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Low-noise correlation measurements based on software-defined-radio receivers and cooled microwave amplifiers

Teemu Nieminen,^{a)} Pasi Lähteenmäki,^{a)} Zhenbing Tan, Daniel Cox, and Pertti J. Hakonen
Low Temperature Laboratory, Department of Applied Physics, Aalto University School of Science,
P.O. Box 15100, FI-00076 Aalto, Finland

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We present a microwave correlation measurement system based on two low-cost USB-connected software defined radio dongles modified to operate as coherent receivers by using a common local oscillator. Existing software is used to obtain I/Q samples from both dongles simultaneously at a software tunable frequency. To achieve low noise, we introduce an easy low-noise solution for cryogenic amplification at 600–900 MHz based on single discrete HEMT with 21 dB gain and 7 K noise temperature. In addition, we discuss the quantization effects in a digital correlation measurement and determination of optimal integration time by applying Allan deviation analysis. *Published by AIP Publishing.* [<http://dx.doi.org/10.1063/1.4966971>]

I. INTRODUCTION

Shot noise is an important measurement quantity in various experiments in mesoscopic physics, providing information of the transport properties and internal energy scales of the sample.¹ Furthermore, shot noise correlations can be used to detect coherence phenomena, including Hanbury-Brown and Twiss effects^{2–4} and Cooper pair splitting.^{5–7} Nevertheless, systems for shot noise cross-correlation measurements have gained little attention in the literature, although some realizations have been described for both analog^{8–10} and high-speed digital detection.^{11–13} Systems based on high-speed digitizers acquire all information contained in the microwave signal, which enables the study of various other physical phenomena, including the quantum vacuum.^{14,15} Multiple phase-locked digitizers operating at different RF frequencies facilitate detection of correlations in cluster states that provide a new kind of scheme for quantum computing.^{16,17} On the applications side, shot noise measurements can be employed for primary thermometry at low temperatures.¹⁸

Wide-band analog and digital systems detect broad-band noise over a wide bandwidth which improves their sensitivity compared to narrow band systems. However, the impedance of a nanoelectronic sample is typically high compared to the 50 Ω impedance of RF measurement systems, thus requiring impedance matching to optimize the coupling of sample noise to the detection system.⁹ Along with increased coupling, impedance matching does limit the available bandwidth, as dictated by the Bode-Fano criterion.¹⁹ Therefore, a narrow-band measurement system can achieve similar performance as a broad-band one when used together with a highly resonant matching circuit.¹¹ Typical PCI/PXI express based wide-band digital measurement systems have data output rates ranging from several hundred MB/s up to several GB/s which set high requirements for on-the-fly data processing and/or storage, often solved using rather cumbersome field-programmable

gate array (FPGA) solutions.¹⁵ In contrast, the data output of a digitizer with sampling rate of a few MS/s (megasamples per second) can be transferred via a regular USB bus and processed on a low-end desktop computer.

We present a correlation measurement system based on two 8-bit, 2 MS/s digitizers modified from two USB-connected Software Defined Radio (SDR) dongles and a simple cooled HEMT amplifier for correlation measurements in the 600–900 MHz range. These cooled, home-made cryogenic amplifiers yield a noise temperature of $T_n \approx 7$ K, which results in 3–4 times longer integration time when compared with high-end commercial amplifiers (with $T_n \approx 4$ K). This article is organized as follows. An overview of the measurement system and data processing scheme is presented in Sec. II. Section III discusses the sensitivity and stability of the system and presents the structure and performance of the home-made amplifier. The effects of quantization noise are described in Sec. IV together with a demonstration measurement of graphene shot noise and possibilities for further improvements of the system. Section V summarizes this study.

II. SYSTEM DESCRIPTION

The SDR dongles in the presented setup are based on Realtek RTL2832U chips with a software modification²⁰ enabling the capture of raw in-phase/quadrature-phase (I/Q) data. We have utilized this software with two dongles modified for coherent reception by removing the 28.8 MHz quartz crystal (XTAL) from one of the dongles and connecting the dongle with capacitive coupling to the crystal of the other dongle to share a common frequency reference between the dongles. Without this modification the receivers become incoherent and are useless for correlation measurements. A block diagram of the system with the modification is presented in Fig. 1.

