

NOISE THERMOMETRY WITH COMMERCIAL SQUIDS

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Abstract

The design and construction of a noise thermometer built around two commercially available SQUID systems is presented. The thermometer is based upon the direct-coupled system of Webb, Gifford and Wheatley but uses a novel digital squarer/integrator to analyse the noise voltage. Preliminary results demonstrate the performance of the thermometer.

INTRODUCTION

Various authors have documented the use of Johnson noise thermometry down to the millikelvin temperature range.¹⁻⁸ This device can be used as an absolute thermometer, without calibration if all of the circuit parameters are measured independently,² or with a calibration point. The noise thermometer can also be used as a measure of the electron temperature in a metal resistor for a variety of hot electron experiments. One such experiment has been reported.⁹ The design of the thermometer involves a careful compromise in order to achieve a low device noise temperature, and high precision with a reasonable averaging time. In this paper we discuss the design criteria for a noise thermometer with two commercially available SQUID systems, and describe a digital circuit design for real-time data analysis.

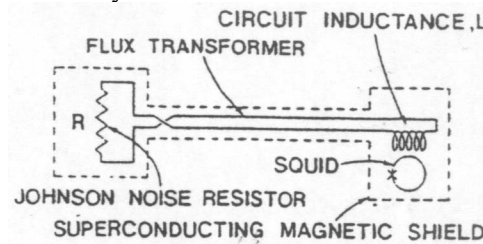


Fig. 1. The Johnson noise thermometer at temperature T, coupled directly by a superconducting flux transformer to a r.f. SQUID at 4.2 K.

The thermometer we have constructed is the direct-coupled system described by Webb, Gifford, and Wheatley² and illustrated in Figure 1. The Johnson noise,

$$\langle V^2 \rangle = 4K_B T R \Delta f \quad (1)$$

drives a mean square current

$$\langle I^2 \rangle = 4K_B T R \Delta f / (R^2 + \omega^2 L^2) \quad (2)$$

in the input circuit, where Δf is the bandwidth, R is the resistance of the Johnson noise resistor and L is the total effective inductance of the input circuit. In turn the mean square flux generated in the SQUID is

$$\langle \phi^2 \rangle_J = S^2 \langle I^2 \rangle \quad (3)$$

where S is the current sensitivity (typically $10 \phi_0 / \mu A$ where $\phi_0 = h/2e$ is the flux quantum). In addition to the flux due to Johnson noise in the resistor there will be device noise due to a variety of causes. This can be expressed as,

$$\langle \phi^2 \rangle_D = n^2 \Delta f \quad (4)$$

where n is the device noise expressed in terms of $\phi_0 / \sqrt{Hz}$, typically $10^{-4} \phi_0 / \sqrt{Hz}$ for an r.f. SQUID and $\leq 10^{-5} \phi_0 / \sqrt{Hz}$ for a D.C. SQUID¹⁰. Since $\langle \phi^2 \rangle_J$ saturates as the bandwidth is increased, at $\Delta f \sim R/2\pi L$, whereas $\langle \phi^2 \rangle_D$ does not, it is usual to apply a low pass filter at $\omega_L \sim R/L$ to the output signal of the SQUID. With such a low pass filter, the effective device noise temperature, derived from equations (2) to (4), is given by,

$$\langle \phi^2 \rangle_D \sim S^2 4k_B T_D \Delta f / R \quad (5)$$

or $T_D \sim n^2 R / 4k_B S^2$

The device noise temperature is shown in Figure 2. for the two commercial SQUIDs which have the parameters given in Table 1.

Table 1. Parameters of the Two Commercial SQUIDS

Model Number	CTF r.f. SQUID*	SHE D.C. SQUID ⁺
	SQP-200	DSQ
Flux Transformer Inductance	3.3 μ H	2 μ H
Sensitivity (Φ_0 / μ A)	10.8	10
Flux Noise ($\Phi_0 / \sqrt{\text{Hz}}$)	1.1×10^{-4}	1.5×10^{-5}

* Canadian Thin Films, Port Coquitlam, B.C., Canada.

⁺ S.H.E. Corp., San Diego, California, USA.

