bridge) is often used in resistance measurements, and lead resistances can be compensated with special bridge connections, as in the “Mueller bridge”\textsuperscript{7}. To improve the temperature measurement, different ways have been investigated in the literature, which decrease the noise equivalent temperature of the system\textsuperscript{8}. Our circuit is derived from low-noise electronics designed for astronomical bolometric measurements. It consists of a square wave bias current source, a low temperature-coefficient resistance, a low-noise preamplifier stage, a lock-in amplifier and a $2^{nd}$ order Butterworth filter. The periodic bias current avoids some contamination of the useful signal by low frequency electrical noise. The square wave bias is used in order to keep constant the electrical power dissipated in the thermistor, unlike the sine wave bias\textsuperscript{9}. Consequently, the temperature of the thermistor does not oscillate. Indeed, in the sine wave bias case, a modulated applied power, at $2f_o$, is dissipated in the thermistor. Moreover, in order to avoid parasitic harmonic components, the modulation frequency $f_o$ has to be higher than the bolometer thermal cutoff frequency $f_{th}$ ($f_o \gg f_{th}$).

The high precision voltage reference AD588 was chosen because of its low output
noise, typically $6 \mu V_{pp}$ over the 0.1 Hz to 10 Hz frequency ranges, of its long-term stability at ambient temperature ($\sim 15$ ppm/1000 hours) and of its low output voltage drift ($\pm 2$ ppm $K^{-1}$). It was configured to provide a $\pm 5V$ reference voltage, noted $\pm V_{Ref}$. A switching electronics follows the voltage reference circuit and then provides the desired square wave voltage at frequency $f_o$ around a few kHz. As shown in Fig. 1, the bias voltage is applied to two high precision resistive dividers, one of which includes the high sensitivity thermistor $R_{th}$, and the other includes a low TCR resistance $R$ ($\leq \pm 50$ ppm $K^{-1}$, a few tens of ohms). It has to be noted that the electrical power dissipated in the thermometer has to be much lower than $1$ mW in order to limit self heating effects. Additionally, the value of the injection resistor, $R_l$, has to be very large compared to the thermistor resistance $R_{th}$ in order to form a current source. This injection resistor was fixed at $6$ k$\Omega$ thus leading to a $0.83$ mA current through $R_{th}$ ($\sim 100$ $\Omega$ impedance) and $R$.

An injection resistor was preferred to a capacitance even if the capacitance is sometimes preferred because it does not generate Johnson noise. In that case, the capacitance is fed using a triangular wave voltage, which induces a square wave current. This triangular wave voltage may be obtained using an active integrator circuit. Although no Johnson noise takes place through the capacitance, the system stability is dependant on the capacitance and the use of an additional amplifier may also increase the noise level. In order to ensure stability together with the low noise requirement, we therefore decided to use a square wave voltage through the low value thermistor and the low TCR reference resistor $R$.

The preamplifier stage is made using an instrumentation amplifier (SSM2017) followed by a lock-in amplifier. The reference signals are applied to the lock-in amplifier built around an AD620 and JFET switching connections. Finally, a Butterworth filter (2nd-order) shapes the useful output signal of the battery-operated power device. Using the $\alpha$ value of an IPRT, the voltage-to-temperature responsivity of the system at the input $R_s$ is then $320 \mu V$ $K^{-1}$. The white noise around the frequency $f_o$ is mainly dominated by the Johnson noise of $R_{th}$ and $R$. It can be estimated as $\epsilon_n = \sqrt{2} \sqrt{4k_B T R}$ with the
$R$ value chosen close to $R_{th}$ at the working temperature. Adding the amplifier noise, we obtain the equivalent noise source $e_n \simeq 2 \text{nV Hz}^{-1/2}$ at room temperature. It follows that the ultimate noise equivalent limit of our system is $T_n = 6.3 \mu\text{K Hz}^{-1/2}$ (i.e. an rms value $T_{eff} = \sqrt{\frac{2}{T}} T_n \simeq 8 \mu\text{K}_{eff}$ in a 1 Hz, and a first order bandwidth), twice the ideal value for an IPRT.

