Microwave Transmission Technique for Accurate Impedance Characterization of Superconductive Bolometric Mixers

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Abstract—The impedance of the superconducting hot-electron bolometer mixer was measured in the range 0.2-4 GHz. A special technique relying on determination of the S_{21} transmission parameter of the device was used. Many advantages of the technique (wide frequency range, *in situ* calibration, low test power, accuracy) are demonstrated. The estimate of the mixer bandwidth from the impedance data using the mixer theory is in good agreement with the results of the direct bandwidth measurements.

Superconductive hot-electron bolometric (HEB) mixers are rapidly becoming a very attractive alternative sensor for low-noise heterodyne spectroscopy applications at terahertz frequencies. It has been demonstrated that for this type of mixer a noise temperature below 1000 K is possible up to 2.5 THz [1]. The mixer sensitivity is not expected to degrade through the entire far-infrared spectral range. One can expect that HEB mixers will be used in practical heterodyne receivers very soon. One of the important issues for designers of bolometric mixers is a careful evaluation of the mixer bandwidth. This parameter is usually set by the intrinsic relaxation time which depends on the material and size of the device. There is an extensive literature on the bandwidth measurements in two major types of HEB mixers (diffusion-cooled and phonon-cooled) [2,3,4,5] and, generally, bandwidths between 1 GHz and 6 GHz has been demonstrated. Lack of widely tunable local oscillators (LO) at terahertz frequencies often requires large mixer bandwidths; up to 10 GHz or 20 GHz when fixed-tuned lasers are used as the LO source. Thus, further improvement of the HEB bandwidth may be necessary.

In this paper we describe an accurate microwave impedance measurement technique which provides a good bandwidth estimate in an HEB device, thus eliminating the need for tunable submillimeter radiation sources which are expensive and/or not always available. These impedance measurements can be performed using a conventional microwave network analyzer. A special calibration routine for S_{21} measurements allows the device impedance to be de-imbedded from the surrounding circuit (and its parasitic) without using any calibration standards. The results presented show the implementation of this technique for a Nb diffusion-cooled mixer in a frequency range up to 4 GHz. The data obtained with the novel technique are in good agreement with the direct measurements of the bandwidth using two tunable 600 GHz sources.

During past decade there has been detailed studies of the mixing mechanism in hot-electron bolometers made from superconductive films. The theoretical basis describing the mixer operation is well established at present [6,7]. According to the theory, the mixer bandwidth is given by

$$\Delta f_{3dB} = \frac{1}{2\pi\tau_0}\left(1 + C\frac{R - R_L}{R + R_L}\right) \quad (1)$$

Here R *is* the device dc resistance at the operating point, R_L *is* the intermediate frequency (JF) load resistance (normally 50 Q), C is the self-heating parameter ($C< 1$), τ_0 is the electron temperature relaxation time in the device measured at the critical temperature T_c when a self-heating is not important (ie: using small transport current).

Using the same notation, the IF impedance of a HEB device is given by the following formula [7]:

$$Z(\omega) = R\frac{1+C}{1-C}\frac{1+j\omega\frac{\tau_0}{1+C}}{1+j\omega\frac{\tau_0}{1-C}}, \quad (2)$$

or in other terms

$$Z(f) = R_d\frac{1+jf/f_2}{1+jf/f_1}, \quad (3)$$

where $R_d=R(1+C)/(1-C)$ *is* the differential resistance of the mixer, $f_1 = (1-C)/(2\pi\tau_0)$, $f_2=(1+C)/(2\pi\tau_0)$. The relationship between frequencies, and f_2 is straightforward, i.e. $f_1 = f_2\ R/R_d$.

According to Eqs. (2) and (3) the impedance is purely real at low frequencies ($Z(0) = R_d$) and high frequencies ($Z(\infty) = R$). In the intermediate region between f_1 and f_2 the impedance has a significant reactive part (capacitive) which is due to an inertial origin of the electron-heating non-linearity. Thus, measuring the IF response of an HEB device in the frequency range wider than f_1-f_2, one can find parameters C and τ_0 and estimate the mixer bandwidth.

