Title
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Journal
Applied Physics Letters, 59(18)

Authors
Mees, J.
Nahum, M.
Richards, P.L.

Publication Date
1991-07-01

LBL-31100
UC-364
Preprint

Center for Advanced Materials
CAM

Submitted to Applied Physics Letters

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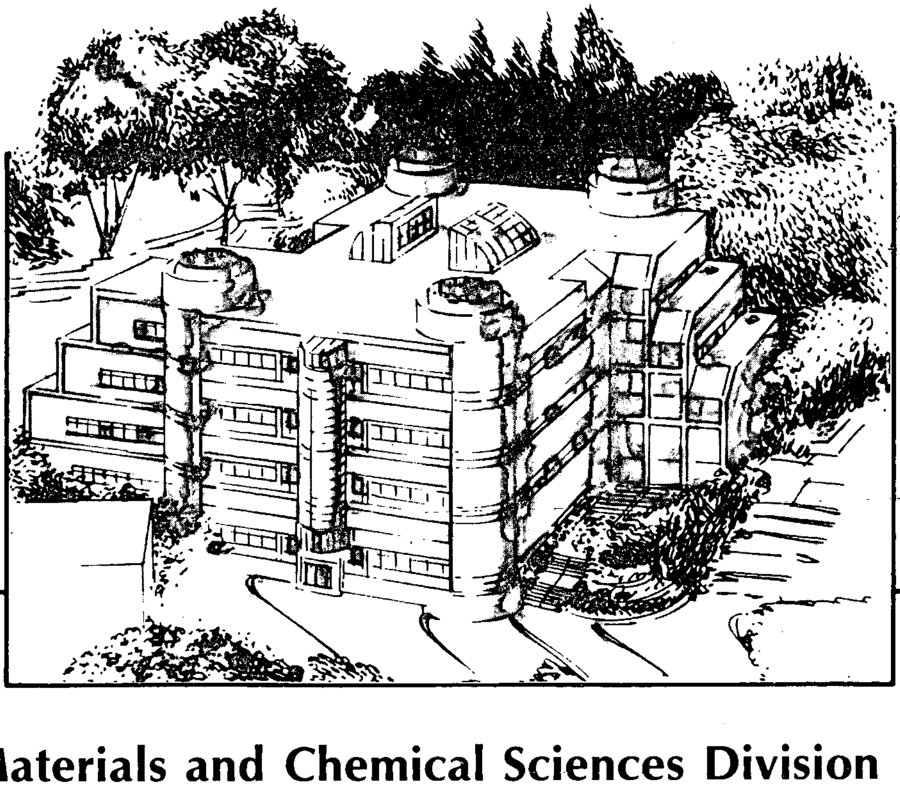
July 1991



Materials and Chemical Sciences Division

Lawrence Berkeley Laboratory • University of California

ONE CYCLOTRON ROAD, BERKELEY, CA 94720 • (415) 486-4755

Prepared for the U.S. Department of Energy under Contract DE-AC03-76SF00098

LBL-31100

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J. Mees,[a] M. Nahum, and P.L. Richards
Department of Physics, University of California,
and Materials Sciences Division,Lawrence Berkeley Laboratory,
Berkeley, California 94720

[a] present address:

Max-Planck-Institut für Radioastronomie
Auf dem Hügel 69, 5300 Bonn 1, Germany

The authors gratefully acknowledge many fruitful discussions with C. A. Mears. This work was supported by the Director, Office of Energy Research, Office of Basic Energy Sciences, Materials Sciences Division of the U.S. Department of Energy under Contract No. DE-AC03-76SF00098, and by the German foundation Friedrich-Ebert-Stiftung (for J. Mees).

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J. Mees,[a] M. Nahum, and P.L. Richards
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Auf dem Hügel 69, 5300 Bonn 1, Germany

Abstract

We propose a novel antenna-coupled low T_c superconducting bolometer which makes use of the thermal boundary resistance and the trapping of quasiparticles at metal-superconducting interfaces. A thin strip of superconductor, whose temperature is regulated at the midpoint of its resistive transition, serves both as a resistive load to thermalize the infrared current from the antenna and as a thermometer to measure the resulting temperature rise. Calculations give an NEP $\approx 7\times10^{-16}T^{5/2}$ WHz$^{-1/2}$ and a time constant $\tau \approx 10^{-8}T^{-2}$ s for a 2x2 μm^2 thermometer area at temperature T (K). Designs for efficient on-chip RF matching and filter networks with well defined bandpasses are presented. These detectors can be used to make frequency-multiplexed array receivers for astronomical observations at near millimeter wavelengths.

