Flat Frequency Response in the Electronic Measurement of Boltzmann's Constant

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***Abstract*—A new quantum-voltage-calibrated Johnson noise thermometer was developed at the National Institute of Metrology to demonstrate the electrical approach that determines Boltzmann's constant k, by comparing electrical and thermal noise power. A measurement with an integration period of 19 h and a bandwidth of 638 kHz results in a relative offset of 1×10^{-6}, from the current Committee on Data for Science and Technology value of k, and a type A relative standard uncertainty of 17×10^{-6}. Closely matched noise powers and transmission-line impedances were achieved, and consequently, the quadratic fitting parameters of the ratio spectrum show flat frequency responses with respect to the measurement bandwidth. This flat response produces a dramatically reduced systematic error compared to that of the National Institute of Standards and Technology measurement of k, in which the relative combined uncertainty was dominated by this error.**

***Index Terms*—Correlation, Johnson noise, Josephson voltage standard, noise measurement, temperature measurement.**

I. Introduction

MEASURING Boltzmann's constant k with comparable uncertainty by use of methods based on physical principles different from each other is required for the redefinition of k [1]. Although numerous efforts were made to meet this requirement, the current Committee on Data for Science and Technology (CODATA) value of k is dominated by the gas-based thermometry measurements [2]. The lowest uncertainty of 1.2×10^{-6} was achieved by the acoustic-gas thermometry, which measured the speed of sound in argon and deduced k from the molar gas constant R [3]. Another gas thermometry approach that measures the dielectric constant of helium gas has achieved an uncertainty of k of 7.9×10^{-6} [4], [5]. Measurements of the refractive index of helium gas result in a value of k with a relative standard uncertainty of 9.1×10^{-6} [6]. The Doppler broadening technique measured the structure of the absorption lines of NH_3 and obtained a relative uncertainty for k of 38×10^{-6} [7]. All of the aforementioned experiments work with some type of gas as the measurement medium.

The Johnson noise thermometer (JNT) measures the thermodynamic temperature T through the Nyquist relation $\langle V^2 \rangle = 4\ kTR\Delta f$, by measuring the mean-squared voltage noise power $\langle V^2 \rangle$ arising from the thermal fluctuation of the electrons in a resistor R over a frequency bandwidth Δf [8]–[10]. Alternatively, by measuring $\langle V^2 \rangle$ across R at a known T, the JNT provides the possibility of determining k. However, the challenge is that the noise voltages are extremely small, ~ 1.2 nV/Hz$^{1/2}$, for a 100-Ω resistor at the triple point of water (TPW). Extremely high-performance electronics and very long integrating times are required to measure the thermal noise with small statistical uncertainty [11].

Recently, the National Institute of Standards and Technology (NIST) has reported the first practical electronic measurement of Boltzmann's constant with Johnson noise thermometry that compared the thermal noise of a resistor at the TPW temperature to the pseudo-noise voltage synthesized by a quantum-voltage-noise source (QVNS) [12]. The result of $k = 1.380651(17) \times 10^{-23}$ J/K was consistent with the current CODATA value. The 12.1×10^{-6} relative combined uncertainty of this previous measurement was dominated by 1) "systematic effects that produce aberrations in the ratio spectra" (10.4×10^{-6}) and 2) random statistical uncertainty (5.2×10^{-6}) by combining two data sets, each having 116.6 h of integration. Neither the systematic error producing the spectral aberration nor the statistical uncertainty is currently believed to represent a fundamental limit of this measurement. Further improvements are necessary to reduce the uncertainty to contribute to the redefinition of k.

