NATIONAL RADIO ASTRONOMY OBSERVATORY

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TITLE: OPTIMUM PARAMETERS OF SIS JUNCTIONS FOR MIXER APPLICATIONS

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In fabricating a superconducting tunnel junction with given materials, only two parameters are adjustable: the junction area and the barrier thickness. The electrical behavior of the junction may also be described by two principal parameters, which I take to be the junction capacitance C and normal state resistance R_N. In this note, I try to determine the best choices of fabrication parameters for devices used in low-noise, millimeter-wavelength mixers. Two sets of materials are considered: Pb-In-Au/In_2O_3/Pb-Bi (IBM process) and Nb/NbO_x/Pb-Bi.

1. Number of Series Junctions, Based on Dynamic Range

In addition to choosing the area and thickness of each junction, the designer can choose to connect several junctions in series. In some cases this may be necessary in order to prevent a mixer from saturating at low signal levels. The series connection is called an "array" in the literature [1,2]. Arrays have the disadvantage that the interconnections add parasitic series inductance, which makes tuning more difficult and restricts bandwidth. Good performance also requires that the instantaneous current be the same in all junctions, which usually means that the array be small compared with 1/4 wavelength. Finally, all junctions should have nearly the same R_N. These requirements may be hard to meet at high frequencies.

The number of junctions required to achieve a given saturation level may be estimated as follows. Assume that the RF saturation power is some fraction ε_1 of the L.O. power absorbed by the junctions:

$$P_{sat} = \varepsilon_1 P_{LO} = \tfrac{1}{2}\varepsilon_1 V_{LO}^2 \, \mathrm{Re}\{Y_{LO}\} \tag{1}$$

where V_{LO} is the L.O. voltage amplitude across the array of junctions and Y_{LO} is the large-signal array admittance at the L.O. frequency. The L.O. voltage is assumed sinusoidal. If we suppose that all of the RF available power is absorbed by the junctions, then a conservative value of ε_1 is .01. (However, best noise performance may well be achieved with mismatched input, in which case larger values of ε_1 may be reasonable.) As a rough approximation, take $(\mathrm{Re}Y_{LO})^{-1} \approx R_A = NR_N$, the total normal resistance of the array.

Now let $P_{sat} = kT_{max}B_{max}$ where T_{max} is the maximum input noise temperature and $B_{max} = \varepsilon_2 f_L$ is the maximum bandwidth for fractional bandwidth ε_2 at L.O. frequency f_L. Then (1) becomes

$$(\varepsilon_2/\varepsilon_1)kT_{max}f_L = V_{LO}^2/2R_A \; . \tag{2}$$

V_{LO} is assumed to be adjusted to maximize the conversion gain. The latter varies approximately as $J_n(eV_{LO}/Nhf_L)$, where N is the number of junctions in series and the dc bias point is $V_{dc} = V_{gap} - (n-\tfrac{1}{2})Nhf_L/e$ (i.e., the nth "photon peak" below the gap). Taking n = 1 and noting that $J_1(x)$ is maximized at $x \approx 1.8$, we have

$$eV_{LO}/Nhf_L = 1.8$$

$$V_{LO} = 1.8\; Nhf_L/e \; . \tag{3}$$

Using (3) in (2) and solving for N^2 gives

$$N^2 = .62\left(\frac{e}{h}\right)^2 \frac{\varepsilon_2}{\varepsilon_1} \frac{kT_{max}R_A}{f_L} \; . \tag{4}$$

Thus, for a given fractional bandwidth and saturation temperature, fewer junctions are needed at higher frequencies. Also, fewer junctions are needed at lower impedances (equivalently, low impedance leads to high saturation levels).

Taking $\varepsilon_1 = .01$,

$\varepsilon_2 = .1$ (10% bandwidth),

and $T_{max} = 300K$,

(4) evaluates to

$$N^2 = (1.50 \times 10^9 \text{ Hz}/\Omega) \frac{R_A}{f_L} \tag{5}$$

TABLE I: Number of Junctions for 300°K Saturation*

f_L =	36	115	220	350	GHz
R_A = 10	1 (.58)	1 (.32)	1 (.23)	1 (.19)	
100	2	1	1 (.74)	1 (.59)	
1000 Ω	6	4	3	2 (1.86)	

*Assumes 10% bandwidth, $P_{sat} = .01\ P_{LO}$, matched input. Values in parentheses are formal results from (5); others are next higher integer.