Bolometers are well known as very sensitive direct detectors for infrared and millimeter waves and are widely used in ground and space-based astronomical observations. Although excellent performance has been available with bolometers operated at ^{3}He temperatures[1] and at 100 mK,[2] there remain opportunities for improved bolometer sensitivity at millimeter and submillimeter wavelengths.

We propose to make very sensitive antenna-coupled superconducting bolometers at LHe and lower temperatures. Efficient radiation coupling is achieved by means of a planar lithographed antenna. A thin strip of superconductor whose temperature is regulated at the midpoint of its resistive transition serves both as the resistive load to thermalize the infrared current and as a sensitive thermometer to measure the resulting temperature rise. Thermal isolation is provided by the thermal boundary resistance and by the reflection of quasiparticles at the interface between a metal and a superconductor. All of the components are deposited directly on the substrate and can be produced in arrays using standard photolithographic techniques. The thermally active volume of this bolometer is $\approx 1\ \mu m^3$ compared with $\approx 10^8\ \mu m^3$ for more conventional millimeter wave bolometers. This letter will focus on two subjects. The first is the thermal design and optimization of the bolometer.[3] The second is the efficient RF coupling of the thermally active region to the antenna and the design of filter networks with well defined bandpasses.

As will be discussed, the sensitivity of the bolometer is ultimately limited by its thermal isolation from the environment. In our geometry the absorbed power is dissipated via two paths. The first is direct heat flow from the superconducting thermometer into the dielectric substrate. At temperatures above ≈ 10 K the thermal conductance is dominated by the bulk spreading resistance arising from the thermal conductivity of the substrate, as is the case for the high T_c superconducting microbolometer.[4] At the lower temperatures of interest here, the thermal boundary resistance contributes significantly to the thermal isolation, and below several Kelvin controls the heat flow. Acoustic mismatch theory has been very successful in explaining thermal boundary resistances at low temperatures.[5] According to this theory, the thermal conductance is given by

$$G=AT^3/B, \qquad (1)$$

where A is the contact area, and T is the temperature. The parameter B depends on the densities and sound velocities of the materials and is usually $\approx 20\ K^4cm^2/W$.

The contact area between the superconducting thermometer and the antenna terminals provides an alternate path for heat dissipation. In addition to the phonon conduction process discussed above, there exists the possibility that unpaired electrons in the superconducting thermometer (which is maintained at the center of the resistive transition) could also transfer energy across this interface. However, if the antenna is made of a superconductor whose T_c is higher than the operating temperature of the thermometer, then the Andreev reflection of electrons at this interface traps the quasiparticles in the active region, and hence reduces their contribution to the thermal conductance to a negligible value.[6]

If we model the thermal circuit as a film of heat capacity C coupled to the substrate through a boundary conductance G, then the thermal time constant is $\tau = C/G$, and for a constant current bias I, the voltage responsivity is[7]

$$S = \frac{IdR/dT}{G(1 + 4\pi^2 f^2 \tau^2)^{1/2}} \,. \qquad (2)$$

Here dR/dT is the temperature coefficient of the resistance and f is the modulation frequency. The noise equivalent power (NEP) has contributions from phonon noise, Johnson noise, and amplifier noise.[7] In the case discussed here, the responsivity is large enough that the sensitivity will be limited by the thermal fluctuations in the superconducting thermometer. Consequently,

$$\mathrm{NEP} = (4k_B A B^{-1} T^5)^{1/2}/\eta, \qquad (3)$$

where η is the optical efficiency. Candidate materials for thermometers include bcc-Ta at 4.2 K, β-Ta or α-Ti at 500 mK, or alloys such as Mo/Ge and Nb/W at lower temperatures.[8] For a 2x2 μm square film with transition width $\Delta T = 0.1T_c$ and a 50% optical efficiency we obtain $S \approx 6\mathrm{x}10^5 T^{-4}$ V/W, and $\mathrm{NEP} \approx 7\mathrm{x}10^{-16} T^{5/2}$ $\mathrm{WHz}^{-1/2}$ up to $2\pi/\tau \approx 6\mathrm{x}10^8 T^2$ Hz. Because this speed of response is much faster than is required for most applications, and because SQUID readouts have negligible noise for the impedances used here, it is possible to consider novel operation modes for this bolometer. The dynamic range for example can be greatly enhanced by using bias current feedback to keep the bolometer temperature constant. The output would then be obtained from the modulation of the feedback current at the signal chopping frequency.