In order to demonstrate the reproducibility of this electrical approach to measuring k and to pursue lower measurement uncertainties, it is important to perform measurements under different conditions, and with different QVNS-JNT systems. The NIST recently presented measurements by doubling the sense resistor value from 100 to 200 Ω to produce higher noise power, although which yielded a lower statistical uncertainty for the same integration period, the fitting results of k showed aberrations with different fitting bandwidth as in the previous measurement [13]. In this paper, we describe the newly developed QVNS-JNT system at the National Institute of Metrology (NIM) [14], [15] and report the recent measurement result that shows flat frequency responses for both the noise-power

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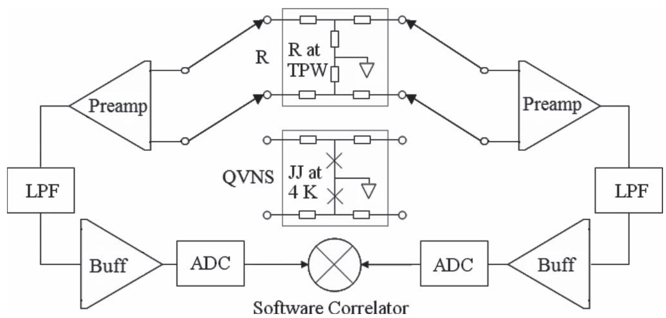

Fig. 1. Schematic of the NIM QVNS-JNT system.

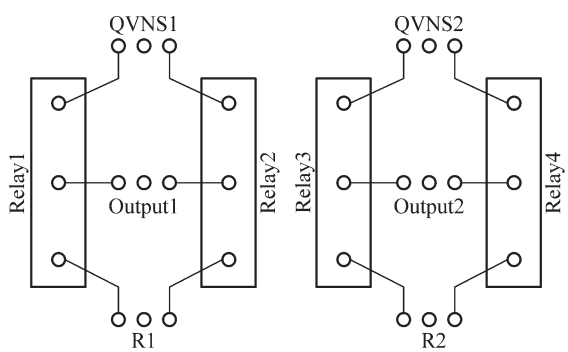

Fig. 2. Schematic of the switch circuit.

ratio and the quadratic fitting parameters. The flat frequency responses indicate the reduction in systematic error in the new system and demonstrate the potential of a relative uncertainty of 6×10^{-6} for measuring k with the JNT.

II. Construction of the NIM QVNS-JNT System

Fig. 1 shows the schematic of the NIM QVNS-JNT system. A switch circuit switches between the QVNS and the sense resistor R so that one of the noise sources connects to the two nominally identical amplification channels (see Fig. 2). Each amplification channel contains a low-noise high-gain differential preamplifier (Preamp), a single low-pass filter (LPF), a buffer amplifier (Buff), and an analog-to-digital converter (ADC). The digitized signals are transferred to computer via optical fibers and then cross correlated by software to remove the uncorrelated noise. All the electronics are sealed in aluminum boxes and powered by batteries to protect the measurement from possible electromagnetic interference (EMI).

The custom sense resistor consists of two Ni–Cr-alloy foils on an alumina substrate with a nominal resistance value of 50 Ω for each. Two pairs of gold-coated leads on the hermetically sealed package are connected to the two ends of the resistor network, and a fifth lead is connected to the center. The resistance of the sensor is measured by a dc resistance bridge through four-wire configuration to be 100.0065 Ω.

The resistor is mounted on a copper header soldered at the bottom end of a thin-wall stainless steel probe. The probe is about 60 cm long, filled with Argon gas and hermetically sealed. Two cables with the same length made of three Teflon-insulated wires transmit the differential noise signals from the resistor to the connectors at the top of the probe. The probe is immersed in a TPW cell. The cell is placed in a stainless steel Dewar filled with ice.

The QVNS chip was fabricated at NIST in Boulder. There are ten superconductor–normal metal–superconductor (SNS) Josephson junctions in each of the two arrays [16]. The critical current of the Josephson junctions is around 6 mA, and the characteristic frequency is about 5 GHz. The chip is mounted on a flexible package in a magnetically shielded probe. The probe is cooled to 4 K in a 100-L liquid-helium Dewar.

