Superconducting terahertz mixer using a transition-edge microbolometer

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We present a new device concept for a mixer element for THz frequencies. This uses a superconducting transition-edge microbridge biased at the center of its superconducting transition near 4.2 K. It is fed from an antenna or waveguide structure. Power from a local oscillator and a rf signal produce a temperature and resulting resistance variation at the difference frequency. The new aspect is the use of a very short bridge in which rapid (<0.1 ns) outdiffusion of hot electrons occurs. This gives large intermediate frequency (if) response. The mixer offers ≈ 4 GHz if bandwidth, $\approx 80\ \Omega$ rf resistive impedance, good match to the if amplifier, and requires only 1–20 nW of local oscillator power. The upper rf frequency is determined by antenna or waveguide properties. Predicted mixer conversion efficiency is 1/8, and predicted double-sideband receiver noise temperatures are 260 and 90 K for transition widths of 0.1 and 0.5 T_c, respectively.

During the past decade there have been significant advances in heterodyne receivers in the frequency range 0.1–2 THz. These advances have been due to developments in superconductor-insulator-superconductor (SIS) mixers,[1,2] cooled Schottky receivers,[2] and integrated antennas.[3] SIS tunnel junctions have been successfully employed to 500 GHz in waveguide receivers. For these and higher frequencies, many designs favor antenna-coupled detectors, due to increasing difficulty with waveguide components and the potential ease of producing focal plane arrays of antenna-coupled devices. NbN films offer a gap frequency above 700 GHz, but have so far exhibited problems with rf loss. Needs for improved THz heterodyne receivers are found in spaced-based,[4] airborne, and balloon-based observatories, and in laboratory studies in diverse fields.

Other approaches for submillimeter wavelength mixers include SIN (N=normal metal) tunnel junctions,[2] Josephson mixers, hot-electron superconductor devices using Nb[5] or NbN[6] films of area $\approx 10\ \mu m^2$, and hot-electron semiconductor devices.[7,8] The hot-electron superconductor devices studied to date[5,6] use electron-phonon coupling to cool the electrons; the electron-phonon interaction time determines the response time. The experiments on Nb devices[5] have used an intermediate frequency (if) ≈ 40 MHz. Past hot-electron semiconductor devices[7] had limited if bandwidth, though this situation may improve.[8]

The microbolometer proposed here is a superconducting transition-edge detector[9,10] using a superconducting microbridge (e.g., Nb with[5] $T_c \approx 4.4$ K, $L=0.2\ \mu m$, $W=0.05\ \mu m$, thickness $d=0.01\ \mu m$, and $R_n=80\ \Omega=2R$). It is biased at a temperature T and resistance R near the midpoint of its transition, whose width is ΔT; see Fig. 1. The leads are thicker nonsuperconducting films of high conductivity. When two rf signals are coupled to the device, its temperature can respond rapidly enough to follow the power variation at the difference (intermediate) frequency. This fast thermal response is due to rapid diffusion of the hot electrons out the ends of the strip into the thicker leads which are at a temperature $T_0 \approx 4.2$ K. The thick leads connect to the rf coupling structure (for a waveguide system) or to the antenna (for an antenna-coupled system). The thick leads serve to heat sink the ends of the bridge at T_0 and to couple in the rf signals with low loss. The new aspect of the present letter is the use of a very short bridge to achieve a large if response using hot-electron outdiffusion, and the development of a basic prediction for receiver noise.

The microbolometer mixer proposed here has a number of attractive features: the impedance is resistive, with negligible capacitance and inductance;[10] the device resistance can match that of the antenna or waveguide struc-

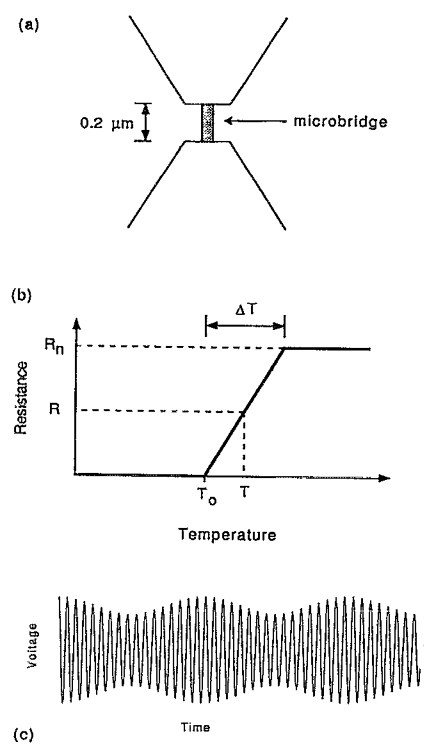

FIG. 1. (a) Layout of microbridge mixer; bridge is shaded, thick pads are unshaded; (b) Resistive transition with bridge biased at R and T; (c) Voltage wave form $V(t)=V_0\cos(\omega_{lo}t)+0.2\ V_0\cos(\omega_s t)$, with $\omega_s=1.06\ \omega_{lo}$; shows amplitude modulation of the power, averaged over a few rf cycles.

ture and the if amplifier; the mixer conversion efficiency is fairly high, up to 1/8; the if can be ≈ 4 GHz, a practical requirement; and the local oscillator (lo) power required is small, $\approx$nW. Finally, the device can be fabricated on a standard quartz substrate.

To predict receiver performance, we first calculate the response at the if. The voltage responsivity S (V/W) to rf power dissipated in the bridge is given by

$$S=I(dR/dT)/G(1+\omega_{\rm if}^2\tau^2)^{1/2}=S_0/(1+\omega_{\rm if}^2\tau^2)^{1/2} \quad (1)$$

with I the bias current, R the *electrical* resistance, G the *thermal* conductance for heat flow to the thick leads, $\omega_{\rm if}$ the angular frequency of the amplitude modulated power, and τ the *thermal* response time. This is given by[11] $\tau=C/G$, with C the electron specific heat for the bridge.[12] Cooling by outdiffusion of hot electrons dominates over other cooling paths, such as electron-phonon cooling, since $\tau_{\rm diff} \approx (L/4)^2/D \approx \tau$ is less than the electron-phonon inelastic time, $\tau_{\rm ep}$.[13] (D is the diffusion constant, ≈ 1 cm^2/s for a Nb film 100-Å thick.[5]) $\tau_{\rm ep} \approx 10^{-8}$ T^{-2} in s.[5]

The if response extends to a -3 dB frequency of $f_{\rm if,3dB}=(2\pi\tau)^{-1}$. The proposed design has a time constant of[14]

$$\tau=C/G=0.04 \text{ ns}. \quad (2)$$

This gives -3 dB if response to 4 GHz. For Nb electron-phonon hot-electron mixers, $\tau=1$ ns (3 dB frequency ≈ 160 MHz) at 4.2 K.[5] In most THz astronomical applications, an if bandwidth $\geqslant 1$ GHz is desired.

We next calculate the mixer conversion efficiency, η_M, which is the if power coupled into a matched if load divided by the rf signal power coupled into the mixer. The local oscillator and signal produce a voltage in the mixer $V(t)=V_{\rm lo}\cos(\omega_{\rm lo}t)+V_s\cos(\omega_s t)$. The instantaneous power is[12] $P_{\rm inst}=V(t)^2/R_n$. The mixer's thermal response is too slow to follow at frequencies of order $2\omega_s$ and $2\omega_{\rm lo}$. The thermal response averages over many rf cycles, responding to the power

$$P(t)=P_{\rm lo}+P_s+2(P_{\rm lo}P_s)^{1/2}\cos\omega_{\rm if}t, \quad (3)$$

with $P_{\rm lo}=V_{\rm lo}^2/2R_n$, $P_s=V_s^2/2R_n$, and $\omega_{\rm if}=\omega_{\rm lo}-\omega_s$. We assume $P_s \ll P_{\rm lo}$. If $\omega_{\rm if}\tau<1$, we can use the dc responsivity S_0 from Eq. (1) to find the voltage at the if. With no if load connected this voltage is given as $V(t)_{\rm if}=S_0 2(P_{\rm lo}P_s)^{1/2}\cos\omega_{\rm if}t$. With a matched if load, $R_{\rm if}=R$, the if voltage is half this value. The dc power coupled into this matched if load is $(1/4)\langle V(t)_{\rm if}^2/R\rangle$, so

$$P_{\rm if}=S_0^2P_{\rm lo}P_s/(2R)$$
$$=2(\delta T_{\rm dc}/\Delta T)(\delta T_{\rm lo}/\Delta T)P_s \equiv \eta_M P_s. \quad (4)$$

We use $dR/dT=2R/\Delta T$ in deriving the above equation. We take the average temperature rise due to the dc current as $\delta T_{\rm dc}=I^2R/G=\Delta T/4$, and $\delta T_{\rm lo}=P_{\rm lo}/G=\Delta T/4$. With these values we find the mixer conversion efficiency is

$$\eta_M=1/8. \quad (5)$$

We have considered a sinusoidal rf signal, so η_M is a single-sideband (SSB) conversion efficiency. The SSB mixer conversion loss, L_M, is given as $-10\log\eta_M=9$ dB. Temperature increases of $\delta T_{\rm dc}$ and $\delta T_{\rm lo}=\Delta T/4$ are close to upper limits since the temperature increase at the center of the bridge is approximately $(3/4)\Delta T$.[12] The overall conversion efficiency is given by $\eta=\eta_{\rm rf}\eta_M\eta_{\rm if}$. The rf coupling efficiency to the mixer's rf resistance$=R_n$ of 80 Ω should be excellent; $\eta_{\rm rf}\approx 1$ for waveguide coupling. The if coupling efficiency to the mixer output resistance $R\approx 40$ Ω is also very good, $\eta_{\rm if}\approx 1$. The overall conversion efficiency for waveguide coupling can thus be $\eta\approx 0.1$, for a conversion loss$=10$ dB; for a planar antenna[3,15,16] or corner reflector,[2] we take $\eta\approx 0.05$.

A receiver using this microbridge heterodyne element will have three sources of noise: (1) intrinsic temperature fluctuations, given by $\delta T_n^2=4$ kT^2/G per Hz, rms at frequencies below $(2\pi\tau)^{-1}$, for the mixer itself; (2) Johnson noise; and (3) if amplitude noise, characterized by the amplifier noise temperature $T_{\rm if}$. Temperature fluctuations create a mean square voltage in the matched if resistor of $[I(dR/dT)/2]^2\delta T_n^2$. (The factor of 2 results from voltage division into the load.) The noise power in the if load is thus given as

$$P_{n,\rm if}=kB[T(T/\Delta T)+T+T_{if}] \quad (6)$$

with $k=$Boltzmann's constant and B the bandwidth; we have set $I^2(dR/dT)/G=1/2$. The double-sideband (DSB) receiver noise temperature, T_R, is the temperature of the rf source which will produce additional noise power $2kT_R\eta B$ at the if equal to the if noise power given by Eq. (6). (The factor of 2 here is for the DSB case.) We find

$$T_R(\text{DSB})=(2\eta)^{-1}[T(T/\Delta T)+T+T_{\rm if}] \quad (7)$$

$$\approx 5T[(T/\Delta T)+2] \quad (8)$$

with $\eta=0.1$, $T_{\rm if}=4$ K, and $T=4.4$ K. For a bridge transition width of $\Delta T=0.1T$, T_R (DSB)$=260$ K; for a transition width $\Delta T=0.5\ T$, T_R (DSB)$=90$ K.[17,18] A surprising conclusion from Eqs. (6) and (7) is that one desires a *broad* superconducting transition. This can be understood. Since δT_n^2 is fixed once G is set, a broad transition reduces the effect of thermal fluctuations. The conversion efficiency does not depend on ΔT, since the lo and dc power are adjusted to give a fixed temperature rise $\delta T_{\rm dc}=\delta T_{\rm lo}=\Delta T/4$. We summarize the device performance in Table I.

We summarize the performance of selected heterodyne receivers in the literature in Table II. SIS technology[15,19,20] is fairly mature for frequencies up to ≈ 350 GHz. LO power is modest for single SIS junctions, of order 10 nW. The Schottky devices[21] require $\approx$mW lo power. They use laser lo sources and corner-reflector mounts above ≈ 700 GHz. Above ≈ 500 GHz, our proposed hot-electron mi-

TABLE I. Device properties for Nb microbolometer mixer, for overall conversion efficiency $\eta=0.1$, appropriate for waveguide coupling.

$P_{\rm lo}$ for $\Delta T=0.1$ T	1 nW
$\Delta T=0.5$ T	5 nW
if response (-3 dB)	4 GHz
T_R(DSB) for $\Delta T=0.1$ T	260 K
$\Delta T=0.5$ T	90 K

TABLE II. Receiver performance with various mixer types. SIS devices are at 4.2 K; cooled Schottky at $\approx$20 K. if is $\gg$1 GHz unless noted. T_R is a measured value, except for Ref. 5.

Mixer	Frequency (GHz)	T_R(DSB) (K)	Coupling	Ref.
Superconductor				
Hot electron[a]	134	175	Waveguide	5
SIS,Nb	$\approx$230	50	Waveguide	19
	426	220	Antenna coupled	15
	492	172	Waveguide	20
Cooled				
Schottky	492	500	Waveguide	21
	800	1200	Corner reflector	b
	1900	4600	Corner reflector	c

[a]if=40 MHz; T=1.6 K; T_R is calculated value.
[b]H. P. Röser (private communication).
[c]A. L. Betz (private communication); T=77 K.

crobolometer mixer looks attractive even for a narrow transition width ΔT=0.1 T. Such a transition width is currently achievable. For the higher frequencies, the small lo power required by the microbridge may allow use of solid state sources to well above 1 THz. Below $\approx$500 GHz, our proposed device can improve on present and near-future SIS devices if we can achieve $\Delta T > 0.2$ T. Our computation of noise temperature does not include any contribution from quantum noise.[1] This will contribute an amount of order T_Q (DSB) $= hf/2k = 24$ K at 1 THz.

An issue in real designs is saturation by background rf. $P_{\rm sat}$ is equal to $(2/3)P_{\rm lo}$ with the dc and lo power levels described above. $P_{\rm lo}$=5 nW for ΔT=0.5 T if $\eta_{\rm rf}$=1. To prevent saturation, on-chip filters[10] may be used. Further, one could employ an array of four microbridges in series parallel. This configuration has the same resistance as an individual bridge, but a lo power of 20 nW.

The upper frequency limit for the microbolometer mixer is not established. Antenna coupling can work well, to at least 2 THz with much larger transition-edge detectors.[22] Good antenna efficiencies, $\approx$50%, have been measured at 16 THz ($\lambda = 19\ \mu$m).[16] The elastic scattering time in the bridge, $\tau_e \approx 10^{-15}$ s, sets an upper limit $\delta \approx 100$ THz on the signal frequency. Bolometric mixing in a different system, a superconducting point contact, can persist to very high frequencies.[23]

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[11] The relation $\tau = C/G$ is appropriate for a discrete thermal element; our device is a distributed thermal element. Simulations of the distributed system by M. Nahum (private commun.) show that the discrete-element analysis presented here matches Eq. (1) to better than 10%.
[12] The rf dissipation occurs in the bridge and is uniform along the length, because the frequency is above the gap frequency of the bridge. This leads to a parabolic profile for the temperature increase, with the peak temperature rise $=3/2$ the average rise. The profile is sharper than parabolic for the temperature rise due to dc dissipation, but we use the simpler parabolic law here also.
[13] The electrons in the bridge need to come to internal thermal equilibrium on a time scale shorter than $\tau_{\rm diff}$. The electron-electron inelastic time for thermal electrons at 4.4 K (E=0.3 meV) is $\approx$0.1 ns; see P. Santhanam and D. E. Prober, Phys. Rev. B **29**, 3733 (1984). The absorbed rf photon has energy 4 meV for ν=1 THz. The initially excited hot electron will share its energy in a time of order 0.01 ns.
[14] For *small* dc power dissipation the thermal conductance is given by $G_0 = \mathscr{L}T/R_{\rm eff}$(W/K), with $\mathscr{L}$ the Weidemann Franz constant, 2.5×10^{-8} W Ω/K^2 and $R_{\rm eff}$ the effective electrical resistance along the paths through which the heat flows out of the bridge, out the *two* ends; $R_{\rm eff} = R_n/12$. We find $G_0 = 1.5\times10^{-8}$ W/K at 4.2 K, for $R_n = 80\ \Omega$. The actual thermal conductance G with *finite* dc dissipation is smaller than G_0 due to self-heating effects. (Refs. 5 and 9) Thus, $G = 0.8\times10^{-8}$ W/K, with the self-heating parameter chosen to be $I^2(dR/dT)/G = 1/2$. The electron heat capacity is given approximately by $C = \gamma TV$, with $\gamma = 7\times10^{-4}$ J/K cm^{-3}, and V the microbridge volume in cm^3. We define G_0 for the average temperature increase with small dc dissipation.
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[17] Our predictions have been developed for $\Delta T \ll T$, but should apply semi-quantitatively for ΔT=0.5 T.
[18] With $\Delta T \ll T$ the dominant noise is due to temperature fluctuations. In this case, the signal and noise are *both* attenuated in the same manner above $f_{if,3db}$, as $(S/S_0)^2$; see Eq. (1). The receiver noise temperature will thus remain constant to above 10 GHz if. (N. R. Erickson, private communication)
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