2.2 Read-out electronic characterization

2.2.1 Noise of the system

First, a study of the read-out electronic system where the thermistor was replaced by an other low TCR resistor has been carried out. The system exhibited an excess low frequency noise regim close to a “1/f” one as clearly seen in the comparative Fig. 1(c). Then, the deduced noise equivalent temperature spectrum (NET) would be comprised between $6/\sqrt{f} \mu\text{K Hz}^{-1/2}$ and $10/\sqrt{f} \mu\text{K Hz}^{-1/2}$ if an ideal IPRT was used at 0.83 mA. The associated rms value is in the range $16 - 30 \mu\text{K}_{eff}$, 2 times higher than expected in Sec. 2.1 in the white noise regime. Deviation from the white noise and the 1/f law in the read-out electronic system could be due to the electronic system temperature that slowly drifts according to the surrounding laboratory temperature. The temperature compensating coefficients were estimated by measuring the electronic boards temperature during heating or cooling the whole electronic systems. Temperature coefficients of the three read-out electronic systems were found to be: $15.1 \pm 3.6 \times 10^{-4} \text{V K}^{-1}$, $5 \pm 2 \times 10^{-4} \text{V K}^{-1}$ and $-7.9 \pm 1.4 \times 10^{-4} \text{V K}^{-1}$, respectively. These coefficients are consistent with the observed output drift. Temperature drifts in the range of $40 \mu\text{K s}^{-1}$ at the PC board level were then estimated. They must be avoided or compensated to minimize the output drifts, which are known to lead to an excess noise differing from a 1/f regime, with a higher $1/f^n$ power.
2.2.2 Room temperature measurement tests

The temperature-coefficient resistance (TCR), noted $\alpha$, and the dimensionless slope of the thermometer $A$, defined as $A = d \ln R / d \ln T = \alpha T$, are generally used to compare different thermometers operated at different temperatures. The Standard Platinum Resistance Thermometer (SPRT) is used to define the ITS90 (International Temperature Scale of 1990) between the triple point of hydrogen (13.8023 K) and the freezing point of silver (1234.94 K) with an accuracy of $\pm 2.0 \times 10^{-3}$ K to $\pm 7.0 \times 10^{-3}$ K. However, the strain-free design of a SPRT limits its use in controlled laboratory conditions.

In the industrial platinum resistance thermometer (IPRT), the platinum wire is encapsulated within a ceramic housing or a thick film sensor is coated onto a ceramic surface. The protection from the environment is increased with a metal sheath. A Corrige IPRT was used in this present work. This class A device is characterized by an accuracy of $\pm 0.15$ K at 273 K and an average TCR of $3.85 \times 10^{-3}$ K$^{-1}$ (European standard) between 273.16 K to 373.16 K (according to DIN CEI 751 norm). This sensing element has a 100 $\Omega$ impedance at 273.16 K (triple point of water). The room temperature setup includes three IPRT thermometers radially set out on a good heat conductor plate (copper), several copper heat shields, and a heater resistance $R_H$. A coolant (oil) is used to minimize temperature gradients across the system. A correlation study between two temperature measurements was then possible using this arrangement. The most important parameters of merit of the thermometers we used are gathered in Table 1.

The temperature measurements were made with two read-out electronic systems, each using an IPRT as thermistor. The third IPRT was measured using the conventional four-wire method resistance measurement, performed with an HP3034A multimeter ($6^{1/2}$ digits, 2 s integration time). It is hereafter called the reference IPRT thermometer. Responsivity estimations at the output $R_o = \partial V_{out} / \partial T$ were made using the slow dc drift of the temperature cell, assuming that the reference IPRT thermometer is well calibrated (see Fig. 3). Note the good temperature stability: the temperature drift $\Delta T$ was only 25 mK for 12000 s measurement. Output responsivity for each read-out electronic system was about 10 V K$^{-1}$. 
A fluctuation estimation has been made using a polynomial fit up to the fourth order to remove the main part of the cell temperature drift. The result is shown in Fig. [3]. The fitting operation explains the quasi-periodic temperature artifacts evolutions, but the fact that the three responses are still coherent between themselves on Fig. [3] indicates that the resolution of the thermometers is better than the amplitude of the artifact modulation, i.e. $\sim 200 \mu K$. To better estimate the resolution of the high sensitivity read-out, we made a second fit limited to 2000 s. A result is shown Fig. [4]. This numerical analysis shows that our read-out system, connected to standard IPRT is compatible with a temperature resolution of $25 \mu K_{eff}$ over a 2000 s integration time. This value appears to be rather higher than that what was expected from the theoretical white noise level as introduced in Sec. 2.1 ($8 \mu K_{eff}$). This is due to the excess low frequency noise below 1 Hz. We note that the final estimation of noise is consistent with the value estimated in subsection 2.2.1 after measurements with a low TCR resistor. This shows that the excess noise of the IPRT is likely not seen in this frequency range. Finally we also note that for a short integration time (< 1s), the resolution obtained is close to the theoretical one below $10 \mu K Hz^{-1/2}$. Such values (on short integration time) were also obtained by Dupuis et al.\textsuperscript{14}, but for 100 kΩ thermistors, the intrinsic voltage noise of which were 31.6 times higher than the 100 Ω thermistor used in this study.