The quantum-voltage-noise waveform is synthesized by the bipolar pulse-driven technique. Although the operation margin of the Josephson junction array is lower than that for the unipolar pulse-driven synthesis [16], the advantage is that there is no dc offset for the synthesized ac waveform. The frequency of the microwave signals that drives the pulse generator is 9.999872 GHz, which is split by the pulse generator to get the sampling frequency $f_s = 4.999936$ GHz. The bit length of the digital code is selected to be $M = 4\,999\,936$; thus, the repetition frequency of the waveform is $f_1 = f_s/M = 1$ kHz. The comblike multitone waveform consists of odd-harmonic tones of f_1 from 1 to 999 kHz with tone spacing of 2 kHz. The tone amplitude is set to 54.9268 nV to ensure that the voltage spectral density $V_{\mathrm{QVNS-calc}} = 1.2282$ nV/Hz$^{1/2}$ closely matches the spectral density of the thermal noise, i.e., $V_{R-\mathrm{calc}} =$ 1.228272 nV/Hz$^{1/2}$, of the 100.0065-Ω resistor at the TPW.

To minimize the nonlinear effects of the preamplifier, it is required that both the total noise powers and impedances of the QVNS and R noise sources are closely matched to each other [17]. Note that four resistors with a nominal resistance of 100 Ω are fabricated in the transmission wires on the Josephson junction chip to match the resistance of the sense resistor noise source. Unfortunately, these resistors introduce extra uncorrelated noise at 4 K. Additional small chip resistors with a resistance of 1.5 Ω are inserted between the signal leads and the transmission wires, for the QVNS and R, to closely match both the uncorrelated noise power and the resistance.

The switch circuit consists of four latching relays mounted on an FR4 printed circuit board (PCB). The relays are controlled by a field-programmable gate array to alternate between the noise sources such that either the QVNS or the resistor is connected to the two cross-correlation channels. In the PCB, the signal traces are symmetrically placed in the top layer, and the total length from the input lead to the output lead is about 15 mm. The copper ground plane is in the bottom layer, in which the area opposite to the signal traces is removed. This is important because the smallest possible shunt capacitance helps to minimize the effect of dielectric losses [12].

The preamplifier in the NIM system is of similar design to that in the NIST system [18], which has a gain of about 70 dB and the input voltage noise of about 1.1 nV/Hz$^{1/2}$ in the 36-Hz to 1-MHz bandwidth and a common-mode rejection ratio of 100 dB at 100 kHz. An offset compensation circuit is included to null the dc offset of the signal before the differential stage to reduce the nonlinear distortion of the preamplifier [19]. Two-tone measurements show that, for waveforms with frequencies of 300 and 301 kHz and amplitude of 54.9268 nV, the second-order intermodulation distortion is -121.5 dBc, and the others are even lower [15].

The 11-pole LPF is of a Butterworth response with a cutoff frequency of 800 kHz, which defines the measurement bandwidth. The buffer amplifier has a gain of 11× and an output resistance of 50 Ω, which in turn drives the ADC.

Commercial ADCs with a high resolution of 20 bits at 2-MHz sampling frequency are used for data acquisition. There is a 1-MHz anti-aliasing filter integrated within the ADC to prevent the aliased high-frequency signals from contributing to the measured noise-power spectra. The ADCs are clocked and triggered by external clocks provided by commercial function generators via optical fibers. The phases of the clocks are carefully adjusted before the measurement so that the two acquisitions are synchronized.

The data are then optically transmitted to the computer. The software performs fast Fourier transforms (FFTs) on the data and cross correlates the data that are acquired for every 1 s, accumulates these data for 100 s and saves the result as one chop, and then switches the relays to measure the other noise source. Note that the thermal noise is a random process with flat power spectral density. The QVNS generates a periodic signal with a minimum frequency defined by the codes memory length. In our experiment, the ADC digitizes an integer number of periods of the pseudo-noise voltage waveform at 2 MHz for 1 s, which ensures an FFT spectrum with all harmonic tones in well-defined frequency bins. It is important that no windowing is necessary for FFTs of both waveforms, so that no error is introduced.