Table I gives some values of N calculated using (5). Apparently, under the stated assumptions, arrays are necessary only in the high impedance cases and below 36 GHz.

2. Desired Normal State Resistance

The argument of the preceding section favors normal resistances of less than 100Ω. I.F. matching considerations also tend to favor low R_N, since under moderate-gain conditions the I.F. source impedance of the junction is several times R_N [3,4], and it is usually necessary to match this to a 50Ω amplifier. On the other hand, RF matching considerations favor high R_N, for easier coupling into waveguides. The latter is especially true since the reactance of the junction capacitance is preferably $R_N/10$ to $R_N/5$, as explained below; this reactance reactance must be resonated out by the imbedding network.

As a reasonable compromise, I suggest $\underline{R_N = 50\Omega}$ (or $R_A = 50\Omega$ for $N > 1$). We will then have a shunt reactance of −5 to −10Ω, for which a tuning reactance can be reasonably implemented in microstrip. If the latter succeeds in being within 20% of resonance (due to uncertainty in the actual junction capacitance as well as microstrip fabrication tolerances), the residual reactance of $\sim 50\Omega$ can be tuned out with waveguide adjustments (after a microstrip-to-waveguide transition).

3. Desired Junction Capacitance

Although the junction capacitance may at first seem to be an annoyance which should be kept small, it actually serves some useful functions in a mixer. It must be tuned out at the signal frequency, but in principal this is always possible to do because (1) there is negligible series resistance, since the leads are superconducting; and (2) the capacitance is constant (not affected by d.c. bias or L.O.). There is no "cutoff frequency" in the sense used for Schottky diodes [where $f_c = (2\pi R_s C)^{-1}$], because the series resistance R_s is essentially zero.

The useful functions served by the junction capacitance are these:

(1) It tends to produce a lossless termination (essentially a short) for the nonlinear element at high frequencies. Of particular importance are the harmonic sidebands, $f_n = f_{IF} \pm nf_L$, for $|n| > 1$.

If these are reactively terminated, then none of the signal power is converted to these undesired frequencies. Also of interest are the harmonics of the L.O.; if these are shorted, then the L.O. voltage waveform at the junction is sinusoidal. This makes theoretical analysis much simpler, although it does not necessarily imply better mixer performance.

(2) The junction capacitance also reduces noise associated with the a.c. Josephson effect. The mechanism of this noise has been roughly explained by Rudner et al. [1]. Experimentally, it is observed that the I.F. noise from a pumped junction increases drastically when the d.c. bias voltage is reduced below a certain threshold, and this extra noise can be suppressed by applying a d.c. magnetic field whose magnitude and direction are chosen to make the Josephson critical current small. It is thought that the threshold voltage is the drop-back voltage V_d plus the amplitude of the L.O. voltage. Theoretically, V_d is proportional to $C^{-\frac{1}{2}}$; thus large values of C are favored.

Whereas the Josephson effect noise can be suppressed by a magnetic field, I will consider only the first point, harmonic sideband termination, in design calculations. (However, it is necessary to verify that the magnetic field, if required, will not be unreasonably large.) This leads easily to the conclusion that $Q > 3$ is desirable, where $Q = 2\pi f R_{RF} C \approx 2\pi f R_N C$, and R_{RF} is the signal-frequency impedance (mostly resistive) of the pumped junction. In published experiments, $Q \approx 1$ resulted in relatively poor conversion [5], whereas the highest gain results were obtained with $Q \approx 10$ [3,6]. Of course, high Q results in a reduction in signal bandwidth and more critical tuning adjustments. As a good compromise, I suggest Q = 5 for most designs.

The desired capacitance is then

$$C = \frac{Q}{2\pi f R_N} = (15.9 \text{ pf GHz})/f \quad \text{for} \quad Q = 5,\ R_N = 50\Omega. \tag{6}$$

4. Junction Area and Barrier Thickness

From the desired electrical parameters, I now deduce the fabrication parameters.

We shall have to assume that the tunneling current is uniformly distributed over the junction area, although this may be difficult to achieve for very small areas or very thin barriers. Under this assumption we can scale the theoretical and experimental results to any junction area.

For a fixed area, the capacitance depends only weakly on the normal resistance; this is because the latter varies exponentially with barrier thickness. Although the specific capacitance should be calculable from the dielectric constant and thickness of the barrier, in practice the barrier may not be uniform in composition and the thickness may be hard to estimate. Fortunately, experimental data is available [7], with these results:

$$C_{Pb}^{-1} = 32.9\ \mu m^2/pf - (1.5\ \mu m^2/pf)\ \ell n(j_1/Acm^{-2}) \tag{7a}$$

$$C_{Nb}^{-1} = 10.5\ \mu m^2/pf - (0.45\ \mu m^2/pf)\ \ell n(j_1/Acm^{-2}) \tag{7b}$$

where C_{Pb} is the capacitance per unit area for Pb-In-Au/Pb-Bi junctions and C_{Nb} is for Nb/Pb-Bi; and j_1 is the maximum d.c. Josephson current density. These results are believed accurate to about 10% for $100 < j_1 < 5000\ A/cm^2$.

The maximum d.c. Josephson current density j_1 is related to the normal conductance per unit area g_N(= $1/aR_N$ for area a) by

$$j_1 = K(\frac{\Delta(T)}{e})\ (g_N - g_j) \tag{8}$$

where Δ is the average of the energy gaps of the superconductors; e is the electronic charge; g_j is a measure of the non-ideal sub-gap conductivity (conventionally defined as $i_2 a^{-1}/2\text{mV}$ where i_2 is the d.c. current at 2mV); and K is a constant which depends on the materials and the temperature. Theoretically [8,9], $g_j = 0$ and

$$K = \frac{\pi}{2} \tanh \frac{\Delta(T)}{2kT}\ , \tag{9}$$

which is close to 1.50 for both Pb-alloy and Nb/Pb junctions at 4.2K. In practice, K = 1.40 for Pb-alloy junctions [10] and K = 1.18 for Nb/Pb junctions [8] are measured. Also, g_j is usually 5% to 10% of g_N. Using these experimental values (with $g_j = 0.1\ g_N$), (8) becomes

$$j_1 = \begin{cases} (1.73\ \text{mV})\ g_N & \text{for Pb-alloy} \quad (10a) \\ (1.51\ \text{mV})\ g_N & \text{for Nb/Pb}\ . \quad (10b) \end{cases}$$

Using (7) and (10), we can now insert the desired electrical parameters, namely $R_N = 50\Omega$ and Q = 5, and deduce the junction area a and Josephson current density j_1. The results are given in Table II for various frequencies. As expected, the niobium junctions require smaller areas and higher current densities to be electrically similar to the lead junctions. In fact, the ratio is close to that of the dielectric constants of Nb_2O_3 and In_2O_3 (8 and 29, respectively).

It should be noted that current densities greater than a few thousand A/cm^2 are difficult to achieve in the Pb-alloy materials using the IBM processing technology. However, $j_c > 10^5$ A/cm^2 is reported as routinely achieved with Nb by using ion beam oxidation [11]. The latter technology is being implemented at the University of Virginia.

TABLE II: Fabrication Parameters for $R_N = 50\Omega$ and $Q = 5$

f_{RF} =		36	115	220	350	GHz
Pb-alloy						
	a	10.7	3.1	1.54	.93	μm^2
	j_1	320	1100	2250	3300	A/cm^2
Nb/Pb						
	a	3.3	.95	.48	.29	μm^2
	j_1	915	3200	6300	10400	A/cm^2

The smallest area required in Table II, 0.29 μm^2, is achievable with edge junctions and photolithography. More advanced techniques, such as electron beam lithography and shadow-masking, do not appear to be necessary for mixer applications. In fact, efforts to make much smaller area junctions seem counter-productive. On the other hand, if one is restricted to planar (non-edge) geometry, $a < 6$ μm^2 is hard to achieve with photolithography, but electron beam lithography would probably be adequate for all cases in Table II.

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