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A SUBHARMONICALLY PUMPED MIXER FOR 230 GHz

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FOR 230 GHZ

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ABSTRACT

A subharmonically pumped mixer has been built for the 200 to 260 GHz band. The design utilizes GaAs Schottky barrier diodes in a back-to-back configuration to obtain an hyperbolic sine law i-v characteristic. A scalar feed is integrated into the mixer block to minimize waveguide losses. The mixer has been tested with L.O. frequencies from 99 to 115 GHz and I.F. of 4.5 to 5.0 GHz. The measured conversion loss and noise temperature in terms of single sideband operation are 13.2 to 14.5 dB and 4700 to 5800°K respectively for L.O. frequencies between 100 and 115 GHz (signal frequencies from 195 to 235 GHz) for room temperature operation.

1. Introduction

A need exists for low noise mixers above 150 GHz for use in radio astronomy[1,2]. The problems of mixer design at these frequencies may be enumerated as:

1) extremely small size limits realizability of designs and magnifies parasitics due to physical size of diode chips, contacting arrangement, etc.
2) high losses per unit length for transmission media
3) approaching limitations of semiconductor diode performance
4) problems related to physical dexterity required in handling and assembling small parts that compose the mixer.

Design, construction and performance of a subharmonically pumped mixer for 230 GHz are described. This mixer uses back-to-back mounted Schottky barrier diodes to obtain the antisymmetric hyperbolic sine law i-v characteristic required for high efficiency subharmonically pumped mixing. A scalar feed for signal frequency input is integrated into the mixer block in order to minimize input waveguide losses.

Some advantages of the subharmonically pumped mixer configuration chosen compared to conventional fundamental mixers in this frequency range are as follows:

1) the local oscillator frequency is about half that required for a fundamental mixer, and, therefore, substantial savings in klystron expense and improvements in operational life are possible (commercial klystrons are not available above 200

GHz, extended interaction oscillators, EIO, or backward wave oscillators, BWO, for this frequency range cost about \$25K to \$35K whereas 115 GHz klystrons cost about \$5K)

2) the contribution of L.O. noise is greatly reduced since the signal frequency is not near the L.O. frequency[3]

3) the signal and L.O. frequencies are easily separated as a result of the large frequency difference.

The electronic design, fabrication and assembly techniques, and observed performance characteristics are described in succeeding portions of this report.

2. Design

Advantages of using antiparallel diodes for harmonic mixing were first elucidated by Cohn et al[4]. This approach to millimeter wave mixer design has also been strongly popularized by the work of Schneider et al[5,6,7]. Briefly, the combination of two GaAs Schottky barrier mixer diodes in a back-to-back configuration yields a current voltage characteristic that is described by a hyperbolic sine law.

$$I = I_s \left(e^{\frac{qV}{\eta KT}} - 1 \right) \text{, single diode} \tag{1}$$

$$I = 2I_s \sinh \left(\frac{qV}{\eta KT} \right) \text{, antiparallel diode pair} \tag{2}$$

The differential conductance of the antiparallel diode pair is correspondingly[4]

$$q = \frac{di}{dV} = \frac{2qI_s}{\eta KT} \cosh \left(\frac{qV}{\eta KT} \right) \tag{3}$$

Examination of these expressions, (2) and (3), reveals that the current is an antisymmetric function and the conductance is a symmetric function of the voltage impressed across the diode pair.

As a direct result of the above symmetry properties, the Fourier expansion for the conductance contains only even harmonics of the local oscillator (pumping) frequency. When a small signal voltage is included in the impressed voltage waveform and the resultant output spectrum is analyzed, it is seen that there is no contribution at the intermediate frequency, I.F., arising from fundamental mixing, i.e. from the direct interaction of the local oscillator with the signal. Rather, production of the I.F. signal proceeds via interaction of the signal frequency with the conductance component at twice the local oscillator frequency. Therefore, for the ideal subharmonically pumped mixer using a hyperbolic sine law pair of antiparallel diodes the following frequency relationships obtain.

$$f_s = 2\, f_{LO} + f_{IF} \tag{4}$$

$$f_i = 2\, f_{LO} - f_{IF} \tag{5}$$

A more detailed theoretical analysis of subharmonic mixers is given in a separate report[15].