The validity of Eqs. (1)-(3) for superconducting HEB devices was demonstrated [6,8,9] for phonon-cooled Nb HEB mixers [6,8,9]. The first experimental verification of the above expressions for the impedance involved only scalar measurements, so only the impedance modulus was measured [9]. Both real and imaginary parts of the impedance were measured later [6]. The results agreed very well with the bandwidth measurements and the theory. Since the bandwidth in a Nb phonon-cooled mixer is about 100 MHz, an accurate calibration of the cryogenic cable loss and dispersion was not required, and straightforward vector S_{11} (ie: reflection) measurements were performed in Ref. 6. The device was simply connected to the end of a coaxial cable and inserted into a dewar. Compensation of the phase shift caused by the cable was done by setting the temperature below T_c, *so* the device was in a superconducting state and the impedance was assumed to be zero. Unfortunately, this approach did not work well for NbN HEB mixer whose bandwidth was of the order of 1 GHz [10]. The parasitic inductance due to wire bonds and capacitance across the device bridge as well as cable loss could not be characterized well enough at high frequencies. As a result, the quality of the data[10] was poor.

In this paper we use a different approach which is free of drawbacks of the S_{11} technique and allows for *in situ* calibration over a very wide frequency range. The schematic of the microwave circuit we used, is shown in Fig. 1 An HP8510 network analyzer was used, and no calibration procedure was performed on this instrument prior to the device measurements. The device was wire-bonded between two pieces of a microstrip line each ≈ 12 mm long. The microstrip lines were soldered to SMA connectors. The rest of the circuit was made with coaxial elements. We used 20 dB O-18 GHz attenuators to simulate 50 Ω terminations. The incident test power was greatly attenuated to avoid any perturbation of the device operating point. After passing through the microstrip test fixture, the test signal was amplified. Achieving sufficient signal-to-noise was very challenging in our case since the bolometers were very sensitive and even a 1 nW test power made a noticeable effect. Thus a cooled wideband amplifier (0.2-18 GHz) with a noise temperature of 80 K was used to achieve the desired results.

Figure 2 shows the equivalent circuit model used to analyze the data. The parasitic impedance of the microstrip fixture is represented by a simple series impedance Z . For very high frequencies a shunt capacitance across the device may become important. fn this case a parallel parasitic impedance $Z_{||}$ needs to be included in the model. Determination of $Z_{||}$ requires a calibration for which the device impedance $Z = \infty$; i.e. the device needs to be destroyed for this in-situ calibration. However, in the frequency range which we used (up to 4 GHz) it was not necessary. All calibrations were made *in situ*, i.e. the device was always in the dewar, and we only changed its state by applying more dc current or setting the temperature either below $T_c \approx 6$ K or above. Thus, only through calibrations were done when the device was either in the superconducting state (Z= O) or in the normal state ($Z = R_n$, where R_n *is* the normal metal resistance which is purely resistive and frequency independent in the range of interest). Having made three measurements of the throughput signal (the above two calibrations and the measurement at the operating point) for the same test power, one can find the complex device impedance vs frequency.

The complex magnitude of the transmitted signal at the output(see Fig. 2) is:

for the superconducting state:

$$v_s = v_0 \frac{Z_0}{Z_p + 2Z_0}, \tag{4}$$

for the normal state:

$$v_n = v_0 \frac{Z_0}{Z_p + 2Z_0 + R_n}; \tag{5}$$

and for the operating point:

$$v = v_0 \frac{Z_0}{Z_p + 2Z_0 + Z}. \tag{6}$$

Normalizing the transmission in the normal state and at the operating point to the transmission in the superconducting state, one can obtain an explicit relationship between the device impedance and the experimental S-parameters $S_{21}^s = v/v_s$ and $S_{21}^n = v/v_n$:

$$Z = R_n \frac{S_{21}^n}{S_{21}^s} \frac{1-S_{21}^s}{1-S_{21}^n}. \tag{7}$$

To test the above concept we used a diffusion-cooled HEB device which was made from a 12 nm thick high-quality Nb film. The device was a planar microbridge 0,3 μm long and 0.15 μm wide patterned by means of e-beam lithography between gold contact pads. The details of the device fabrication are given elsewhere[11]. For the impedance measurements the device temperature was adjusted to obtain a monotonic current-voltage characteristic (see Fig. 3) with a large differential resistance. Since the range of the frequency dependent effect is of the order of R_d-R (see Eq.(2)), it was important to make this difference of at least several tens ohms.