The dongles are capable of capturing data at a continuous uninterrupted rate of around 5 MB/s/dongle, which is higher than the rate provided by most oscilloscopes, but significantly

^{a)}T. Nieminen and P. Lähteenmäki contributed equally to this work.

FIG. 1. Block diagram of the essential parts of the dongle and the measurement setup.

less than what can be accomplished by PCI Express based digitizers. Nevertheless, this modification provides a reasonably powerful alternative for correlation measurements. The RTL2832U chip in the dongle is used as an 8-bit ADC for digitizing the filtered signal which has first been mixed down by a separate tuner chip that also determines the tunable frequency range (typically 22–2200 MHz). The RTL-chip also handles communications such as USB down to PC and I2C to the tuner, while most of the other standard functions of the chip are ignored in our setup. The particular devices in our system can be flexibly tuned by software to 48.25–863.25 MHz. They accept low (aerial antenna) level signals and as such require less amplification compared to an oscilloscope or PCI Express based digitizer. We operate the dongles at a sample rate of 2 MS/s to ensure reliable operation, since the digitizers may drop individual samples at higher sample rates, leading to a time shift of the correlation peak in the time domain correlation array and thus to a significant reduction in correlation value.

In a typical application of measuring the correlated noise power from a nanoscale sample, the captured data are continuously stored on a hard disk and processed simultaneously, a single 1 s (2 MS) buffer at a time. To remove effects caused by slightly alternating gains in amplifier lines, we normalize the cross-correlation with the geometric average of the autocorrelations from both channels. Thus, the normalized cross-correlation of N IQ demodulated samples represented by arrays of raw complex voltages a and b is defined as

$$z_{\text{cor}} = \frac{\sum_{i=1}^N a_i \bar{b}_i}{\sqrt{P_a P_b}}, \quad (1)$$

where autocorrelations P_a and similarly P_b are obtained as

$$P_a = \sum_{i=1}^N a_i \bar{a}_i. \quad (2)$$

We compensate the effect of differing cable lengths and slight differences in capture start times between the two dongles by calibrating the system with a strong correlated

noise source (e.g., by increasing the current bias of the measured sample) prior to each measurement sequence. The duration of each measurement sequence with on-the-fly processing at 100% duty cycle can be up to several hours, after which a re-calibration is conducted to ensure the simultaneity of the measurement. Computationally more demanding quantities, such as power spectra or second order coherence, can be calculated after the measurement from the stored raw data.

III. RESULTS

A. Digitizers

The sensitivity of the setup was determined by measuring correlation with two channels separated (producing ideally zero correlation). Raw data were collected for 1.5 h after which the data were integrated over a time period τ . Sensitivity was determined as the standard deviation of 50 successive integrated samples of period τ . Theoretically the sensitivity of a radiometer follows the formula $\Delta T = T_s / \sqrt{B\tau}$, where T_s denotes the system noise temperature, B is the frequency bandwidth, and τ is the integration time.²¹ In digital radiometers the product $B\tau$ is replaced with the amount of digitized samples, N , resulting in the formula $\Delta T = T_s / \sqrt{2N}$, where the detection of both quadratures (I and Q) results in an additional factor of two.⁸ By varying τ and normalizing with total power, we obtain sensitivity relative to amplifier noise temperature as a function of integration time, presented in Fig. 2. The theoretical and measured relative sensitivities at 100 s integration time are 5.0×10^{-5} and 6.7×10^{-5} , respectively, the latter corresponding to temperature sensitivity of 470 μK or 4.7 mK/ $\sqrt{\text{Hz}}$ with 7 K system noise temperature. It can be seen that the sensitivity of cross-correlation keeps improving as the integration time is increased, while autocorrelation starts to suffer from $1/f$ noise and drift above 1 s integration times.

However, standard deviation is problematic when determining the optimal integration time, i.e., the longest integration time not affected by $1/f$ noise and drift. The standard

FIG. 2. Relative sensitivity of the setup determined from standard deviation of 50 successive integration periods of varying lengths. Inset: Relative Allan deviation of the setup with the blue curve corresponding to autocorrelation of the original time series, red curve to time series with drift periods removed, and yellow curve corresponding to cross-correlation.

