Quasiparticle-tuned superconducting mixer

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We have successfully employed the recently observed quasiparticle admittance of a superconducting tunnel junction as the tuning element for a superconductor-insulator-superconductor mixer. This is the first demonstration of an electronic on-chip tuning element for a superconducting detector. We observe an increase in the dynamic resistance and an improvement in mixer gain and noise temperature as the dc bias voltage of the tuning junction array is varied. We compare our results with the quantum theory of mixing.

Mixers based on the SIS (superconductor-insulator-superconductor) tunnel junction are now the dominant technology for astronomical receivers in the millimeter-wave band and there is substantial effort to extend their operation into the submillimeter range and to construct imaging arrays.[1] Mechanical tuning elements in the form of waveguide backshorts have been used successfully to tune out the capacitance at frequencies up to 500 GHz, but become increasingly cumbersome above 200 GHz. On-chip thin-film inductors have provided tuning in some designs, but are not variable, so one cannot adjust the resonant frequency or compensate for variations in microfabrication.[2,3]

There have been previous attempts to obtain electronic tuning with a superconducting circuit. Use of the Josephson reactance as a tuning element has been proposed, but not tested with success.[4] The Josephson inductance is highly nonlinear, and could introduce large noise sidebands, as in Josephson effect mixers. We demonstrate in the present work an electronic on-chip tuning element where the *quasiparticle* susceptance provides a voltage controlled inductance.

The current flowing in a tunnel junction in the presence of an applied voltage has four components.[5] Two of these components are due to the flow of Cooper pairs (Josephson effect) and two are due to quasiparticles. In this letter we will concern ourselves only with the two quasiparticle currents, dissipative and reactive, and assume the two Josephson currents are suppressed by a magnetic field. For an applied voltage $V(t)=V_0+V_\omega \cos(\omega t)$, with V_0 the dc voltage, the real part of the admittance $Y(V_0,V_\omega)$ is the quasiparticle conductance G_Q and the imaginary part of the admittance is the quasiparticle susceptance B_Q:

$$G_Q(V_0)=\frac{1}{V_\omega}\sum_{n=-\infty}^{\infty} J_n(\alpha)[J_{n+1}(\alpha)+J_{n-1}(\alpha)] \times I_{\rm dc}(V_0+\hbar\omega/e) \tag{1a}$$

and

$$B_Q(V_0)=\frac{1}{V_\omega}\sum_{n=-\infty}^{\infty} J_n(\alpha)[J_{n+1}(\alpha)-J_{n-1}(\alpha)] \times I_{\rm KK}(V_0+\hbar\omega/e), \tag{1b}$$

where $\alpha=eV_\omega/\hbar_\omega$, $I_{\rm dc}(V_0)$ is the dc current at V_0 with $V_\omega=0$, J_n is the Bessel function of order n, and $I_{\rm KK}(V_0)$ is the Kramers–Kronig transform of $I_{\rm dc}(V_0)$.[6]

Figure 1 shows the quasiparticle branch of the dc I-V curve of a SIS tunnel junction along with the calculated quasiparticle conductance and susceptance at 90 GHz for a low input power. As seen there, close to the gap voltage the quasiparticle response of a tunnel junction can be modeled as a voltage-variable susceptance, $B_Q(V_0)=-j/\omega L(V_0)$, in parallel with a constant conductance. This susceptance and conductance have recently been experimentally measured.[7,8]

In order to resonate the capacitance of the SIS element at the frequency of operation, one needs to place the mixer in rf parallel with a tuning inductor. In our design we use a series array of four tunnel junctions as the mixing element. The inductance is provided by a separate series array

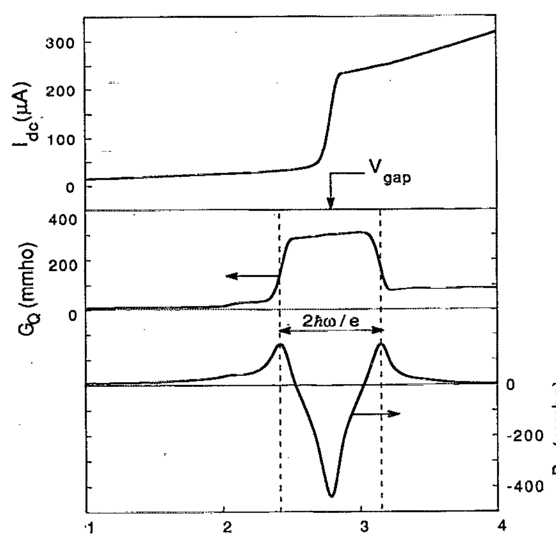

FIG. 1. dc I-V curve of a single tunnel junction (top) and the calculated quasiparticle conductance and susceptance (bottom) for a small input power ($\alpha<0.3$) at 90 GHz. The dotted lines show the voltages corresponding to a photon energy below and above the gap voltage.

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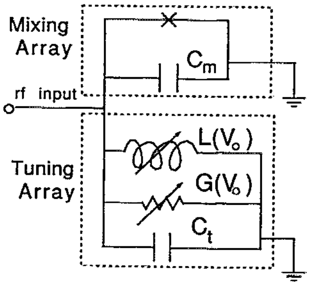

FIG. 2. The rf equivalent circuit. We have used a four junction array for mixing and a seven junction array for tuning.

of seven tunnel junctions which is referred to as the tuning array. This tuning array is independently dc biased and is in rf parallel with the mixing array (Fig. 2). The junctions of the tuning array have the same current density and area as the junctions of the mixing array. In order to decrease the absorption of photons from the input signal by the tuning array, we have used about twice as many tuning junctions as mixing junctions.

The waveguide mixer mount and the receiver system are described elsewhere.[3] Briefly, the microstrip-line circuitry consists of a $\lambda/4$ section which feeds the tuning array in parallel with the mixing array. The mixing array and the tuning array are terminated with separate 90° radial stubs. These provide two dc- and intermediate-frequency (if)-isolated broadband rf-ground terminations at nearly the same geometrical location. This allows the tuning array to be independently dc biased. By measuring the if output for different hot/cold load temperatures we can determine the noise and the gain of the receiver. The Nb/Al-O_x/Nb trilayer tunnel junctions used in this work each have an area of 4 μm^2 and a resistance $R_n = 11.5\ \Omega$.[9]

We have used several methods to test the performance of this tuning approach and have compared our results quantitatively with the quantum theory of mixing.[6] The shape of the photon-induced tunneling steps depends on the amplitude of the rf signal and the admittance presented to the device by the embedding circuitry. As one changes the embedding admittance from capacitive to inductive, the shape of the individual steps becomes better defined and the dynamic resistance in the middle of the first photon step increases (i.e., the step becomes flatter). For optimum receiver performance, the capacitive susceptance of the mixer should be resonated by the inductive embedding susceptance of the tuning array.[10]

Experimentally obtained *I-V* curves with tunneling steps at 90 GHz are shown in Fig. 3 for two different biasing conditions for the tuning array. When the tuning array is biased at the gap voltage (point *a* in the inset) the dynamic resistance of the mixing array in the middle of the first photon step is a factor of 2 larger than for bias voltage *b*; this is evident in Fig. 3. This shows that we have altered the embedding admittance seen by the *mixing* array by changing the bias voltage of the *tuning* array.

From the shape of the tunneling steps in Fig. 3 we calculate the embedding admittance seen by the mixing array.[11] The embedding admittance is

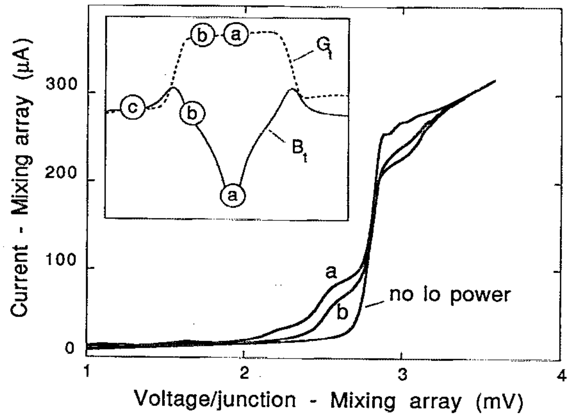

FIG. 3. Measured *I-V* curves of the mixing array for two different bias voltages on the tuning array. Curves *a* and *b* are with low power applied. The inset shows the conductance and the susceptance of the tuning array at bias voltages *a*, *b*, and *c* on the tuning array.

$$Y_{\text{emb}} = [G_t + G'] + i[\omega C_m + \omega C_t + B_t(V_0) + \delta]. \qquad (2)$$

This includes the conductance of the tuning array G_t, the microstripline conductance G', the geometric capacitance of the mixing array C_m and the tuning array C_t,[12] the dc-voltage-dependent susceptance of the tuning array $B_t(V_0)$, and δ, which accounts for the susceptance of the radial stubs and other stray susceptances. Values of B_t and G_t are obtained by fitting the *I-V* curves. The susceptance change between curves *a* and *b* is $\Delta B_t = 42 \pm 9\ \text{m}\Omega^{-1}$. (The uncertainty arises from the range of embedding admittances which result in similar *I-V* curves.) The theoretical value

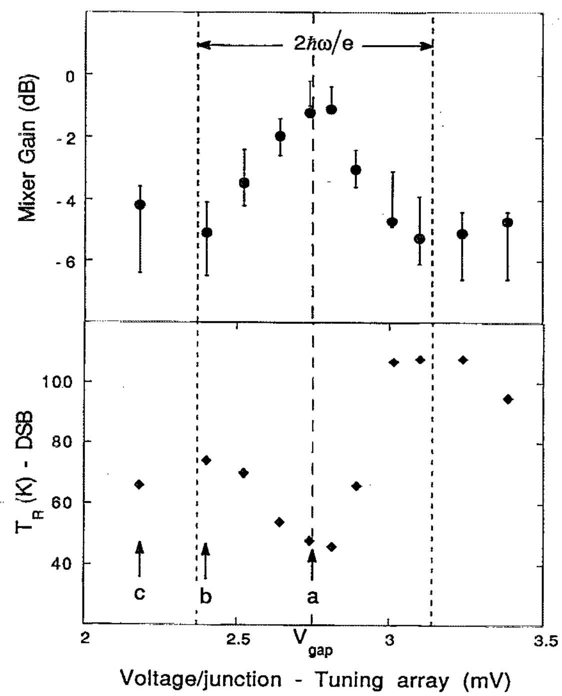

FIG. 4. Experimental mixer gain (top) and DSB receiver noise temperature (bottom) at 90 GHz as a function of the tuning array bias voltage. The bars indicate the range of theoretical mixer gains for the embedding admittances derived from the measured *I-V* curves.

TABLE I. Values of B_t and G_t inferred from fitting the rf-pumped *I-V* curves in Fig. 3, experimental and theoretical double-sideband (DSB) mixer gain and experimental receiver noise temperature for three different bias voltages of the tuning array. The bias voltages are shown in the inset of Fig. 3. Without the tuning junctions the calculated mixer gain is -3.3 dB.

Bias point V_t	a V_g	b $\approx[V_g-(\hbar\omega)/e]$	c $<[V_g-(\hbar\omega)/e]$
B_t (mΩ^{-1})	-42	0	0
G_t (mΩ^{-1})	35	35	3
g_m (dB) Calc.	-0.6 ± 0.4	-5.3 ± 1.2	-5 ± 1.4
g_m (dB) Expt.	-1.1	-5.1	-4.3
T_R (K) Expt.	47	76	65

is $\Delta B_t(\text{calc})=49$ mΩ^{-1}, from Eq. (1). This is in good agreement with the experimentally derived value.

Figure 4 shows the mixer gain g_m and receiver noise temperature T_R of this mixer at 90 GHz as a function of the dc-bias voltage of the *tuning* array. The double-sideband receiver noise depends on the mixer gain as

$$T_R=T_m+T_{\text{if}}/g_m, \tag{3}$$

where T_m is the mixer noise temperature and T_{if} is the if amplifier noise temperature, $T_{\text{if}}=8$ K. The negative susceptance of the tuning array increases as the dc bias voltage (V_t) approaches the gap voltage, where it resonates the capacitance. Above the gap voltage there is an increase in the receiver noise due to the shot noise added by the dc current in the tuning array. The conductance G_t of the tuning array does not appear to play a major role in determining the mixer gain, as seen by comparing columns b and c of Table I.

The theoretical mixer gain was calculated for various values of V_t, using the embedding admittances inferred from the pumped *I-V* curves. The trends seen in Fig. 4 for the calculated values of g_m agree rather well with those observed experimentally, including the peak at $V_t\sim V_g$. This agreement and the agreement of the experimental value of ΔB_t with the theoretical value verify the validity of the model we have employed of a separate quasiparticle tuning element. The mixer gain is significantly improved compared to the case of no tuning junctions.

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