In order to reduce the temperature drifts during long time measurements, another experimental setup at liquid-nitrogen temperature was built in association with a home-made optimized PID controller which will be described in the next section.

### 3 LOW TEMPERATURE SAMPLE HOLDER AND EXPERIMENTS

#### 3.1 Sample holder design

The sample holder depicted in Fig. [5] was designed in order to improve long time temperature measurements by introducing an optimized temperature control. It is known
that the temperature control of massive systems, even at temperature as low as 77 K, can be made difficult because of the inherent time delay between the heat production (or heat removing) and the temperature rise (or temperature decrease) of the sample. It follows that the overall gain of the servo loop cannot be made arbitrarily large in order to ensure a sufficient stability (Nyquist criterion). The present sample holder will serve as the temperature controlled heat sink of a future active cavity radiometer (ACR). Its mass must be much higher than the receiver one, say $\sim 10^3$, as in the NIST (National Institute of Standards and Technology, USA) prototype, in order to act as a heat sink. We chose a mass of 20 mg for the receiver and 20 g for the sample holder. In order to minimize the heat travel between heating resistor and the thermometers glued on the copper plate, the heating resistor, noted $R_{H_1}$, was wound in a spiral grooved at the rear of the plate. The resulting delay $\tau_D$ is 0.4 s. A second heating resistor, noted $R_{H_2}$, was wound, intertwined and identical to the first one. It constitutes an easy and reproducible way of applying an accurate heat perturbation to the system, either in open loop or closed loop configuration. The use of this auxiliary input led to a very convenient identification of the parameters of the servo-loop models and greatly helped the full characterization of the system. A large perturbation in the open loop configuration was then applied in order to determine the thermal model parameters gathered in Table 2. They are those of a first order thermal circuit with a small delay. Using these values and the signal characteristics of our read-out electronic system, we built a SPICE model in order to derive a set of parameters for the PID controller feedback circuit.

3.2 Full system characterization and performances

The schematic of the experimental arrangement used to fully characterize the temperature control of the sample holder is reported in Fig. 6. We used three high TCR resistors made using a 200 nm thick high quality $YBa_2Cu_3O_{7-\delta}$ (YBCO) film, patterned in $40 \times 600 \ \mu m^2$ strips. Their $R$ versus $T$ characteristics and their derivatives are plotted in Fig. 7. They exhibit the same critical temperature and very similar shapes. One of
these thermometers was used to sense the temperature in the servo loop, the two others giving two independent observations of the temperature evolution of the copper plate. Finally, a heat perturbation could be easily applied by means of the second heating resistor $R_{H2}$. Data plotted in Fig. 8 show examples of this perturbation method with a large signal applied on $R_{H2}$. Decreasing the voltage oscillation amplitude enables the recording of small temperature oscillations in the two high-sensitivity thermometers as shown in Fig. 8. These recordings, performed in longer measuring times, were used to calculate the noise spectra of our system. The frequency doubling in Fig. 8 is of course associated with the squaring of the applied voltage producing the Joule heating $V^2/R_{H2}$. From these data, it is easy to extract the rejection efficiency of the servo loop at these frequencies for the heat perturbations coupled to the sample holder. We find a value of 400 at DC. This means that the stability of the sample holder temperature is improved by $\sim$ 400 by closing the loop, and because the temperature drifts can be 70 $\mu$K s$^{-1}$ in open loop, we are expecting a temperature stability of 0.175 $\mu$K s$^{-1}$.