Thus, the task of designing a subharmonically pumped mixer separates quite naturally into the design of frequency separation networks and impedance matching elements at the widely separated signal, local oscillator, and intermediate frequencies. To minimize propagation losses, the signal and L.O. transmission media were chosen to be waveguides. The I.F. output, at 5 GHz for the present system, is coaxial for convenience of size and interconnection to existing I.F. circuitry. The millimeter wave transmission medium chosen in which to fabricate the frequency selective filters to interconnect the signal, local oscillator, and I.F. frequencies is suspended stripline of CrAu metallization on fused quartz[7,8,9].

A cross-sectional view of the suspended stripline is shown in Figure 1. This section, with a 10 mil. wide conductor, has a characteristic impedance Z_0 of about 55 ohms. It can be seen that the dimensions are very small, making accurate machine work difficult. Also, the 3 mil. thick fused quartz circuit is quite delicate and is usually curved somewhat from the tension of the metallization. The cutoff frequency of the cross-section shown in Figure 1 for the next highest order propagating mode is given by[8]

$$f_c = \frac{c}{2a} \left(1 - \frac{h\ (\varepsilon_r - 1)}{b\ \varepsilon_r} \right)^{1/2} \tag{6}$$

where ε_r is the relative dielectric constant of the substrate and a, b and h are as shown in Figure 1. For the dimensions chosen for the present mixer, the cutoff frequency f_c = 297 GHz.

The signal filter is designed to reject a signal in the band 200 to 260 GHz but to pass the local oscillator frequency between 97.5 and 127.5 GHz. The I.F. frequency at 5 GHz is also passed. This filter is realized as a cascade of alternate capacitive and inductive sections of the suspended stripline transmission line as shown in Figure 2. The design is a 230 GHz scaled version of a design published by McMaster et al[5]. A 25:1 scale model was tested in C and X bands where the performance was verified. Electrically, this filter exhibits $/S_{11}/ < 0.2$ (VSWR < 1.5:1, part of this is due to the coax-to-stripline transitions used to test the scale model) for frequencies below 130 GHz. For frequencies between 187 and 260 GHz the filter exhibits $|S_{21}| < 0.02$ (34 dB rejection) with no observed spurious transmission. The filter input impedance is series resonant at about 240 GHz (capacitive below, inductive above). The location of this filter is in alignment with the top wall of the signal waveguide to provide the desired series impedance.

The local oscillator rejection filter (choke) is very similar in design to the L.O. filter for the NRAO 80-120 GHz mixer[10]. It is realized as a cascade of six sections of alternating low and high impedance transmission lines. The dimensions of this filter are also shown in Figure 2. The low impedance end of the L.O. filter is positioned just at the top (I.F.) edge of the L.O. waveguide. A scaled version of this filter in suspended stripline has not been measured to date.

The two filters discussed above are connected by a section of 10 mil. wide ($\sim 55\Omega$) stripline. This section comprises the waveguide-to-stripline transition for the local oscillator. The transition is inductive in shunt with the L.O. waveguide characteristic impedance and the L.O. backshort's variable susceptance. The tunability of the backshort permits matching the local oscillator pump power into the diode pair with low loss. The circuit across the signal waveguide is completed by the antiparallel diode pair which are mounted in the ($\sim 125\Omega$) reduced-height signal waveguide between the bottom wall and the stripline circuit.

A cutaway drawing of the mixer block is shown in Figure 3 which clarifies the relative locations of the signal, local oscillator, and I.F. ports. The I.F. port is taken out through a bellows contact to the end of the stripline to an OSM connector to connect with the I.F. amplifier. The signal is fed into the signal waveguide through a scalar feed horn that is integral with the mixer block. The scalar feed is fabricated by electroforming as shown in Figure 4 and incorporates a circular to rectangular waveguide transition and step impedance transformer as shown in Figure 5. Since the signal waveguide is beyond cutoff for the local oscillator, it provides about 45 dB isolation which prevents leakage of the L.O. out the feed horn.

The GaAs mixer diodes have either a 1.5 or a 2 μm diameter Schottky barrier contact. Junction capacitance is about 4 to 8 fF and series resistance is in the range of 6 to 13 ohms. The diode ideality factor η is about 1.20 to 1.25 for these devices. Whisker inductance is in the range of 100 to 140 pH. The equivalent circuit used to model the antiparallel diode pair is shown in Figure 6. The whisker inductance dominates the impedance transformation of the diode-stripline-waveguide junction. A more complete theoretical analysis of the impedance matching and performance of this mixer is given in a separate paper[15].