deviation at integration time τ is determined from a sequence of $50 \times \tau$, which is affected by drifts occurring in time scales similar and significantly longer than τ , making it difficult to identify the optimal integration time. To overcome this problem, we employed Allan deviation which is defined as the deviation between differences in successive samples as

$$\sigma_y(\tau) = \sqrt{\frac{\sum_{i=1}^{N-n} [T_{n,i+1}(\tau) - T_{n,i}(\tau)]^2}{2(N-n)}}, \quad (3)$$

where $T_{n,k}(\tau)$ denotes an average over n individual measurements from index k to $k + n - 1$. Allan deviation was originally developed in 1966 for characterizing the frequency stability of atomic clocks.²² More recently the same method has been generalized for radiometer stability.^{23,24} Allan deviation allows identification of the dominating noise mechanism at a given time scale by determining the slope of $\sigma_y(\tau)$ curve on a log-log plot. A slope of -0.5 is a sign of white noise (ideal radiometer case), while zero slope denotes the dominance of $1/f$ noise and a positive slope results from drift.²⁵ These regions can be identified in the inset of Fig. 2, where cross-correlation is dominated by white noise at all integration times while autocorrelation reaches its best sensitivity at $\tau \approx 10$ s. The slope of autocorrelation at $\tau > 10$ s changes from positive to zero when clear drift periods due to varying ambient temperature are removed from the beginning and end of the sequence.

The final performance of the cryogenic setup in our case was limited by excess correlation with a magnitude of 4×10^{-3} relative to system background noise. This could easily be improved by better isolation of channel cross talk. At the input we used single cryogenic circulators to isolate amplifier input noise both from the sample and the other amplifier, although their limited isolation of 20 dB is sufficiently low to explain the measured excess correlation. Some grounding issues might have contributed to the excess as well. These issues were absent in the tests done on a simpler room temperature setup.

B. Amplifier

A cryogenic low noise amplifier operating in the 600–900 MHz range was designed based on single Avago ATF-36077 ultra-low-noise Pseudomorphic High Electron Mobility Transistor (pHEMT). This component was chosen based on recent reports²⁶ indicating that it could achieve excellent noise characteristics, wide band gain, and low power consumption while cooled down to deep cryogenic temperatures. This was quickly tested, but no obvious improvement was observed when moving from 4 K environment to 700 mK. However, this was not the main focus of the research and no further attention was given to this problem. The HEMT is connected in common source configuration with a lumped element matching circuit on a printed circuit board (PCB) made of Rogers RO4350B substrate. The PCB is soldered to a copper enclosure, equipped with SMA connectors for RF input and output, while feed-through pi-filters are used for gate and drain DC bias.

The circuit schematic of the designed amplifier is presented in Fig. 3 and the component values are listed in Table I.

FIG. 3. Circuit schematic of the home-made HEMT amplifier.

The impedance matching network is a simple L-section of L1 and C3. Resistors R1–R3 stabilize the amplifier while, in addition, the gate bias voltage (with negligible current) is fed through R1. DC-blocking capacitors C1 and C5 were chosen so that their parasitic inductive impedance compensates capacitive impedance at the operational frequency to minimize the attenuation.²⁷ Similarly, L2 exhibits maximal impedance at the operational frequency. Power supply bypass capacitors C2 and C4 filter the bias voltages and serve as a grounding path at intermediate frequencies. The fabricated amplifier employed Johanson Technology S-series capacitors with NPO dielectric and Murata LQW18A series inductors.

The amplifier was tested at the 4 K stage of a dry dilution refrigerator. Lowest noise temperature was reached at a drain voltage of 0.8 V and gate voltage of -0.15 V. At this bias point, the drain current is 9.4 mA, resulting in power consumption of 7.5 mW. A gain of $G = 21$ dB was measured around 750 MHz with 350 MHz half-power band width at 570–920 MHz. Noise temperature of $T_N = 7$ K was obtained using the standard Y-factor method by feeding attenuated noise from a calibrated avalanche diode noise source at room temperature while the amplifier was cooled down to 4 K. Measured noise temperature and gain are plotted in Fig. 4 as functions of frequency.

IV. DISCUSSION

Representing an analog quantity with a finite number of discrete levels involves quantization error due to rounding to the nearest quantization level. These quantization errors are observed as noise in digital continuous sampling systems. This quantization noise has variance of the form $\sigma_e^2 = \frac{\Delta^2}{12}$, where Δ is the quantization step size.²⁸ It should be noted that the variance, as well as the noise power, is proportional to the square of the quantization step size. The spectral density of quantization noise is approximately uniform, and thus it can be considered as white noise.