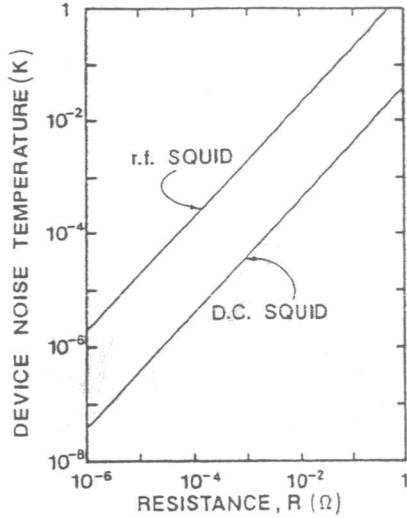


Fig.2. The device noise temperature, T_D , calculated from the SQUID flux noise, as a function of the Johnson Noise thermometer resistance, R .

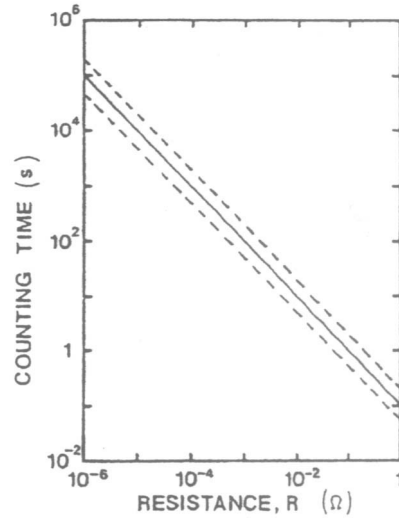


Fig.3. The approximate counting time, t , to obtain 1% precision in a temperature measurement as a function of the Johnson noise thermometer resistance, R .

The precision of the temperature measurement has been evaluated by Webb et al.² The result is dependent upon the setting of the low pass filter but for $\omega_L \sim R/L$, the relative uncertainty of the measurement is,

$$\delta T/T \sim 2(L/Rt)^{1/2} \quad (6)$$

where t is the averaging time. The approximate averaging time for 1% precision as a function of R , is shown in Figure 3. It is clear from equations (5) and (6) that the device noise temperature improves (decreases) with a decrease in R whereas the averaging time improves (decreases) with an increase in R . Any real design therefore involves a compromise.

DESIGN OF THE THERMOMETER

System Parameters

The initial use of our thermometer is for calibration of thermometers for a simple 100 mK dilution refrigerator and a device noise temperature of 5 mK was considered adequate. This set $R=10^{-3} \Omega$ for the r.f. SQUID, $L/R \sim 3.3$ ms. and $\omega_L \sim 300 \text{ s}^{-1}$. In turn the averaging time for 1% precision is ~ 100 s. For the D.C. SQUID, with the same resistance (R), the device noise temperature would be 50 μ K.

Low Temperature Assembly

The low temperature assembly is shown in Figure 4. The resistor was made of copper wire with the ends spot welded to niobium tabs which in turn were spot welded to two niobium wires. These wires were twisted to decrease self inductance and fed through a niobium tubular shield to the input coil of the SQUID. The SQUID was immersed in the 4K bath. The copper wire resistor was wound non-inductively and varnished with GE7031 to a copper former which was in turn thermally anchored to the mixing chamber of the dilution refrigerator by a thick copper wire. The resistor assembly was completely shielded from stray magnetic fields by a small soft-solder-coated brass chamber.

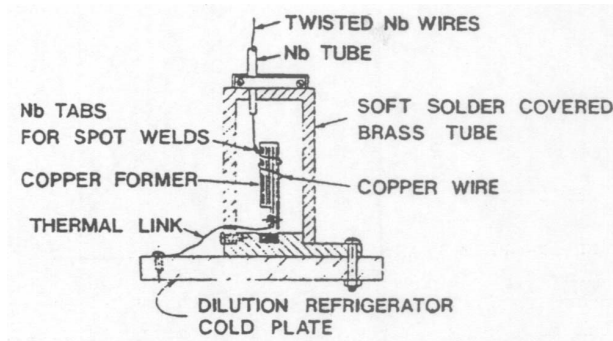


Fig. 4. The low temperature assembly consisting of the copper wire noise thermometer, the superconducting magnetic shield and the flux transformer.

