Johnson Noise Thermometry Measurement of the Boltzmann Constant With a 200 Ω Sense Resistor

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***Abstract*—In 2010, the National Institute of Standards and Technology measured the Boltzmann constant k with an electronic technique that measured the Johnson noise of a 100 Ω resistor at the triple point of water and used a voltage waveform synthesized with a quantized voltage noise source (QVNS) as a reference. In this paper, we present measurements of k using a 200 Ω sense resistor and an appropriately modified QVNS circuit and waveform. Preliminary results show agreement with the previous value within the statistical uncertainty. An analysis is presented, where the largest source of uncertainty is identified, which is the frequency dependence in the constant term a_0 of the two-parameter fit.**

***Index Terms*—Boltzmann equation, Josephson junction, measurement units, noise measurement, standards, temperature.**

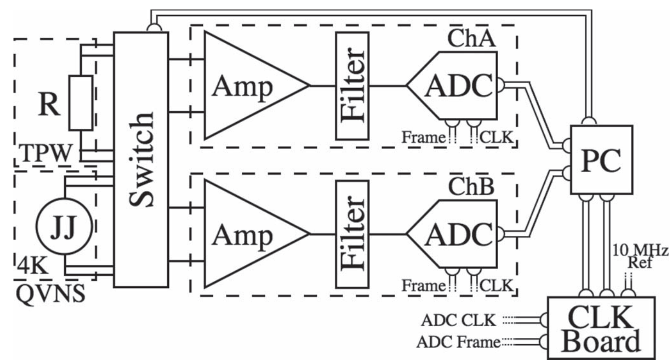

Fig. 1. Schematic diagram of the Johnson-noise two-channel cross-correlator.

I. Introduction

THE Johnson–Nyquist equation (1) defines the thermal noise power (Johnson noise) $\langle V^2 \rangle$ of a resistor in a bandwidth Δf through its resistance R and its thermodynamic temperature T [1], [2]:

$$\langle V_R^2 \rangle = 4kTR\Delta f. \tag{1}$$

Therefore, it is possible to obtain the value of the Boltzmann constant k by measuring the Johnson noise and the resistance value of a sense resistor at a defined temperature such as the triple point of water (TPW). The National Institute of Standards and Technology (NIST) goal of this experiment is to contribute to the redetermination of the Boltzmann constant by achieving a combined uncertainty [3], [4], below 5 μK/K, of the same order as that reached by acoustic gas thermometry [5].

Cross-correlation measurement techniques are required because the voltage-noise spectral density that has to be measured is very small (1.2 nV/$\sqrt{}$Hz) and of the same order as the voltage noise of the electronics used to amplify the signal [6]. NIST developed a Johnson-noise thermometer (JNT) system based on cross-correlation electronics that optimizes the measurement of the noise power of a sense resistor at the TPW [7]. The measurement electronics are calibrated by using a pseudonoise voltage waveform synthesized with the quantized voltage noise source (QVNS) that acts as a spectral-density reference [8], [9].

Fig. 1 shows the experimental schematic. The two channels of the cross-correlator simultaneously amplify, filter, and convert from analog to digital each of the two noise-source voltages. The switch-board relays alternately connect one of the noise sources to the input amplifiers. The acquired digitized voltages are optically transferred to the computer (PC), where the data are fast Fourier transformed, cross-correlated, averaged, and stored.

The QVNS is composed of two symmetric Josephson-junction lumped arrays with a low-inductance common ground. Each of the four output leads of the Josephson chip contains inductive coils that provide high impedance to the high-frequency bias pulses and a resistance whose value matches that of the sense resistor [10]. The chip is pulse driven, and the calibration signal obtained is intrinsically accurate because it is produced with perfectly quantized pulses that have time-integrated areas equal to the flux quantum $h/2e$, where h is the Planck constant and e is the electron charge.