The 2-MHz sampling frequency and 1-s acquisition period for every chop result in a 1-MHz bandwidth and a 1-Hz resolution power spectrum. The averaged cross-correlation power spectra of the thermal and electrical noise are then summed and compared over discrete 2-kHz intervals centered at the frequency of the tones of the QVNS-synthesized waveform. Boltzmann's constant k can be calculated from

$$k = \left. \frac{\left\langle V_R^2 \right\rangle}{\left\langle V_{\mathrm{QVNS}}^2 \right\rangle} \right|_{f=0} \frac{V_{\mathrm{QVNS-calc}}^2}{V_{R-\mathrm{calc}}^2} k_{2010} \qquad (1)$$

where k_{2010} is the current CODATA value of k [2].

There are a number of the differences of the present NIM system from the one at NIST. 1) The NIST system uses a unipolar pulse-driven technique to synthesize the quantum-voltage-noise signal, whereas the NIM system uses a bipolar pulse drive, which has the advantage of having no dc offset in the waveform. 2) NIM maintains the TPW cell in an ice bath to avoid possible EMI from the thermoelectric cooler that is used in the NIST system. 3) Our switching circuit uses only one relay to connect each input lead to one output lead, whereas in the NIST switching circuit, there are two relay connections. 4) Our amplifier is of similar design to that of NIST, except our preamplifier uses a transistor instead of a junction field-effect transistor for the mirror current source. 5) Instead of using two 11-pole LPFs to prevent the aliasing signals from contributing to the measured noise power in the NIST system, there is only one LPF in the NIM setup. The high-frequency signals are attenuated by use of the anti-aliasing filter integrated within the ADC. 6) In the NIM system, the signals are digitized by commercial ADCs with 20-bit resolution and 2-MHz sampling frequency. While in the NIST system, 16-bit custom ADCs are used for sampling at 2.08 MHz. One can learn more details of the NIST system elsewhere [12], [16], [18], [19].

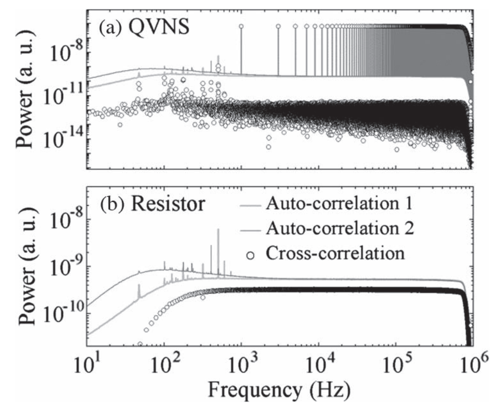

Fig. 3. Measured FFT spectra from noise sources of (a) QVNS and (b) sense resistor R. The gray lines are the auto-correlation spectra for each of the two amplifier channels, and the black circles represent the cross-correlation spectra between channels.

III. Measurement Results

In contrast to the digest summary paper [14], this paper features a measurement with a longer integration period of about 19 h, over which 342 chops were acquired for the QVNS and R, respectively. The averaged auto- and cross-correlation spectra for all of the 342 chops are shown in Fig. 3.

Although there are some EMI signals at frequencies lower than 1 kHz and some tones at around 900 kHz, no EMI signals were seen between 1 and 800 kHz. The amplitude of the spectra and the noise floors of the amplifiers are flat and of almost the same amplitude in the frequency range of a few kilohertz to a few hundred kilohertz. The curved behavior of the noise floor in the low-frequency range is a combined result of the $1/f$ noise of the electronics and the 36-Hz high-pass filter integrated in the amplifier. The amplitude of the spectra visibly decreases above 700 kHz, due to the frequency response of the LPF. To eliminate the visible effect from the EMI and the nonideal behavior of the electronics, we defined our measurement bandwidth from 5 to 650 kHz, but not to 800 kHz.

Fig. 4 shows the relative differences of the auto- and cross-correlated noise-power spectra between the QVNS and the sense resistor. It can be seen that the two auto-correlation differences are matched to better than two parts in 10^3 within the 650-kHz measurement bandwidth, and the relative difference shows the quadratic dependence with frequency. The cross-correlation noise difference spectrum is matched to better than one part in 10^3 and shows the flat frequency response. This was achieved by carefully trimming the length of the transmission lines that connected the noise sources to the cross-correlation electronics.