Because the dimensions of the thermally active region are much smaller than the wavelength to be detected, a planar lithographed antenna can be used to provide efficient coupling. Self-complementary log-periodic or log-spiral antennas are very broadband and have a frequency independent real antenna impedance $Z_{ant} = 377[2(1 + \varepsilon)]^{-1/2}\ \Omega$ that depends only on the dielectric constant ε

of the substrate.[9] When deposited on quartz, $Z_{ant} \approx 120\ \Omega$. Since a planar antenna located on a dielectric surface radiates primarily into the dielectric, the signals are introduced through the back surface of the dielectric which is often placed on the back side of a dielectric lens.[9] Another approach is to place the bolometer in a waveguide and to couple it with thin film elements similar to those used for SIS mixers. The ratio of the power coupled into the thermometer to the incident power at the antenna terminals C_{RF} can be obtained in the usual way from antenna and thermometer impedances and is unity when the antenna impedance equals the complex conjugate of the thermometer impedance.[10]

It is necessary to understand the RF properties of the thermometer in order to design an efficient match to the antenna. Some information can be obtained from the simple two fluid model which gives a surface impedance[11]

$$Z_S = \frac{i\omega\mu_0 \coth[t(1/\lambda^2 + 2i/\delta^2)^{1/2}]}{(1/\lambda^2 + 2i/\delta^2)^{1/2}}. \qquad (4)$$

Here $\lambda = \lambda_0[1-(T/T_c)^4]^{-1/2}$ is the superconducting penetration depth, $\delta = [2/\mu_0\sigma_n(T/T_c)^4\omega]^{1/2}$ is the classical skin depth calculated for the normal carriers below T_c, σ_n is the conductivity in the normal state, t is the film thickness, and ω is the RF frequency. This model should provide a good description far enough below T_c that pair breaking is not important, but where kinetic inductance effects must be considered. It is also valid in the normal state if we set $T/T_c = 1$ and $\lambda = \infty$. When the film thickness $t \ll \delta$, the normal state surface impedance from Eq. (4) is just the dc resistance per square of the strip $Z_S = (t\sigma_n)^{-1}$. If we assume that sample inhomogeneities determine the transition width and that the bolometer is biased at the midpoint of the transition, then all photons with energies larger than the gap at $\approx T_c - \Delta T/2$ can break pairs. A complete understanding of Z_S at the midpoint of the transition for higher frequencies would require a full Mattis-Bardeen calculation including pair breaking for an inhomogeneous superconductor with finite transition width. In practice, it should be sufficiently accurate to use some value of Z_S that is intermediate between the predictions of Eq. (4) above and below T_c.

The most straightforward thermometer design is a simple strip of length l, width w, and thickness t, located directly between the antenna terminals. The impedance of such a strip is

$$Z_{bolo} = Z_S l/w + Z_L \,, \tag{5}$$

where Z_L is the impedance due to the geometrical inductance of a rectangular thin strip over a dielectric half plane,[12]

$$Z_L = 2\times10^{-7} i\omega l \left[\ln\frac{l}{(w+t)} + 1.193 + 0.2235\,\frac{(w+t)}{l} \right], \tag{6}$$

with l, w, and t in μm, and Z_L in ohms. When typical superconductors with large values of σ_n are used, the RF resistance of the thermometer will be much lower than the ≈ 120 Ω real terminal impedance of typical self-complementary antennas or the even higher impedances available for matching to waveguides. This mismatch can be minimized by the use of low conductivity metals or by the use of long thin films. For long films, the geometrical inductance of the isolated thermometer strip can become significant above ≈ 100 GHz, which further complicates the matching problem. Some of these difficulties can be avoided by integrating the thermometer in a superconducting microstrip line.