Moreover, the YBCO thermistors have a much higher sensitivity than the platinum ones, as shown in Table 1, leading to a responsivity $\Re_t \sim 42$ mV K$^{-1}$ at 0.83 mA bias current. It follows that the temperature drifts occurring on the PC board will have a relative effect $\sim$ 200 times lower than that obtained using an IPRT. Then, considering the Johnson noise of the YBCO thermistors only, the ideal noise floor would be about 10 nK$_{eff}$ in a 1 Hz bandwidth. If the read-out electronics is not ideal, the noise floor would be 30 nK$_{eff}$. However, the overall performances are limited by the excess low frequency noise of the YBCO thermistors, the level of which is generally dependent on microstructural properties. Fig. 10 shows a recording at the output of two high sensitivity thermometers during a measuring time of 400 s. They are clearly correlated, and show a mean temperature drift of 87 nK s$^{-1}$. This latter value is consistent with the expected one, because the temperature fluctuations and drifts of the liquid-nitrogen bath are likely associated with the fluctuations and drifts of the atmospheric pressure with a maximum rate of the order of 1 Pa s$^{-1}$ on windy days and a conversion coefficient of 83 $\mu$K Pa$^{-1}$ from Clapeyron’s law$^{15}$. Finally, using the two simultaneous records of Fig. 10, we also plotted the instantaneous difference between the two thermometers,
in order to get an estimation of the actual rms noise of the thermometers. We have got a standard deviation $\sigma = 2.8 \, \mu K_{\text{eff}}$ during this integration time. This value is much higher than the ideal one obtained using only the Johnson noise assumption. It clearly indicates that the noise process in our YBCO samples, at the superconducting transition, is dominated by excess low frequency fluctuations of the resistance. The associated spectrum has been plotted in Fig. 11. Note that the calibration of this spectrum was conveniently made using a reference signal at 5 mHz obtained with a known applied heat perturbation. Its amplitude was $95 \, \text{mV}$, producing a temperature oscillation of the sample holder of $30 \, \mu K_{pp}$ in closed loop operation (see Fig. 3). However, despite this large excess noise, we show in the following section that these performances are still useful to design an ACR.

4 ESR PERSPECTIVES

The achieved low noise temperature control could be used to regulate the temperature of the heat sink of the ESR. The NIST prototype described by Rice et al.\textsuperscript{10} will be considered below as an example. In order to estimate the performances, we will consider the same conditions as those used in the NIST prototype. The ESR cavity is supposed to receive radiant flux ranging from the microwatt level to the milliwatt, coming from an extended-source blackbody at $T = 300$ K. It was shown that the measurement could be done with a contribution to the standard random uncertainty below $20 \, \text{nW}$. As explained in Sec. 3.1, we reduced the mass of the heat sink, which implies the reduction of that of the receiver down to $20 \, \text{mg}$. This condition is thought to be very well fulfilled using a plane membrane receiver and an integrating sphere above it. To derive our following noise calculations, we assume that the incident radiation $P_i$, to be converted in electrical power $P_{CR}$ in order to keep the receiver at constant temperature, is chopped at a frequency $f_o = 10 \, \text{Hz}$. The simple electrothermal circuit we used to describe the effects of thermometer noise on the receiver is reported in Fig. 12 where $G_{ESR}$ and $C_{ESR}$ are the ESR thermal conductance and thermal capacitance, respectively. We introduced the associated noise sources to three contributions having roughly the same
order of magnitude:

- $T_n(f)$, which is the spectral density of the temperature fluctuations of the heat sink. It is shown from $10^{-4}$ Hz to $10^{-1}$ Hz in Fig. 11(b). Above $10^{-1}$ Hz the servo loop of the heat sink is not suitable anymore, and we just extrapolated the quasi $1/f^2 T_n^2(f)$ dependence. The spectral density $T_n(f)$ can be roughly fitted by $T_n(f) = 8 \times 10^{-7} f^{-0.9}$ K Hz$^{-1/2}$ from Fig. 11, we estimated its value at 10 Hz to be about 100 nK Hz$^{-1/2}$.