The quantization noises of two separate channels, however, are not correlated, and thus quantization noise is less significant in correlation receivers. A 1-bit correlator requires 2.43 times the integration time of an ideal analog correlator to achieve equivalent accuracy. The difference in integration

TABLE I. Component values used in the amplifier.

C1	47 pF	L1	24 nH
C2	1 nF	L2	270 nH
C3	0.5 pF	R1	10 kΩ
C4	1 nF	R2	33 Ω
C5	47 pF	R3	1 kΩ

FIG. 4. Gain and noise temperature of the HEMT amplifier measured at bias point $V_g = -0.15$ V, $V_d = 0.8$ V, and $I_d = 9.4$ mA. Dashed vertical lines denote the -3 dB band width.

times becomes insignificant as the bit depth increases, being only 2% for a 4-bit correlator.²⁹

Because of the relatively small difference in integration times, the early digital radiometers were equipped with one- or two-bit correlators.^{30,31} More recently, low-bit-depth correlators have been used in multichannel applications and when high sample rates are required. The drawbacks of low-bit-depth correlators include the requirements of linearization of the results and accurate adjustment of the digitizer threshold levels. It should also be noted that the results mentioned above apply only for Gaussian signals.²⁹ Power- or V^2 -correlations and other more involved data statistics require bit depths in excess of one. In addition, autocorrelation of a single channel is affected by the additive quantization noise. Because of these reasons, a low-bit-depth system may experience unexpected performance degradation when the measurement conditions differ from the above criteria. While having low bit depth has its drawbacks, the effective number of bits (ENOB) of our setup (≈ 7) is still sufficient for analysis of microwave shot noise, with negligible loss of sensitivity.

The setup depicted in Fig. 1 was used successfully in a dry dilution refrigerator to collect shot noise data from a graphene sample mounted on the mixing chamber plate at 50 mK.³² A few sweeps of cross- and autocorrelation data from this experiment have been plotted in Fig. 5 for reference. Calibration of the system was done by taking the value for Fano factor of diffusive systems, $F = 1/3$, yielding an accuracy of $\pm 30\%$ for the autocorrelations.

There are several possible future improvements to our system regarding the frequency reference and limitations in measurement bandwidth. The dongle used in our setup has a frequency reference with ± 100 ppm stability, while similar SDR dongles are available with a more stable frequency reference. Alternatively, we have also successfully connected the dongles to an external frequency reference. The sample rate of our system (2 MS/s) limits the measurement bandwidth and consequently the sensitivity of noise power measurement. To improve sensitivity, the USB dongles can be replaced with a data acquisition card with higher sample rate. As the sample rate is increased, the data processing and/or storage performance needs to be considered to maintain a reasonably high duty cycle during measurements. Moreover, several phase-locked dongles can be used to measure multiple frequencies simultaneously, achieving similar phase stability as standard microwave measurement devices phase-locked

FIG. 5. Autocorrelation ($P_{a,b}$) expressed in noise equivalent temperature and absolute value of normalized complex cross-correlation ($|z_{cor}|$) of shot noise from a graphene sample as a function of DC bias current applied across the two-terminal sample, measured in a real experimental setup in a dry dilution refrigerator at 50 mK temperature around 800 MHz frequency with 1 s sample averages. Cryogenically cooled home-made LNAs were used as the first stage amplifiers. The offset due to system noise background is removed from the autocorrelation curves.

with a 10 MHz reference signal. This is useful when addressing correlations in quantum mechanical cluster states.^{16,17}

V. CONCLUSIONS

We have presented an original microwave correlation measurement system based on SDR dongles and cooled HEMT amplifiers. A sensitivity of $4.7 \text{ mK}/\sqrt{\text{Hz}}$ was achieved using the amplifiers with 7 K noise temperature and 21 dB gain. The software tunable system can measure a 2 MHz band of frequency spectrum up to 860 MHz. The limited number of quantized levels in the digital system was observed to have negligible effect in noise measurement performance. In addition, we discussed the advantages of Allan deviation as a tool for determining the optimal integration time and dominating noise mechanism in a noise measurement system. Altogether, our simple system shows performance comparable to other correlation measurement systems.^{9,11} The system can be used to obtain full data statistics in a single measurement, enabling the determination of several physical quantities.

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