Figure 4 shows the de-embedded device impedance in a 4 GHz frequency range. The device normal resistance R_n = 60 Ω was used in Eq. (7) for the calculations. Also shown is a best fit to the data by Eq. (3) (solid line). The device resistance R = 44 Ω and differential resistance R_d = 80 Ω determined from the current-voltage characteristic (Fig. 3) were used. These values give C = 0.29 and the ratio f_2/f_1 = 1.8. Therefore, only one fitting parameter had to be used in Eq. (3). The best fit for the real part of the impedance gave f_1 = 1.0 GHz and f_2 = 1.8 GHz. Using Eq. (1) the mixer bandwidth Δf_{3dB} was estimated to 1.3 GHz. The agreement between the experimental and calculated imaginary parts was slightly poorer, though still satisfactory. We believe better signal-to-noise ratio could provide more accurate calibration and, hence, better coincidence of the experiment and the theory.

The parasitic impedance which was calibrated out in the course of measurements was found from the following expression:

$$Z_p = \frac{R_n S_{21}^n}{1 - S_{21}^n} - 2Z_0 \tag{8}$$

The frequency dependence of Z_p is shown in Fig. 5. One can see that Z_p is essentially inductive and is most likely associated with the wire-bonds.

In order to estimate the accuracy of the bandwidth prediction from the impedance measurements, we directly measured the bandwidth for a similar device integrated with a submillimeter twin-slot antenna. The details of the mixer device design are given in Ref. 1. The measurements were performed using two tunable 600 GHz monochromatic sources. One was a ×6-multiplier fed by a 100 GHz Gunn oscillator, the other was a backward-wave (BWO) oscillator which served as the LO. The radiation from both sources was combined in a single beam using a thin dielectric beamsplitter. The frequency of the LO was smoothly tuned. At each frequency point the LO power was adjusted by an attenuator in order to maintain the same bias current at the operation point. This ensured a frequency independent optical coupling to the mixer. The mixer operating temperature was chosen lower than in the impedance measurement, but the LO power level was adjusted to create a current-voltage characteristic very similar to that of Fig. 3. The IF signal at the mixer output was coupled out of the cryostat through a stainless steel cable, then amplified with two 30-dB wideband amplifiers and measured with a spectrum analyzer. The readout system was calibrated with a precise frequency synthesizer and the error from the calibration correction was 0.5 dB.

Figure 6 shows the mixer response vs intermediate frequency. The experimental points fit well to a one-pole spectral curve with a cut-off frequency of 1.2 GHz. This value is in excellent agreement with the above estimate of the bandwidth from the impedance data.

In summary, wc have demonstrated that a transmission impedance measurement technique can be a useful tool for characterization of the bandwidth of HEB superconducting mixers. The advantages of the S_{21} technique over alternative S,, techniques is a much wider frequency range and the availability of precise in-situ calibration. The proposed techniques is expected to work up to at least ≈20 GHz.

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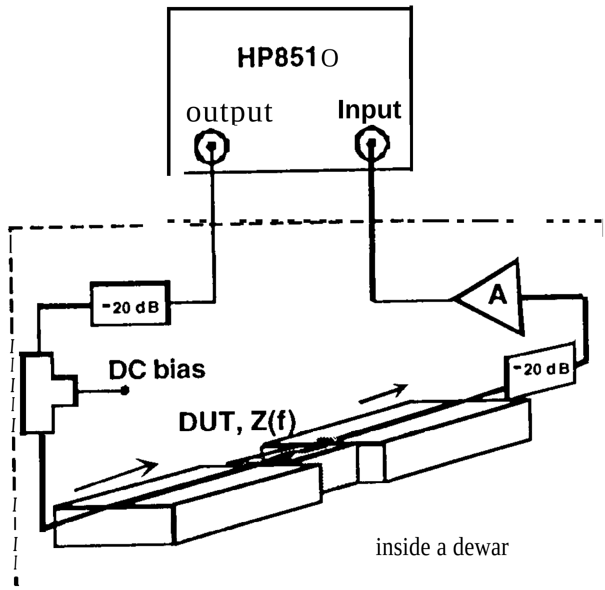

Fig. 1. The schematic of the impedance set-up.