- $T_{n,ESR}(f)$, which is the spectral density of the YBCO thermometer attached to the ESR. It would act as the error detector of the servo loop driving the ESR cavity. Because we intend to use the same material and detecting electronics as that of the heat sink, we adopt the same spectral density $T_{n,ESR}(f) = T_n(f)$, which should hold up to about 100 Hz, before joining the Johnson level. As a matter of fact, recordings of the thermometer output, using a wide opening (300 Hz) of the lock-in amplifier unit, led to this spectral density.

- $T_{n,A}(f)$, which is the equivalent noise associated with the amplifier. At 10 Hz, a reasonable value is given by the ratio of the voltage noise $e_{nA} \simeq 1$ nV Hz$^{-1/2}$ of the amplifier to the thermometer responsitivity $R_t \simeq 42$ mV K$^{-1}$ at 0.83 mA bias.

Furthermore, in the circuit of Fig. 12, $S$ represents the thermometer sensitivity including gain (in V K$^{-1}$). We assume $S = 100$ V K$^{-1}$ and AOP is an operational amplifier, closing the servo loop and feeding the heating resistor $R_H$ of the ESR cavity. Assuming the AOP ideal, the inverting operational amplifier input writes $V_- = V_{REF}$, implying the thermometer circuit input to be $V_{REF}/S$. Standard circuit analysis leads to:

$$T_{ESR}(f) = \frac{T_n(f) G_{ESR} + P_{in} + P_{CR}}{G_{ESR} + jC_{ESR}\omega}$$ (1)

where the heat flow between the thermometer and the ESR body is neglected. Finally, The “Kirchhoff” law gives:
The deduced feedback power is written in two parts:

\[ P_{CR} + P_m = (G_{ESR} + jC_{ESR}\omega) \frac{V_{REF}}{S} \]  \hspace{1cm} (3)

is the ideal part, if we consider that there is no noise, and:

\[ P_{CR,n} = -G_{ESR}[(T_{n,ESR}(f) + T_{n,A}(f))(1 + j\tau_{ESR}\omega) + T_n(f)] \]  \hspace{1cm} (4)

is the noisy feedback power.

In the bandwidth of the ESR any change in \( P_m \) will induce a change in \( P_{CR} \) in order to keep the receiver temperature constant and \( P_{CR} \) accounts for the output signal. The second part accounts for the noise on the feedback power, the minus sign indicating that this part of the feedback power reacts so as to balance the various noise sources. Eq. 4 is deduced from a small signal analysis applied to Eq. 2, i.e. by nulling its right member, and to Eq. 1, i.e. with \( P_m = 0 \). Assuming the noise sources to be uncorrelated we get the final spectral noise density of \( P_{CR,n} \):

\[ |P_{CR,n}| = G_{ESR} \times \sqrt{T_n^2(f) + \frac{T_{A,n}^2(f) + T_{n,ESR}^2(f)}{1 + j\tau_{ESR}^2\omega^2}} \]  \hspace{1cm} (5)

We may now use the values attached to the noise sources to estimate whether or not we fulfill the ESR requirements. For clarity, we examine independently the effects of \( T_n^2(f) \) and \( T_{n,ESR}^2(f) + T_{n,A}^2(f) \). The numerical values are calculated at a chopping frequency of 10 Hz, assuming a measuring time of 250 s with a first order post filtering. Before performing these calculations we need to know the numerical value of the
thermal conductance $G_{ESR}$. The latter is related to the dynamic range of the ESR, i.e., the largest input signal $P_{in,Max}$ that can be measured. In the chopped mode of the incoming power $P_{in}$, the feedback power will oscillate between the on-off states of $P_{in}$, leading to the condition of Eq. 3 to be fulfilled by $G_{ESR}$:

$$G_{ESR} = \frac{S \times P_{in,MAX}}{V_{REF}}$$

Choosing $P_{in,MAX} = 10\,\text{mW}$, $V_{REF} = 10\,\text{V}$, $S = 100\,\text{V K}^{-1}$ gives $G_{ESR} = 100\,\text{mW K}^{-1}$. Such a thermal conductance value can reasonably be achieved at 90 K using silicon micromachining techniques\textsuperscript{16,17}: for calculations we assumed a $10 - 30\,\mu\text{m}$ thick membrane of $3 \times 3\,\text{mm}^2$ area. It follows that the noise associated with that of the heat sink should be $G_{ESR} \times T_n(10\,\text{Hz})$ of the order of $10^{-8}\,\text{W Hz}^{-1/2}$, i.e., a standard deviation of the order 800 pW$_{eff}$ with a measuring time of 250 s. To estimate the noise associated with that of the receiver thermometer we need to know the receiver time constant. Because of the small active volume of the receiver and of the high thermal conductance value, the thermal time constant would be much shorter than that of the NIST prototype, about 0.1 ms at 90 K. We then deduce the order of magnitude of the associated noise to be close to $G_{ESR} \times T_n,ESR(10\,\text{Hz})/2\pi \simeq 1.6\,\text{nW Hz}^{-1/2}$, which means an rms value of 130 pW$_{eff}$ for 250 s observing time. We then conclude that these estimated values, although estimated using a rough model, are very promising, within an order of magnitude below the demonstrated values by Rice \textit{et al.}\textsuperscript{10}. Similar results were presented by Libonate and Foukal\textsuperscript{18}, reporting a root-mean-square noise level of 1.6 nW and a 30 s time measurement for an allowable input-power level of up to $\sim 2\,\text{mW}$. These experimental results are of the same order of magnitude as ours, calculated for an input-power level of up to 10 mW. No specific optimization in order to reduce the excess noise in our YBCO films was made, which explains why the noise level is rather large, if compared to other works\textsuperscript{17,19}. As shown by Neff \textit{et al.}\textsuperscript{17}, the use of a thin gold layer onto the YBCO layer can lead to very low noise equivalent temperature at 10 Hz. The use of such levels in our model would
have led to dramatic reduction of the noise floor of the sample holder, which means improvements in the ESR performances. We would like to point out that the use of a small membrane and an integrating sphere does not allow a proper primary radiometer operation, but the very short time constant would permit the use of a much higher shopping frequency. This means working conditions in the white noise domain of the YBCO sensors, implying a possible reduction of two orders of magnitude of the noise floor level, which appear still very attractive. Such a system would then have performances close to the liquid-helium-cooled ESR of Reintsema et al.\textsuperscript{20}. Finally, to eliminate the problem of the integrating sphere, we have done preliminary calculations in order to design a 1 cm aperture, 1.5 cm long pyramidal cavity in silicon, with wall thicknesses ranging between 5 to 50 $\mu$m. A high conductance value of $\sim 100$ mW K$^{-1}$ is feasible together with a time constant of the order of 0.1 s. Such a system would then be very attractive because it could include full-integrated heating resistors and thermometers. Thus correlation techniques and a perturbation heating method could be used as well to improve model identification and signal processing.

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Tables

Table 1: Parameters of merit of the studied thermometers: $T_o$ is the operating temperature, $\Delta T$ is the measurement temperature range, $\alpha$ is the temperature coefficient resistance (TCR) and $A$ is the dimensionless slope of the thermometer.

| studied thermometers | $T_o$ [K] | $\Delta T$ [K] | $\alpha$ [K$^{-1}$] | $A$ [dimensionless] |
|----------------------|-----------|----------------|---------------------|---------------------|
| IPRT sensor          | 273.16    | 13.8023 to 1234.94 | 0.00385            | $\sim$ 1           |
| $YBa_2Cu_3O_{7-\delta}$ (typical values) | around 90 | few K | 1 to 5 | 90 to 450 |
| (present work)       | 90.25     | 2 – 3 | 1.7    | 150     |

Table 2: Thermal parameters and heating resistors for the liquid-nitrogen temperature measurement setup.

| parameter                                      | symbol | value       |
|------------------------------------------------|--------|-------------|
| mean thermal time constant of the sample holder| $\tau_{Th}$ | 100 s       |
| thermal conductance between heat sink and copper plate | $G_{Th}$ | 77 mW K$^{-1}$ |
| delay between applied electrical heating and temperature plate | $\tau_D$ | 0.4 s |
| main heating resistor                          | $R_{H1}$ | 30 $\Omega$ |
| auxiliary heating resistor                     | $R_{H2}$ | 30 $\Omega$ |
FIGURES CAPTIONS

FIG. 1. Read-out electronic system: a square wave voltage is applied to two voltage dividers, one includes the thermistor $R_{th}$, the other includes a low TCR resistance. A true differential amplifier (SSM2017) followed by a lock-in amplifier and a Butterworth filter (2nd-order) shapes the useful output signal of the battery-operated power system.