Because the thermometer may not be small enough that it will appear as a lumped element at RF-frequencies, we will treat it as a lossy microstrip. The bolometer impedance Z_{bolo} can be calculated from[11,13]

$$Z_{bolo} = (Z/Y)^{1/2} \tanh \gamma l, \tag{7}$$

where

$$Z = (i\mu_o \omega d + Z_{S_a} + Z_{S_b})/kw, \tag{8}$$

is the series impedance per unit length of the transmission line,

$$Y = i\varepsilon_o \varepsilon \omega kw/d, \tag{9}$$

is the shunt admittance per unit length of the line, and

$$\gamma = (ZY)^{1/2}, \tag{10}$$

is the propagation constant. Here d is the dielectric thickness, Z_{S_a} and Z_{S_b} are the surface impedances of the lower groundplane and upper thermometer strip respectively, k is the fringing factor of the stripline,[14] and ε is the dielectric constant. When $\omega/2\pi$ is small compared to the energy-gap frequencies of the groundplane ($\approx$ 740 GHz for Nb at T = 0 K), Z_{S_a} is well approximated by Eq. 4. The surface impedance of the thermometer strip can be described by the normal state limit of Eq. 4, as previously discussed.

An additional advantage of integrating the thermometer in a superconducting microstrip line is that it becomes possible to include efficient matching networks, made from transmission line elements in the bolometer design. When microstrip lines produced by conventional lithographic techniques are used for the network elements, the optimum range for the thermometer resistance is between 2 Ω and 30 Ω. Combining microstrips with co-planar waveguides would extend the useful resistance range up to more than 100 Ω, and therefore enlarge the range of applications. The benefits of such networks are not restricted to badly matched thermometers. A major advantage of this configuration is that lithographed filter elements can be incorporated with relative ease. These lithographed filters can have superior performance over conventional filters which tend to exhibit high frequency leakage at submillimeter wavelengths.

As a specific example of a poorly matched thermometer we consider a thermometer element made from a film of bcc-Ta with $T_c = 4.2$ K, $l = 10$ μm, $w = 2$ μm, and $t = 1000$ Å. The thermometer is incorporated in the microstrip configuration with a $d = 6000$ Å thick SiO_2 dielectric and a thick Nb ground plane. For this specific example the real part of Z_{bolo} is 2.7 Ω above T_c, which presents a very poor match to the antenna. The imaginary part of Z_{bolo} is 4.7 Ω above T_c and 6.6 Ω in the kinetic inductance region at 300 GHz. For calculations we use $Z_{bolo} = (2.7 + i5.4)$ Ω at 300 GHz. Using conventional microwave theories[15] we have designed quarter-wave transformers with several elements. Fig. 1 shows the calculated coupling coefficient C_{RF} versus frequency for a) the above thermometer on a ground plane but without tuning elements, b) with one transforming element and c) one transforming element plus an additional stub that cancels the reactance of the thermometer. As a further demonstration of the power of these techniques we have designed 7-element Chebyshev bandpass filters with 30% and 10% bandwidths.[15,16] These are shown in d) and e) respectively. A typical filter layout is shown at the right. These calculations were done with lossless microstrip elements coupled to the lossy thermometer. Little change is seen in the filter characteristics when microstrip losses as large as 600 dB/m are included in the calculations. Losses below 10 dB/m have been measured in Nb trilayer technology at 100 GHz.[17]

Potential applications of these devices can be understood by considering applications of conventional composite bolometers. They are used in ground and space-based telescopes as sensitive broadband detectors for measurements of molecular line emission from interstellar gas and planetary atmospheres, continuum emission from point sources, and for measurements of the anisotropy of the cosmic microwave background. The estimated sensitivity of the proposed bolometers can be higher than that of the conventional ones. Also, it is possible to construct multiplexed systems by designing on-chip networks with different bandpass filters fed from a single antenna. Using well developed reproducible lithographic techniques it should be possible to make powerful array receivers with multiple spectral bands. Related design concepts can also be used to provide efficient coupling and RF filtering for SIS quasiparticle detectors and mixers.

The authors gratefully acknowledge many fruitful discussions with C. A. Mears. This work was supported by the Director, Office of Energy Research, Office of Basic Energy Sciences, Materials Sciences Division of the U.S. Department of Energy under Contract No. DE-AC03-76SF00098, and by the German foundation Friedrich-Ebert-Stiftung (for J. Mees).