In spite of the closely matched noise powers and impedances, different transmission lines may still result in different transfer functions and thus the deviation of the noise-power ratio from the unity response. Furthermore, there remain errors due to

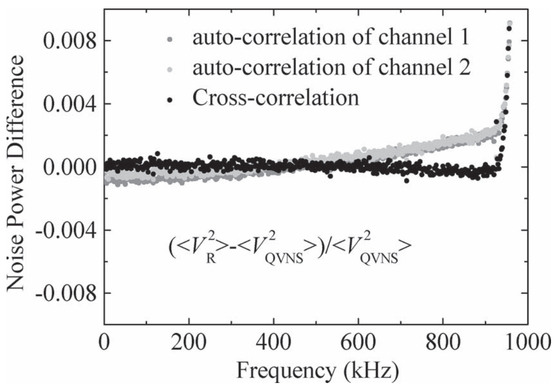

Fig. 4. Relative differences of the auto- and cross-correlated noise-power spectra between the QVNS and R noise sources.

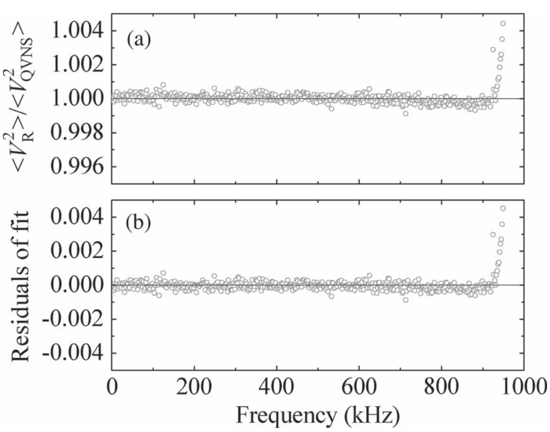

Fig. 5. Frequency response of (a) the power ratio and (b) the residuals of the two-parameter fit.

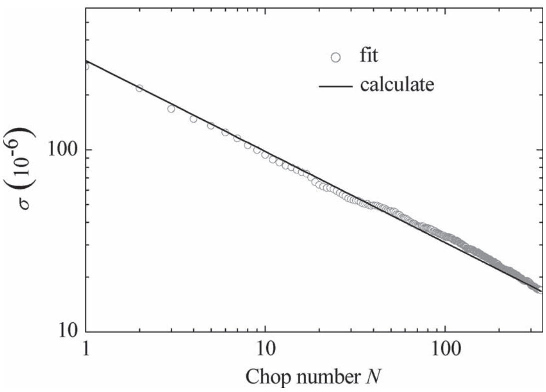

Fig. 6. Variations of the standard deviation σ of a_0 with increasing integration period for a fitting bandwidth from 5 to 643 kHz. The straight line represents the theoretically expected statistical uncertainty [11].

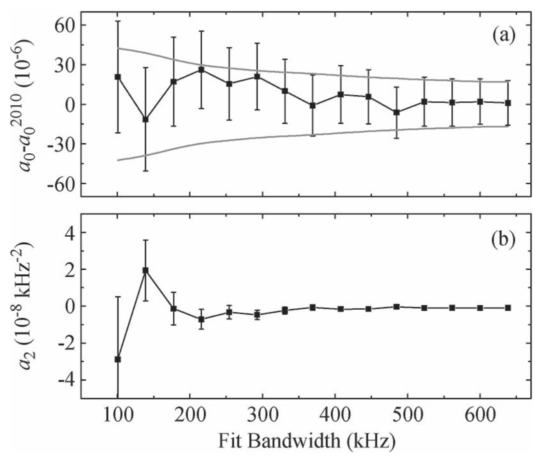

Fig. 7. Flat frequency responses of the fit parameters, i.e., (a) a_0 and (b) a_2, on the fitting bandwidth. The gray boundary lines in (a) represent the $\pm\sigma$ of a_0.