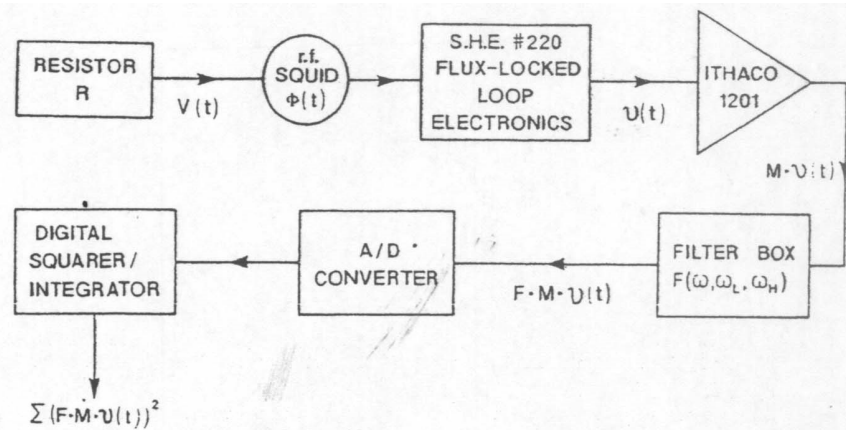


Fig. 5. Noise thermometer electronics

Data Acquisition

A block diagram of the measurement system for the r.f. SQUID is shown in Figure 5. The SQUID is operated with the SHE Model 220 flux-locked loop electronics module, producing an output voltage across a feedback resistor proportional to the flux in the SQUID. The signal is amplified with an Ithaco model 1201 low noise preamplifier, filtered with a combination of high and low pass fifth-order Butterworth filters and converted to a digital signal (at 104 kHz) with a Datel Systems, Inc. model ADC-EH12B3 A/D converter. This digital signal is then sampled at a variable rate (between 16 and 1024 Hz) and digitally squared and integrated. The squarer/integrator can square a 32-bit binary number and can accept an integral of 84 bits; at the maximum sampling rate of 1024 Hz and with a full 32-bit noise reading, the maximum integration time is 1000 s. Finally, the integral must be divided by the integrating time and a calibration constant to yield the temperature.

The design of the squarer/integrator is as follows: The digital number, to be squared and then added to the sum in the 84-bit memory, is first moved into 2 shift registers, of 32 and 64 bits, respectively. The number in the first register is used to trigger latches that add the number in the second register to the 84-bit memory. The operation proceeds in the following way. If the rightmost digit in the first register is a 1, the number in the second register is added to the sum. The number in the first register is then shifted right, losing one digit while the number in the second register is shifted left, adding a 0. Again, if the rightmost digit in the first register is a 1, the number in the second register is added to the sum. This is repeated for a total of 31 shifts. The result of the operation is that the square of the number is added to the sum although that square is not evaluated as a distinct number.

The shift registers, made up of 74198 8-bit registers, run at 250 kHz, under the control of a 1 MHz clock. The numbers in the 64-bit register and in the 84-bit memory are added with a series of 74LS83 4-bit adders and this new sum replaces the old sum by way of a 2-step latch process to avoid synchronization problems that could arise with a 1-step latch. The 84-bit latches were made up of 74LS174 6-bit latches. The 250 kHz shift speed allowed plenty of time for the square/integrate process for the maximum 1 kHz digital sample speed.

RESULTS

Figure 6 shows some preliminary results obtained with the CTF r.f. SQUID. The resistance R was $10^{-3} \Omega$. The high pass filter was $\omega_H = 1.2 \text{ s}^{-1}$, a range of low pass filters was used, and the count time was 100 s. The temperature, approximate only, was deduced from an uncalibrated Speer 220Ω carbon resistor. There was a problem with vibration in the cryostat which added noise in the frequency range above 120 s^{-1} and prevented the use of the most efficient low pass filter frequency, $\omega_L \sim R/L \sim 300 \text{ s}^{-1}$. Below 120 s^{-1} , the measured noise scaled with ω_L (or Δf) as expected for Johnson noise.

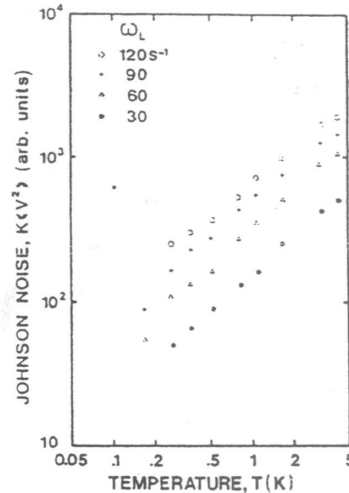


Fig. 6. Preliminary results of Johnson noise versus temperature for a range of low pass filter settings. The temperature scale is approximate only.

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