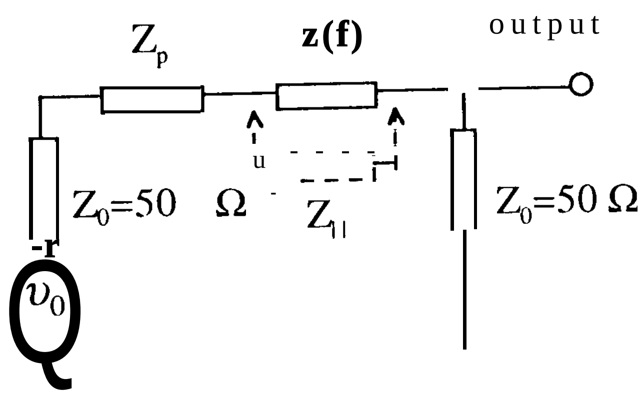

Fig. 2. The equivalent circuit

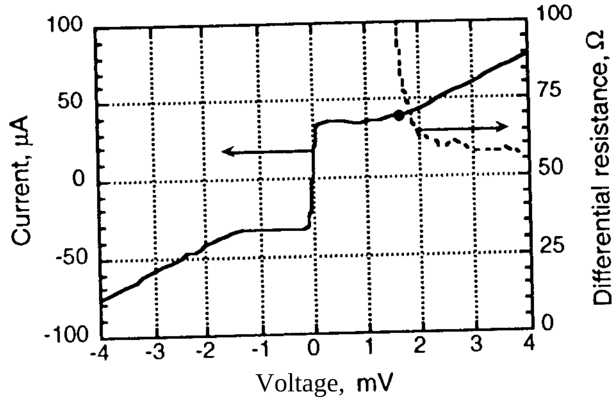

Fig. 3. Current-voltage characteristic.

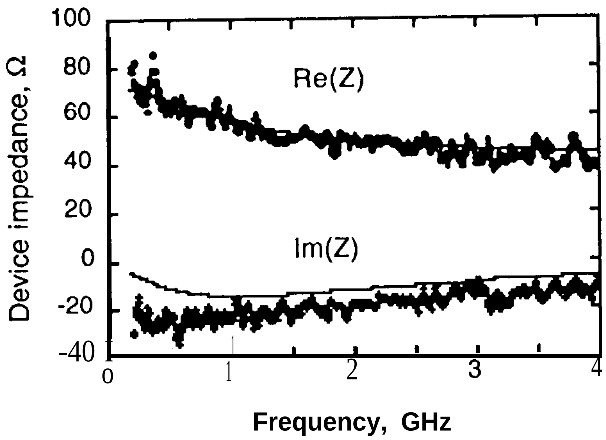

Fig. 4. Frequency dependence of the HEB impedance.

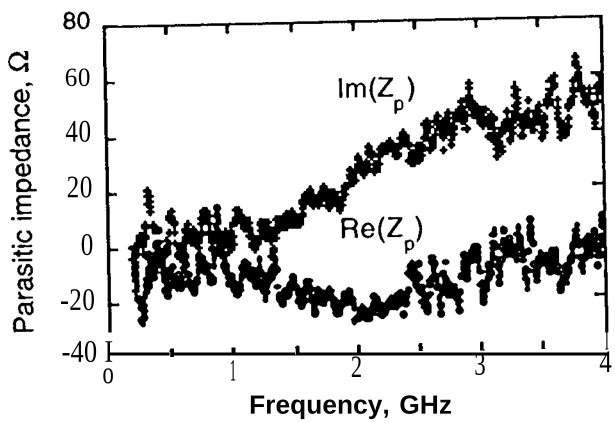

Fig. 5. Frequency dependence of the parasitic impedance.

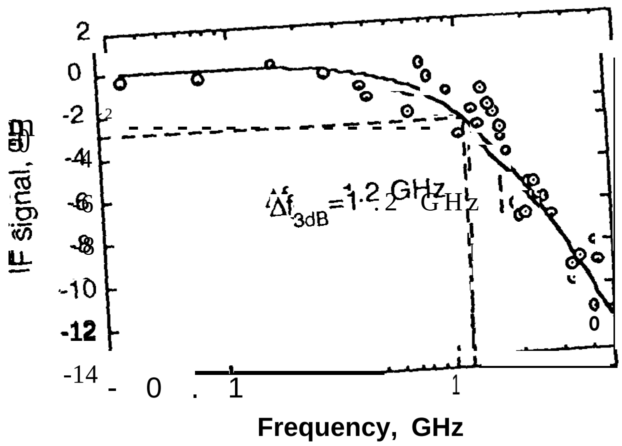

Fig. 6. Mixing signal of the HEB device using two 600 GHz monochromatic sources.