The calibration signal is calculated and selected to closely match the sense-resistor noise power. The signal is a frequency comb of odd harmonics that are equally spaced in frequency. All the tones have the same amplitude but have random relative phases. The mean-squared power spectral density of the tones is equal to $\langle V_Q^2 \rangle = (h/2e)\ Q^2 N_J^2 M f_S$, where Q is a dimensionless factor used to match the tone amplitudes to the power spectral density of the sense resistor, N_J is the number of Josephson junctions, M is the bit length of the code used to drive the chip with pulses, and f_S is the clock frequency of the pulse generator. Waveforms of two different code lengths that produced frequency combs with correspondingly different tone densities and amplitudes were used. The first code, of length

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3×2^{23} bit, produced tones spaced at 800 Hz. The second code is eight times longer and has tones spaced at 100 Hz.

The use of the QVNS provides several advantages with respect to traditional Johnson noise thermometry. First, it allows the measurement of absolute temperature. Second, it enables the simultaneous matching of the noise power and impedance of the reference and the sense resistor [9].

With the QVNS-JNT system, the Boltzmann constant k is obtained from the ratio of the resistor to the QVNS-measured noise powers, the temperature T of the TPW, the parameters used to synthesize the calibration signal, and the measured resistance X_R expressed in units of the von Klitzing resistance [11], as follows:

$$k = \frac{\left\langle V_R^2 \right\rangle}{\left\langle V_Q^2 \right\rangle} \frac{hQ^2 N_J^2 M f_S}{16 T X_R}. \tag{2}$$

After improving the system by reducing nonlinearities, electronic distortion, and EMI coupling [12], [14], it was possible to obtain the first practical electronic measurement of the Boltzmann constant [15]. The value of k is obtained from fitting the ratio between the noise power spectra measured across the sense resistor and that measured across the QVNS with a two-parameter equation $a_0 + a_2 f^2$, where k corresponds to a_0 at zero frequency. This previous measurement was done by using a 100 Ω sense resistor, and we obtained a difference of 0.6×10^{-6} between the measured k and the 2006 CODATA value [16], with a relative combined total uncertainty $u(k) = 12.1 \times 10^{-6}$.

The 10 MHz reference signal is traceable to the NIST frequency standard. It synchronizes the QVNS clock with the cross-correlation sampling electronics. The clock board generates the signals and reference frequencies that synchronize the data transfer between the analog-to-digital converters (ADCs) and the computer. It also multiplies the 10 MHz with a phase-locked-loop chipset to produce the main 50 MHz clock (CLK). The CLK is divided with a field-programmable gate array that results in the 2.08 MHz sampling frequency (FRAME). A programmable delay line sets the delay of each FRAME signal in order to ensure synchronization between both ADCs. These signals are distributed through optical fibers in order to minimize electromagnetic interference in the analog electronics. The electronic apparatus is also described in detail elsewhere [7], [13], [14], [17].

II. 200 Ω Sense Resistor

In this experiment, we test the JNT measurement technique under different measurement conditions by doubling the sense-resistor value to 200 Ω to produce higher noise power, in the expectation that this will yield a lower statistical uncertainty for the same integration period or equivalently yield a decrease in the integration period to achieve the same uncertainty [18].

The use of a higher value sense resistor can affect the bandwidth and has the potential to generate an undesirable signal that has a quadratic frequency response. First, the higher resistance produces a larger RC time constant for the input circuit that decreases its bandwidth. For example, the 100 Ω sense resistor, the 60 pF JFET capacitance, and the 50 pF lead capacitance yield a 14 MHz cutoff frequency for that input circuit. Likewise, the 200 Ω resistor, the same JFET capacitance, and the 70 pF lead capacitance yield a 6 MHz input-circuit bandwidth. These cutoff frequencies are both larger than either the 2 MHz intrinsic bandwidth of the JFET preamplifiers or the 650 kHz cutoff frequency of the low-pass filters in the correlator amplifier chain. This latter value defines our measurement bandwidth.