the amplifier noise currents. When connected to the R, the preamplifier noise currents pass through the sense resistor and are measured by both channels, leading to extra correlated noise power, which does not exist when connected to the QVNS. Fortunately, both of these errors are relatively small and have f^2 dependence. The resulting expected quadratic frequency response of the noise-power ratio can be corrected by least-squares fitting with a two-parameter formula $a_0 + a_2 f^2$. Fig. 5(a) shows the noise-power ratio of our measurement, from which one can see the weak quadratic dependence of the ratio on the frequency. Fig. 5(b) shows the residuals of the two-parameter fitting on the bandwidth from 5 to 643 kHz. The residuals appear to be reasonably flat. Absence of significant curvature in the frequency dependence of the residuals indicates very good linearity of the electronics.

The coefficient a_0 and the associated standard deviation from the fitting are used to determine k and its relative statistical uncertainty. The coefficient a_2 characterizes the remaining quadratic response. For the current measurement with an integrating period of 19 h and a fitting bandwidth from 5 to 643 kHz, the relative difference of k from the 2010 CODATA value is found to be 1×10^{-6}, and the relative statistical uncertainty approaches 17×10^{-6}. In Fig. 6, the variation of the standard deviation σ of a_0 with increasing integration period is tracked and plotted, which linearly decreases with increasing chop number on a logarithmic scale. Indeed, it closely follows the straight line that represents the theoretically expected $1/N^{1/2}$ behavior of the statistical uncertainty [20].

IV. Reduced Systematic Error

For the previous Boltzmann's constant measurement with the NIST QVNS-JNT system, the small spectral aberrations were found for the variation of a_0 when the data were fitted over different bandwidths. The aberrations were probably caused by systematic errors, and they dominated the measurement uncertainty [5]. To check the contribution of the spectral aberrations in the present measurement, in Fig. 7(a) and (b), we plot the results of fitting the data with different bandwidths over different ranges starting from 5 kHz and ending at successively higher frequencies. The difference $a_0 - a_0^{2010}$ represents the relative offset of k from the 2010 CODATA value, and $a_0^{2010} = (V_R/V_Q)^2$. It is important to note that both a_0 and a_2 show flat behavior as a function of fit bandwidth. The $\pm\sigma$ standard deviation of a_0 is also plotted in Fig. 7(a) to show the boundary that allows the variation of $a_0 - a_0^{2010}$. One can see that the values of $a_0 - a_0^{2010}$ are consistent with each other within the error bars for the standard deviations of the different fitting bandwidths. The flat responses demonstrate the self-consistency of the data and indicate that the systematic error that leads to the spectra aberrations in [12] is greatly reduced in the present measurement.

V. Conclusion

With the new QVNS-JNT system developed at NIM, a measurement with an integration period of 19 h resulted in a relative offset of 1×10^{-6}, from the 2010 CODATA value of k, and a type A relative standard uncertainty of 17×10^{-6}. Most importantly, the quadratic fitting parameters of the power-ratio spectrum showed extremely flat frequency responses for different fitting bandwidths, indicating that the systematic errors that produced the spectral aberrations in the NIST system are greatly reduced.

For the previous measurement with the NIST system, the combined relative uncertainty of k was dominated by the systematic effects that produce the spectral aberrations (10.4×10^{-6}) and the random statistical uncertainty (5.2×10^{-6}). For the present measurement, based on the flat frequency response of a_0, we estimate that the contribution of the spectral aberrations to the combined relative uncertainty reduces to less than 1×10^{-6}. Experiments are ongoing at NIST to understand the sources of the spectral aberrations [13] and at NIM to reduce the statistical uncertainty. With further improvements, we anticipate reaching an electronic measurement of k at a combined relative uncertainty of 6×10^{-6} with both systems.

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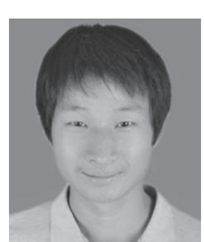

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