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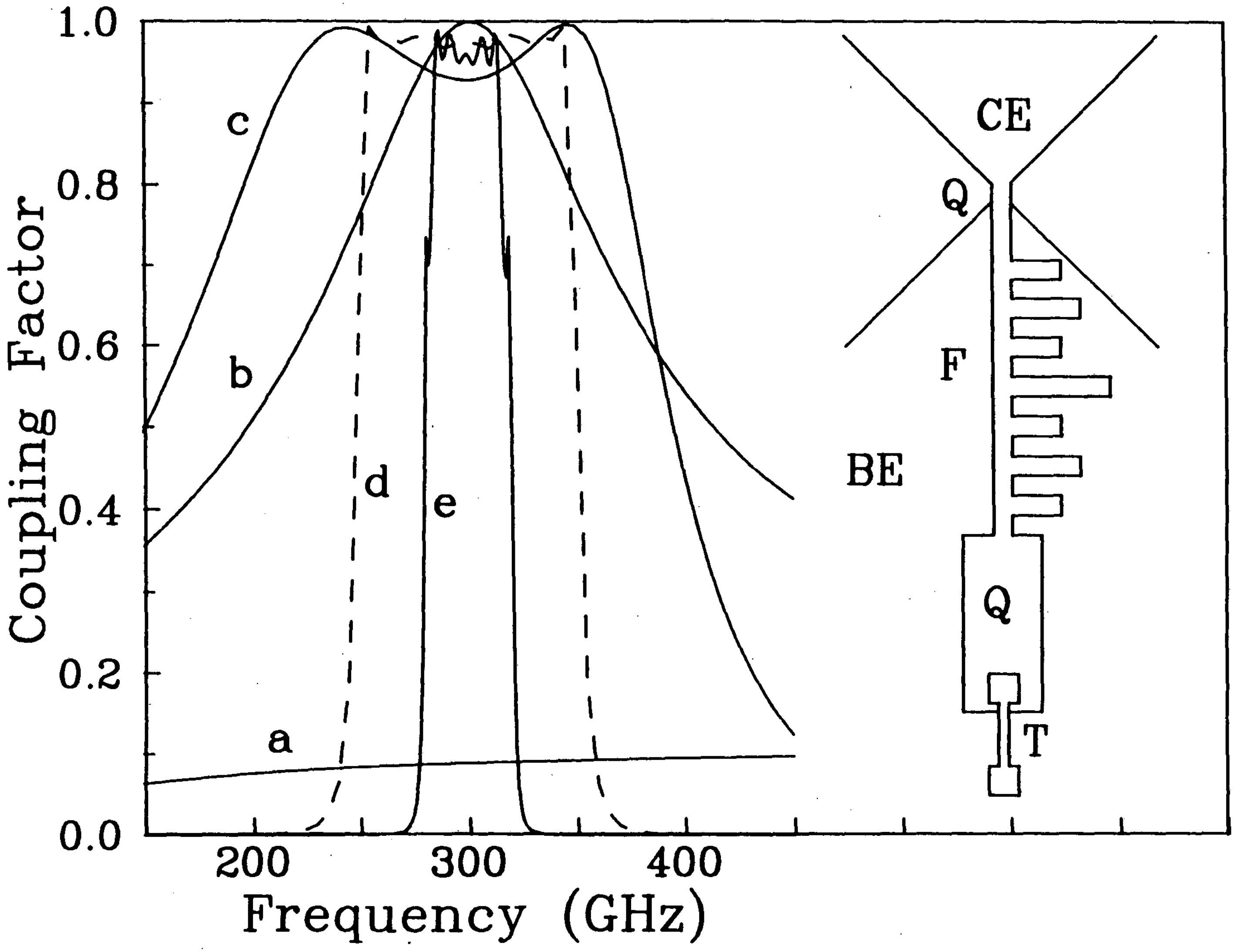
Figure Caption

Fig. 1. The frequency dependent RF-coupling factor for several examples: a) For the thermometer on a ground plane but without any matching network. b) The same thermometer with one quarter wave transforming element. c) Same as b) plus an open-ended stub which cancels the reactance of the bolometer. d) Thermometer with a 7-element Chebyshev filter and a 30% bandwidth. e) Same as d) but with a 10% bandwidth. A typical layout of a Chebyshev filter with transforming sections is shown at right. The antenna terminals are labeled base electrode BE, and counter electrode CE. The other elements are two impedance transformers Q, the filter F and the thermometer T. The dielectric layer which covers the base electrode is not shown.


1.0
0.8
0.6
0.4
0.2
0.0
Coupling Factor
200
300
400
Frequency (GHz)
a
b
c
d
e
CE
Q
F
BE
Q
T

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LAWRENCE BERKELEY LABORATORY
CENTER FOR ADVANCED MATERIALS
1 CYCLOTRON ROAD
BERKELEY, CALIFORNIA 94720