Second, current noise that is intrinsic to the JFET in each preamplifier produces an undesirable voltage signal that is quadratic in frequency and increases with resistance [19]. For room-temperature resistances up to 1 kΩ, we have not observed any quadratic response in the autocorrelated signals that could be associated with a current-noise signal. We conclude that the current noise for our preamplifiers is much smaller than the white noise within the 2 MHz JFET bandwidth and that the minimal rolloff of the input circuit may allow us to successfully measure sense resistors of a few kiloohms.

III. Transmission-Line Matching

In order to measure the noise of the two times higher 200 Ω sense resistor, the voltage amplitude of the QVNS synthesized reference was increased by $2^{1/2}$ (by increasing Q) to match the resistor-noise power spectral density. Equalizing the power of the two noise sources is necessary to ensure similar behavior of the correlator electronics, such as gain and nonlinear response, during the measurement of both noise sources.

Different QVNS circuits were also required, namely, each output lead's series matching resistor was increased twofold to 200 Ω. The total lead resistance in the QVNS circuit must always be double that of the sense resistor, because the Josephson junctions act as a perfect voltage source whose short circuits essentially decouple the two cross-correlation channels. Fortunately, the superconducting integrated circuits were designed with this experiment in mind, because each lead resistor was constructed from four series-connected pairs of parallel 50 Ω resistors. To double the lead resistances, one resistor from each pair was mechanically opened.

In order to have well-matched transmission lines, the QVNS lead resistances must be within 1% of the sense resistance [10]. This required soldering chip resistors of a few ohms to each of the four output leads of the QVNS flex cryo package. Metal film resistors were used because their values do not change significantly when immersed in liquid helium. The QVNS lead resistors produce uncorrelated noise that must be compensated on the sense-resistor transmission line. The value of these compensation resistors is calculable from the differences between the autocorrelated resistor and QVNS signals of each of the two amplification channels. In this case, 2.2 Ω compensation resistors were mounted on each of the four sense-resistor leads.

The last fine-tuning necessary was to match the transmission-line impedances [9], [14]. This modification was necessary for several reasons. First, the new QVNS chip was mounted in a new probe that is longer than the one previously used. Second, the longer QVNS cable lengths and the different sense resistors changed the total impedance connected to the amplifiers.

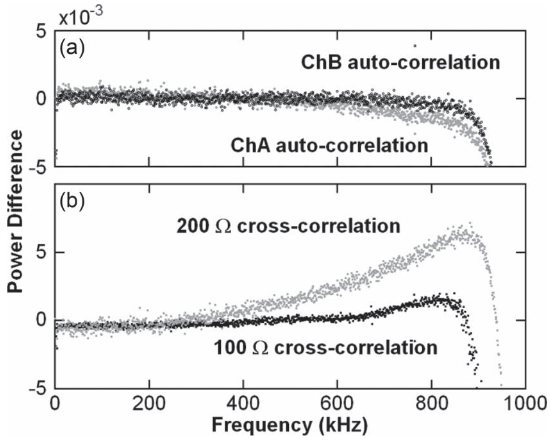

Fig. 2. Differences in power spectra $[\langle V_R^2\rangle - \langle V_Q^2\rangle]/\langle V_R^2\rangle$ between the sense resistor and QVNS versus frequency for (a) the autocorrelation of channels A and B for the 200 Ω sense resistor and (b) the cross-correlation for 200 and 100 Ω sense-resistor measurements. Each point is integrated over the QVNS tone spacing of 100 Hz.

Matching was obtained by changing the lengths of the wires that connected the probes to the amplifiers and by modifying the dielectric between the conductors.

IV. Measured Data

Reasonably good matching of both the correlated and uncorrelated noise sources and the transmission-line impedances was achieved, as demonstrated in Fig. 2, which shows the trend of the acquired data. The larger scatter in the 200 Ω resistor data is due to the higher standard deviation of the eight times shorter 15 h length of the measurement as compared with the 116 h measurement at 100 Ω.