FIG. 2. Simultaneous temperature evolution of two read-out electronic systems (curves b and c), each using an IPRT sensor as thermistor: the slow temperature drift enables the evaluation of the responsivity $R$. A reference IPRT thermometer (curve a) is used to calibrate the temperature deviation.

FIG. 3. Temperature measurements after the substraction of the main temperature drift, showing the correlation between the three thermometer responses: read-out electronic systems, each using an IPRT as thermistor (symbols), and a smoothing fit of reference IPRT thermometer (line).

FIG. 4. Fluctuation temperature during 2000 s after a linear fit, measured with two read-out electronic systems each using an IPRT as thermistor. A standard deviation $\sigma$ of 25 $\mu$K is demonstrated (for system II). A 100 $\mu$K maximal temperature difference is observed between the two thermometer read-outs.

FIG. 5. The liquid-nitrogen temperature measurement setup: a reference IPRT thermometer and YBCO thin film thermometers are glued on the copper sample holder. Two heating resistors $R_{H_{1,2}} = 30 \Omega$ were made using constantan wires and distributed evenly across the sample holder cross-section.

FIG. 6. Schematic of the home-made PID controller: the three boxes named P, I and D depict the functions proportional, integral and derivative, respectively. The temperature measurements were made with three read-out electronic systems, each using an YBCO thermometer: one was used for regulation purpose (thermistor C) and two for correlation investigations (thermistors A and B). Two heating resistors $R_{H_{1,2}} = 30 \Omega$ were made using constantan wire.

FIG. 7. Resistances $R$ (closed symbols) and their derivative $dR/dT$ (opened symbols).
as a function of temperature $T$ for three high resolution YBCO thermometers at 1 mA bias.

**FIG. 8.** A sine wave voltage, called perturbation heating, is applied to the heating resistor $R_{H_1}$ (curve a). The PID output reacts as shown in curve b (note the frequency doubling). The resulting temperature deviation is measured by the two read-out electronic systems, each using an YBCO thermometer (curves c and d, right axis), which show very similar responses.

**FIG. 9.** Thermometric responses of the read-out electronic systems associated with a low temperature oscillation provided by a 5 mHz perturbation heating ($\sim 95$ mV amplitude) applied to the sample holder.

**FIG. 10.** Temperature measurement without perturbation signals using two read-out electronic systems, each using an YBCO thermometer: the two temperature deviation are plotted (left axis). The instantaneous difference (right axis) gives an estimation of the actual rms noise of the thermometers. For a 400 s measurement, a standard deviation equivalent of $2.8 \mu K_{eff}$ is obtained.

**FIG. 11.** Input spectral noise density $e_n$ and noise equivalent temperature (NET) $T_n$ for a read-out electronic system without thermistor (curve c) and with an YBCO film (curve b) as well as the theoretical white noise value (curve d) are plotted. The NET of the reference IPRT thermometer is reported for comparison (curve a on right axis only).

**FIG. 12.** Simple electrothermal circuit used to describe the effects of thermometer noise on the receiver: $G_{ESR}$ et $C_{ESR}$ are the ESR thermal conductance and thermal capacitance, respectively, $P_{in}$ and $P_{CR}$ are the incident radiation and the electrical power, respectively. $T_n(f)$ is the spectral density of the temperature fluctuations of the heat sink. $T_{n,ESR}(f)$ is the spectral density of the YBCO thermometer attached to the ESR. $T_{n,A}(f)$ is the noise equivalent temperature associated with the amplifier.
Figure 1:
Figure 2:
Figure 3:
Figure 4:
Figure 5:
Figure 6:
Figure 7:
Figure 8:
Figure 9:
Figure 10:
Figure 11:
Figure 12: