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FEASIBILITY OF MILLIMETER-WAVE SIS DIRECT DETECTORS

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I. Summary and Conclusions

The sensitivity of NRAO present and projected millimeter-wave receivers are compared with expected sensitivities of an SIS detector receiver. The results are summarized in the NEP_d column of Table I. In general the expected SIS sensitivity is approximately a factor of two better than our present best continuum receiver at 90, 230, and 345 GHz, but are a factor of two worse than new mixer receivers expected at 90 and 230 GHz. However, the fabrication of a focal-plane array (say, 100) of SIS detectors is more feasible than for mixers and for this reason the development should probably be pursued. Compared to doped germanium bolometers SIS detectors operate at 4.2K rather than 0.3K and are more narrowband (~ 5%) which alleviates confusion with spectral lines. The achievement of useful sensitivity with SIS detectors requires junctions with low sub-gap current (< 1 μA for R_N = 100) and high resistance (30 R_N) in the sub-gap region. Symmetrical square-wave bias of the junction, use of a cryogenic transformer, balanced-bridge techniques, and very careful attention to layout, grounds, and shielding may all be necessary to observe the nanovolt output signal of the detector.

II. Sensitivity Comparison

It is instructive to compute the detector noise-equivalent-power (NEP) which would give continuum sensitivity equal to the NRAO present Schottky and SIS mixer continuum sensitivity, projected (to 1989) mixer sensitivity, and to the present NRAO doped germanium bolometer. Using subscript d for detectors, m for mixers and equating receiver rms antenna temperature sensitivity, ΔT (which is proportional to minimum detectable flux, $S_{min} = 2k\Delta T/\eta A$) gives:

$$\Delta T_d = \frac{2NEP_d}{\sqrt{2}\ kB_d} = \Delta T_m = \frac{2NEP_m}{\sqrt{2}\ kB_m} = \frac{2T_m}{\sqrt{B_m}}$$

where T_m is the double-sideband mixer receiver noise temperature. The $\sqrt{2}$ factor in the denominator arises because ΔT is for one second integration while NEP is for 1 Hz video bandwidth. The 2 factor in the numerator is due to switching. Using values of T_m, B_m, and B_d for present and projected receivers, the values of NEP_d to give equal sensitivity and also the values of $\Delta T_m = \Delta T_d$ and S_{min} for stated antenna efficiency, η, are all given in Table I.

Also shown in Table I is the expected NEP of an SIS detector with assumptions as stated in IV.

TABLE I. NEP_d and Bandwidth, B_d, of an SIS Detector to Give Sensitivity Equal to Various NRAO Receivers

	f GHz	T_m °K	B_m GHz	ΔT_m, ΔT_d °K	B_d GHz	NEP_d $\times 10^{-15}$	η	S_{min} Jy
Present Mixer	90	140	0.6	.011	5	0.56	.50	0.5
Future Mixer	90	70	2.0	.003	5	0.15	.50	0.14
Future SIS Detector	90				5	0.50	.50	0.35
Present Mixer	250	250	0.6	.019	10	2.0	.30	1.5
Future Mixer	250	150	2.0	.006	10	0.6	.30	0.5
Present Bolometer	250			.018	10*	2.5	.30	1.4
Future SIS Detector	250				10	1.3	.30	0.7
Present Mixer	345	800	0.6	.065	15	9.5	.12	13
Future Mixer	345	400	2.0	.018	15	2.6	.12	3.6
Present Bolometer	345			.018	15*	3.7	.12	3.6
Future SIS Detector	345				15	1.8	.12	1.8

*This is the SIS detector bandwidth; bolometer bandwidth is 80 GHz, and NEP is 20×10^{-15} $watts \cdot sec^{1/2}$.

III. SIS Detector Responsivity

An excellent review of the theory and past experimental work on SIS detectors is given by Tucker and Feldman [1].

An upper bound to the current responsivity of an SIS detector is e/hf; this is one electron per photon and is equal to 2685 amps/watt = 2.7 $pA/10^{-15}$ watt at 90 GHz. Experimental responsivities close to this bound have been observed. This value is high but should be considered in the following context:

1) One electron per photon is not the total story; the energy of the electrons or how much voltage potential the electrons can be pushed through also matters. A typical video output resistance of 3,000 ohms gives a voltage responsivity of 8.1 $nV/10^{-15}$ watt at 90 GHz.

2) The current responsivity of an ideal Schottky diode is e/2kT = 20 amps/watt at T = 300K. This is a factor of 134 times less than that of an SIS detector at 90 GHz - not an enormous factor in view of the several orders of magnitude difference in sensitivity of diode detectors and diode mixers. (Cooling a Schottky diode detector to 20K will increase its responsivity by a factor of ~ 5.) The lower noise of the SIS detector provides greater sensitivity beyond the responsivity advantage provided that a sufficiently low-noise video amplifier can be realized.

IV. SIS Detector NEP

The noise equivalent power, NEP, of an SIS detector is given in Table I under assumptions which are given in this section and use of the following equations:

$$NEP = \frac{\sqrt{e_j^2 + e_a^2}}{R_V \ \varepsilon \ e/hf} = \frac{\sqrt{i_j^2 + i_a^2}}{\varepsilon \ e/hf}$$

where e_j is the rms voltage noise of the SIS junction, e_a and i_a are the rms voltage or current noise of the video amplifier[1] driven by detector video resistance, R_V, ε is the efficiency of coupling to the junction, and e/hf is the ideal current responsivity.

The junction quality enters into the NEP equation through R_V and e_j or i_j. For excellent quality niobium junctions, which are expected in the next few years, values of R_n = 100, R_V = 3000, and leakage current, I = 1 μA appear

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[1]Either e_a or i_a (not both) represent the noise in the video amplifier driven from a specified source impedance R_V; they are not the two noise sources which together represent the video amplifier noise for any source impedance.

appropriate. This gives a shot noise current $i_j = \quad 2eI = .56$ pA or $e_j = R_V\, i_j = 1.7$ nV or a junction video noise temperature, $T_J = 17K$. An amplifier with noise current, voltage, and temperature of 0.28 pA, 0.85 nV, and 4K is feasible with a silicon junction FET (2SK147 or 2N6550) at 300K and contributes only 25% to the NEP. It is assumed that AC bias of the junction at a frequency of 1 kHz will be utilized so that 1/f noise is not a problem and also transformers can be utilized for impedance transformation and ground isolation.

The values of NEP_d for an SIS detector at 90, 250, and 345 GHz are given in Table I for the junction and amplifier parameters given above and a coupling efficiency, $\varepsilon = 0.5$. It should be noted that in the case of a junction with four times higher leakage current and 1/4 the video resistance (4 μA and 750 ohms) the NEP is only doubled with an amplifier having noise current, voltage, and current of .56 pA, .42 nV, and 4K. This is the same amplifier noise temperature as the previous case except it is now at a lower source resistance as could be obtained with a low-loss transformer.

V. Circuit Considerations

The use of AC square-wave bias of the junctions has been suggested by Richards [2] as a method of avoiding 1/f noise in the video amplifier. A problem which then arises is the separation of the AC bias signal from the detector output signal which may be 10^6 times weaker (3 mV vs. 3 nV); this is a very large dynamic range for a synchronous detector. (In the DC bias case, the coupling capacitor separates bias from signal.) One remedy would be to cancel the bias signal by subtraction after a preamp as shown in Figure 1. A second possibility would be a bridge circuit with a second SIS junction to balance the bridge as shown in Figure 2. The second junction could either be "in the dark" or could be looking at an adjacent patch of sky. The NEP of the detector is reduced by a factor of two by this arrangement since the second junction adds noise. If the second junction was replaced by a low source-resistance bias voltage source at low temperature, the added noise would be reduced but the cancellation would then be a function of bias voltage magnitude.

The circuits of Figures 1 and 2 illustrate some other principles which may be used to advantage:

1) The amplifier may be designed to have nearly zero input impedance (by feedback) without changing its noise performance. This may be necessary to prevent saturation for high input temperatures. For a junction with $R_N = 100$, $R_V = 3000$, $B = 5$ GHz, and $\eta = 0.5$ at 90 GHz, saturation of 1% occurs at 1200K for an amplifier input impedance, R_L, of 3,000 ohms and 72,000K for $R_L = 50$ ohms. (See Feldman [4], equation 11, and paragraph regarding communication from A. D. Smith.)

2) The bias generator must be mismatched to the junction to avoid signal power loss. This may be accomplished by a bias generator source-impedance either much lower or much higher than the junction video resistance. Low impedance (i.e., a voltage source) is probably more stable.

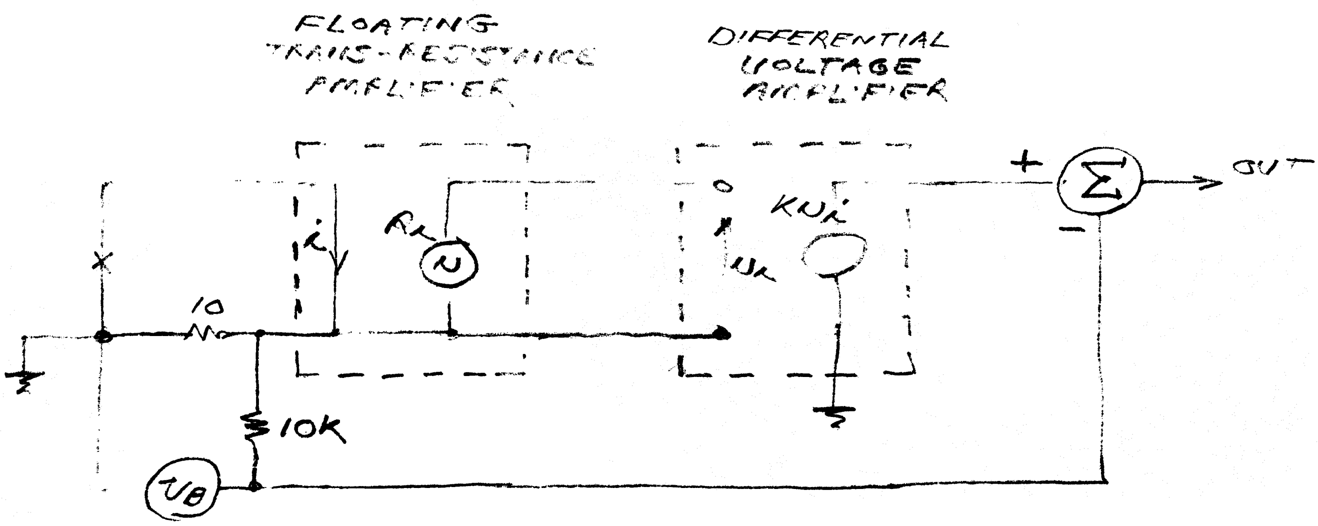

Fig. 1. Detector bias and amplifier configuration with low-input impedance amplifier and bias cancellation circuit. The amplifier can have low-input impedance and yet have very low noise when driven by a high video output impedance detector.

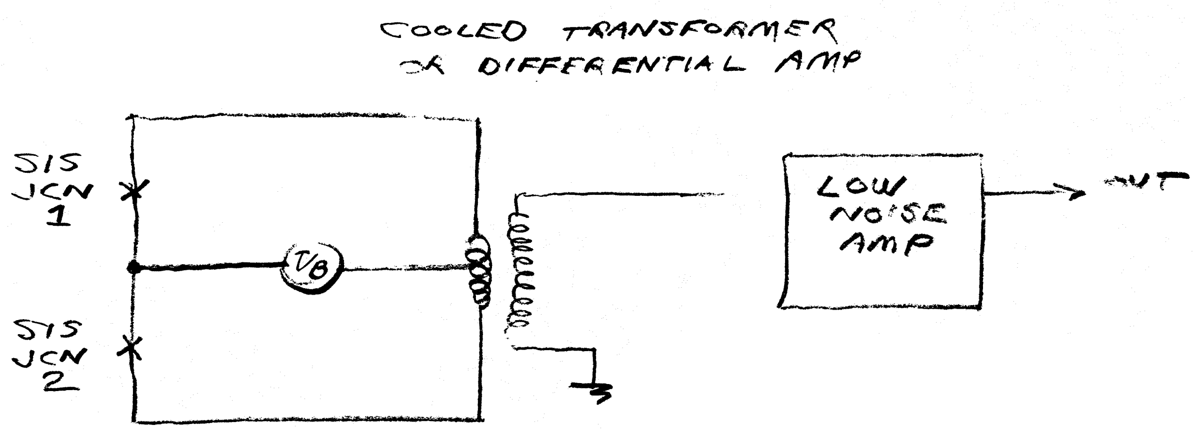

Fig. 2. Bridge circuit to cancel AC bias. The second junction could either have no RF excitation or be in another waveguide (or polarization) looking at adjacent sky.

3) Cooled transformers can be used to advantage. An excellent article by Hannah [3] on this subject (and also a low-noise amplifier) shows that Metglas 2826 (a metallic glass manufactured by Allied Chemical) or Supermalloy (Arnold Engineering #2T-S1) core material have high permeability at 4.2K. It may be an advantage to float both sides of the junction from chassis ground; a low-pass filter-choke on both ends of the junction would be used.

4) Finally, the avoidance of ground-noise generators, thermocouple voltage noise, RFI, and extraneous magnetic fields is of the utmost importance. Thermocouple voltages are very large (10^5 to 10^6 nV per K) compared to the voltages of interest. RFI from VLF transmitters (16 kHz and lower) may leak through most RF feed-through filters. It may be necessary to house a transformer and amplifier in a superconducting box located very close to the detector mount.

REFERENCES

[1] J. R. Tucker and M. J. Feldman, "Quantum Detection at Millimeter Wavelengths," Rev. Mod. Phys., vol. 57, no. 4, pp. 1067-1070, October 1985.

[2] P. Richards, private communication, 1985.

[3] E. C. Hannah, "Low Temperature Magnetic Cores and a Preamp for Low Impedance Cryogenic Sources," Rev. Sci. Inst., vol. 52, no. 6, pp. 1087-1091, June 1981.

[4] M. J. Feldman, "Saturation of Superconductor Quasi-Particle Direct Radiation Detector," to be published, J. of Applied Physics, accepted June 17, 1986.

APPENDIX I. Comments by Larry R. D'Addario

1. The so-called "quantum limit" of one electron per photon may not be a real limit. It is certainly not fundamental like the quantum noise limit (which is based on the Uncertainty Principle); that is, there is nothing fundamental to prevent a photon from inducing the tunneling of two or more electrons. I suspect, however, that in most practical circumstances the probability of single-electron tunneling is much higher than multiple-electron tunneling. Someone more versed in the quantum physics than I should look into whether there are conditions under which the 2-,3-,...electron tunneling probabilities are enhanced. I have mentioned this to Mark Feldman, but have not yet heard his opinion of it.

2. Your comment on page 2 that the current responsivity is not the whole story is certainly right, but I would express it differently. A better figure of merit would be the conversion efficiency (as in mixers) or power gain, given by $r_i^2 * R_v/4$, where r_i is the current responsivity and R_v is the video resistance; this is just the ratio of available power at the output (video or d.c.) to the RF power absorbed.

3. Your coupling factor ε seems out of place. If it is meant to represent ohmic losses in the mount, then there also needs to be a noise term proportional to the temperature of these losses. If it represents mismatch losses, the exact consequences are more complicated, depending on how much noise is radiated from the detector. If it represents geometric losses in the optics preceding the detector, then it seems more logical to include these in the aperture efficiency of the antenna. Your later choice of 0.5 (-3 dB) for this factor seems high for either ohmic or mismatch losses in a reasonably designed mount.

4. Your memo concerns the design of only the video side of the detector, and neglects the RF side. What type of mount do you intend to use for your tests? Your estimated bandwidths seem rather pessimistic (only 3% to 5%). The choice of junction normal resistance (you suggest 100 ohms) is based mainly on RF matching considerations. The input impedance will be considerably different than for a mixer using the same junction, in view of the absence of pumping, and also considering that for a detector we want a good match whereas for a mixer (with gain) we generally want a high source resistance. Your later comment that it may be advantageous to float both sides of the junction from ground could make broadband RF matching very difficult. As I have been saying for years, the configuration in which the device is placed all the way across a waveguide (even a reduced height waveguide) leads to large parasitic reactances and narrow bandwidth. Besides, I don't see what the advantage of d.c. floating would be; both of your circuits allow one side of (each) junction to be grounded.

5. I have been trying to check your proposed numbers for junction parameters. By "leakage current" you apparently mean the quiescent current at the operating bias. This, and also R_v, depends critically on where the bias point is. Normally, the highest responsivity occurs at about V_g - hf/2e, where V_g is the gap voltage. Note the frequency dependence. For most practical junctions, this puts you a bit up on the knee at 90 GHz, but well below it at 345 GHz. Thus, your "leakage current" will vary drastically over this frequency range. Furthermore, this

current is a strong function of physical temperature. Cooling below 3K should give big improvements, even for Nb junctions. (In mixers, sub-gap current is much less important because the current is dominated by the pump-included tunneling, and because the gain reduces the effect of the shot noise when referred to the input.)

The bottom line is that I expect the achieving of 1 uA bias current at 90 GHz to require a very high quality junction indeed, and possibly cooling to 2-3K; whereas at 345 GHz, much less than 1 uA should be possible, especially at low temperatures. All this assumes R_n = 100 ohms.

As to your estimate of R_v = 3000 ohms, this seems to be just 3mV/1uA, and is consistent with the above ONLY IF the junction sharpness is good enough so that you are biased well below the knee. Again, this may be a problem at 90 GHz unless the junction quality is especially good and/or you cool it enough.

6. Niobium junctions keep getting better, but I believe that the sharpest I-V curves are still obtained with lead alloys, especially those tailored to this objective; see Gundlach _et al._, _Appl. Phys. Lett._, vol.41, p.294, 1984. Taking data from this paper, scaled to R_n = 100 ohms, they get a current of about 3 uA when biased for 90 GHz and 0.9 uA for 345 GHz, both at 4.2K; but around 0.2 uA for 90 GHz and < 0.1 uA for 345 GHz at 2K.

7. The mechanism for sub-gap current greater than that of an ideal junction (which is non-zero but very small - see e.g., Feldman, eqn. 15) is not known. It may be best modeled as a resistor in parallel with an ideal junction, in which case the shot noise treatment would be wrong. Instead, the video noise temperature would be nearly equal to the physical temperature, rather than the 17K you calculated from shot noise theory.

8. Single-junction devices are best for detectors, whereas we strongly prefer series arrays for mixers. Do you intend to make special single-junction devices for these tests, or will you use mixer devices? Perhaps Tony has included some singles on his Hypres wafers; I don't know. But the design is more difficult, for several reasons. Topology favors an even number of junctions, for most processes. The total capacitance is larger for a given junction size, leading to difficulty in implementing on-chip tuning. Pushing to smaller areas to reduce the capacitance tends to degrade junction quality.

9. The balanced bridge configuration suggested in Figure 2 is mainly useful if the junction noise turns out to be much less than you predict, for reasons suggested above. If used, however, I think the reference junction must be kept in the dark; the sky noise from an "adjacent patch" would dominate (for earth-bound telescopes). However, one could not rely on this to achieve accurate bias cancellation. It is difficult to get junctions to be equal to 1% in critical current, even on the same chip, and especially if the lithography is near its limit, as would be necessary at the higher frequencies.

10. Of course, the bias cancellation need not be complete; only enough to keep within the dynamic range of later amplifier stages is needed. The critical thing is the STABILITY of the difference between the signal and reference in the face of fluctuations induced by junction temperature changes, power line voltage changes, amplifier gain changes (if cancellation occurs after some gain), etc.