The differences between the QVNS and the sense-resistor signals are shown for the following: 1) the autocorrelated signals for each channel [Fig. 2(a)] and 2) the cross-correlated signals for the two measurements of the two different resistors [Fig. 2(b)]. The frequency response of interest is over a 640 kHz measurement bandwidth from 10 to 650 kHz (the low-pass filter cutoff frequency). One can see that the two autocorrelation differences are matched to better than two parts in 10^3 for the current system that was optimized for the 200 Ω sense resistor.

Fig. 2(b) shows the resistor–QVNS power differences of the cross-correlation signals for both the 200 and 100 Ω measurements. The cross-correlation difference for the 200 Ω measurement shows a stronger quadratic behavior with increasing frequency as compared with that of the 100 Ω measurement. At the beginning, we believed that the quadratic behavior of these differences was due to the correlated noise in the sense resistor from the current noise of the input circuit, which is larger for the 200 Ω source. After further investigation done at room temperature, we can state that this effect is caused by a small mismatch that has not been compensated. The cross-correlation differences for the two measurements do not perfectly scale because the 100 Ω measurement had less perfect matching and showed additional frequency dependence due to the issues described in Section V.

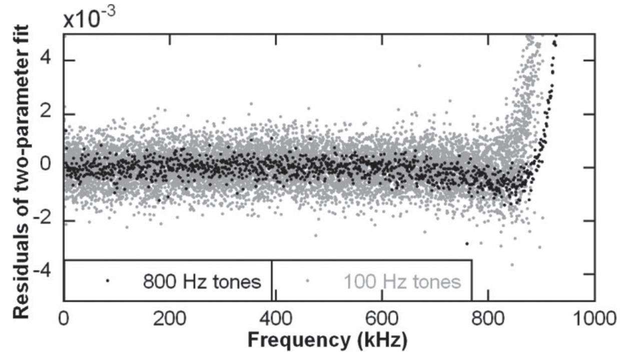

Fig. 3. Residuals of the two-parameter fit, $a_0 + a_2 f^2$, of the ratio between the Johnson noise of the resistance and QVNS noise for the 800 and 100 Hz tone spacing of the two QVNS codes. Each data set is integrated over a 15 h period using a 200 Ω sense resistor.

The ratio of the cross-correlated resistor and QVNS signals used to determine a_0 and k is also quadratic [20]. The ratio fits a two-parameter quadratic function so well that the residual of the fitted ratio of the data sets for both waveforms displays negligible frequency dependence across the 640 kHz measurement bandwidth, as shown in Fig. 3. A quadratic fit is expected and necessary to remove frequency dependence due to the electronic response, current noise, and imperfect impedance matching [18], all of which are quadratic in frequency. The flat frequency response of the residuals is particularly visible on the data for the noise widely spaced at 800 Hz tones because the wider frequency bins produce smaller scatter. We note that the same two-parameter fit that was previously used in the 100 Ω measurements works equally well for the 200 Ω measurements, even though the sense resistor is subject to a slightly different impedance matching. Other measurements with higher resistance values are planned for the future in order to better understand the limits of the two-parameter fit and the possible effects of current noise.

V. Frequency Response

Even though the fit residuals in Fig. 3 appear relatively constant, it is important that the a_0 constant term of the fit remain independent of frequency. The ratios of the sense resistor and QVNS power spectral densities are fitted over variable bandwidths, each starting at 10 kHz and stepping the end frequency from 320 to 650 kHz. This calculation is completed, and the fit parameters are determined for each bandwidth to analyze the frequency response, explore the self-consistency of the data, and identify effects that may affect the measurement uncertainty.

Fig. 4 shows the difference between the a_0 coefficient obtained from fitting the data and the a_0^{2010} value calculated from the CODATA 2010 [21] Boltzmann constant value. The error bars represent the measured statistical uncertainty that is effectively the standard deviation of the mean of a_0. As for the previous measurements of the 100 Ω sense resistor, the main source of the combined uncertainty in the Boltzmann constant measurement should be this statistical uncertainty. The fit parameter a_0 should remain frequency independent regardless of the sense-resistor value. Unfortunately, a_0 is frequency

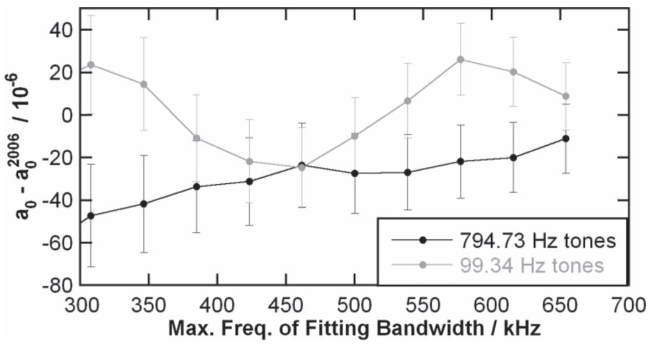

Fig. 4. Differences of the constant term a_0 from the calculated value a_0^{2010} based on the 2010 CODATA value of k. The difference is calculated over different fitted bandwidths for both tone spacings.

dependent for the present measurements, just as it was in the previous measurements in which this systematic error dominated the combined uncertainty.

Measurements at NIM with a similar QVNS–JNT system showed a flat frequency response for a_0 [22], [23]. This system used commercial ADCs and a different clocking scheme. These results inspired us to analyze our ADCs, clocking scheme, and data collection electronics for data-related errors that may be the cause of the frequency dependence of our measurements.

We have carefully investigated the frequency dependence and searched for a cause in our measurement system. We present evidence below that it is due to corrupted data produced by incorrect data transfer due to an intermittent error or jitter associated with the clock distribution. To find this error, we collected data from numerous overnight measurements and modified our data collection algorithm to reduce the number of chops to ten (from 100) that are integrated before storing the fast Fourier transform (FFT) on the PC hard drive. A single chop represents a 200 s integration period, half of which is dedicated to measuring the QVNS noise and the other half for the resistor noise. The data were averaged in groups of ten chops by increasing the starting chop by one for each iteration. This data-collection algorithm allowed us to observe how a_0 depended on chop. By investigating the FFTs of various groups of chops, we identified that the problem was due to intermittent unpredictable jitter in either the CLK or FRAME of Channel B (ChB).

Fig. 5 shows two examples of how a_0 becomes significantly frequency dependent for a group of ten chops that were affected by jitter. The frequency dependence of a_0 shown in Fig. 5(a) is clearly flat, which implies that the ten chops of this calculation are well synchronized and that the clocking signals were not corrupted by jitter. However, Fig. 5(b) shows a large frequency dependence in a_0 because a bit was lost during the data transfer of at least one chop. If the number of lost-bit chops is a majority of the ten chops that are averaged, then the resultant FFT contains a significant spectral aberration (i.e., abnormal frequency dependence). This case has not been reported in our previous measurements and analysis.

A straightforward way to observe that a bit was lost in the acquisition is by checking the high-frequency noise floor of the autocorrelation power spectral density. Fig. 6(a) shows the noise floor and spectral tones of the two channels for a

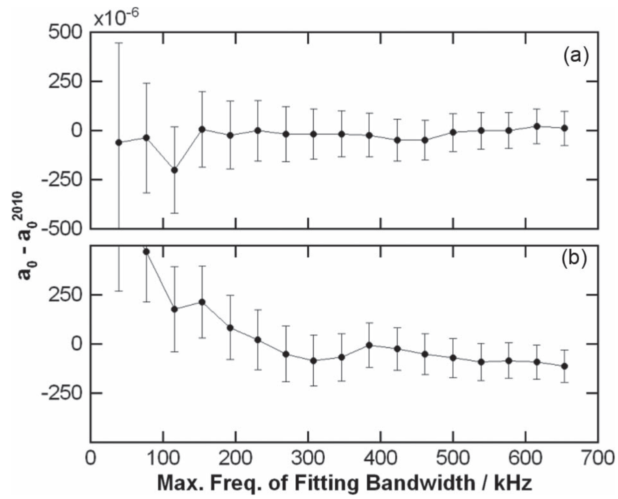

Fig. 5. Difference of a_0 from the 2010 CODATA value fitted over variable bandwidths. (a) A flat frequency response is obtained when the two channels are synchronized, whereas (b) random jitter produces curvature in the frequency response. Each graph is the result of a ten-chop calculation.

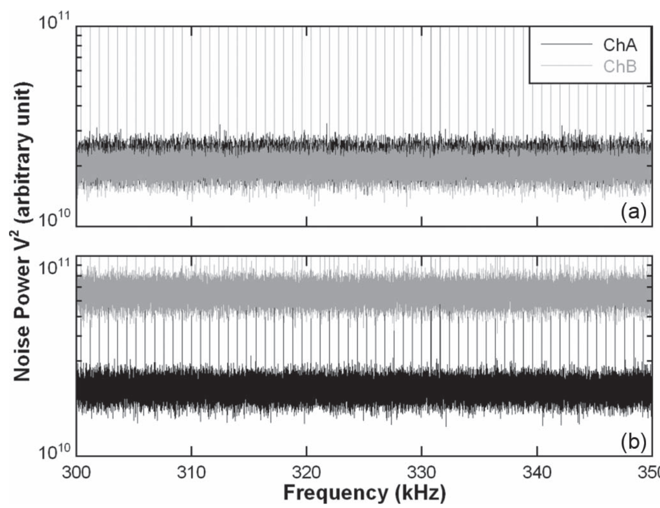

Fig. 6. Comparison of the autocorrelation FFTs for both channels, ChA and ChB, showing (a) well-matched noise levels when all ten chops are perfectly synchronized and (b) a significantly higher level in ChB when there is timing jitter affecting one of the ten chops.

chop in which all the data are synchronized with no timing jitter; the floors differ only due to the small unmatched gains of the amplifier chains for each channel. Fig. 6(b) shows a measurement in which the noise floor of ChB is significantly increased because of a timing error in which a bit is lost due to jitter in the ChB synchronization signals.

We also analyzed the time-domain behavior of the acquired signals in order to find the exact moment of the jitter and to search for possible repetition in the jitter events. Unfortunately, because the time-domain signal includes all of the noise from every signal in the amplification chain, input circuit, and switchboard, it was not possible to directly observe the jitter errors in the time domain. We have not yet been able to eliminate the jitter problem in the present clock board because there are limitations in the clock and data transmitter/receiver hardware.

VI. Conclusion

Our QVNS–JNT measurements of the 200 Ω sense resistor show differences of -10×10^{-6} and $+9 \times 10^{-6}$ between the measured and 2010 CODATA values of k for the 800 and 100 Hz tone spacings, respectively. For both data sets, the statistical uncertainty is 16×10^{-6}. The larger current noise of this higher resistance JNT measurement does not appear to affect the measurement of k, probably because all associated effects are removed by fitting, as are the other quadratic effects.

As in our previous measurement [15], the combined relative uncertainty must account for variations in a_0 with fitting bandwidth due to spectral aberrations. Unfortunately, this frequency dependence still dominates our measurement results. Motivated by the frequency-independent measurements at NIM [22], we have found data errors due to intermittent timing jitter on one of our cross-correlation channels. We expect to eliminate this error with new ADCs and a clock board that are now under development and consequently reduce the uncertainty of our next measurement of the Boltzmann constant.

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