3. Construction and Assembly of the Mixer

The mixer block was machined out of brass. Both the scalar feed horn and the signal backshort waveguide were fabricated by electroforming onto aluminum mandrels as described earlier. The electroformed parts were machined externally and press fit into the brass block. The L.O. waveguide was machined with one half in each side of the block so that the split occurs in the center of the top and bottom broad wall. The channel for the suspended stripline microcircuit and for the dual miniature whisker pins were machined as a final operation. After completion of machine work the block halves were cleaned and a thin gold electroplating was applied.

The assembly of the stripline circuit and antiparallel diode pair into the mixer block is accomplished through a sequence of operations. The 3 mil. thick by half inch square fused quartz blanks are vacuum coated with CrAu on one surface. The circuit patterns are etched using a photoresist technique. The individual circuits may be separated either by scribing and breaking or by diamond sawing. The mask layout for the diamond sawn part permits the position of the signal choke to be varied by several mils. (by adjusting the saw cut position) for fine tuning purposes.

The stripline circuits are prepared for assembly by bonding small pieces of 1 mil. thick gold foil to the end sections. At the I.F. end, the gold is folded over the end to provide good electrical contact to the bellows. At the signal end, the gold foil completes the first low impedance section of the signal choke and a diode chip and whisker are mounted side by side beneath the gold foil to mate with the chip and whisker that are mounted on the miniature whisker pins.

The technique for contacting the two antiparallel chips simultaneously is illustrated in Figure 7. The stripline circuit is cemented in place in its channel using Eastman 910 contact cement. One miniature whisker pin is advanced until the whisker on the end of the stripline circuit contacts a Schottky dot on the surface of the GaAs chip mounted on the whisker pin. Electrical contact is monitored using a curve tracer. Once the first diode is contacted and deflected, the other miniature whisker pin is advanced until the chip on the stripline is contacted in like manner. The whisker pins are held in position using a drop of contact cement along the channel edge side. Whiskers are fabricated of 12.5 μm diameter phosphor bronze wire which is electropointed in NaOH solution and then gold plated. Whisker bending is accomplished by hand using simple jigs.

Two tunable backshorts are provided for this mixer. The L.O. backshort is used to optimize transfer of L.O. power to the diodes. The signal backshort is tuned to match the signal into the diodes at a particular frequency. The waveguide dimensions are .080 X .020 and .043 X .006 inches respectively for the L.O. and signal waveguides at the backshort. A standard split finger backshort fabricated from BeCu alloy and gold plated is used for the L.O. backshort. Due to the extremely reduced cross section of the signal waveguide, the backshort is fabricated from two pieces of .002

inch BeCu sheet. The individual halves of the signal backshort are fabricated by a photoetch process and are subsequently polished to remove sharp edges, gold plated, and bent to provide intimate contact with the waveguide walls. One half of the 230 GHz backshort is shown in Figure 8.

Once the diodes have been whiskered, the halves of the mixer block are assembled. Steel dowel pins aid in precise alignment of the block halves. The I.F. connector is attached, making contact to the stripline through an electroformed bellows. Finally, the backshorts for the signal and local oscillator are inserted and the micrometer backshort drives attached and the mixer is complete and ready for testing.

4. Measurement of Mixer Performance

The mixer performance was tested by the hot and cold load technique using a noise test set described previously[11]. The local oscillator was provided by klystrons in the 95 to 120 GHz range. For tuning purposes, a swept source in the signal frequency range was assembled using a BWO sweeper from 95 to 115 GHz in combination with a varactor frequency doubler. The detected I.F. output was displayed on an oscilloscope in synchronism with the sweep frequency. Also displayed was the swept detected current from the doubler diode. Frequencies were identified by the dip induced in the detected current by a frequency meter in series with the BWO output.

A range of sideband conversion ratios were observed. At most frequencies it was possible to tune the mixer for double side band operation. At some frequencies, it was possible to tune for substantial difference in sideband conversion ratios by adjustment of signal backshort position. Once the mixer was tuned for a given frequency and mode of operation, the performance was measured by application of hot (room temperature) and cold (77^{o}K) loads at the input to the scalar feed horn. The change in I.F. output temperature recorded permits determination of mixer conversion loss and noise temperature.

The noise test set is calibrated using room temperature and 77^{o}K loads. The I.F. gain and offset are adjusted to obtain readings of 80.5^{o}K and 297.5^{o}K for the cold and hot calibration loads respectively.

The slightly high temperature setting for the cold load ($80.5^{o}K$ versus $77^{o}K$) is to compensate for a small amount of loss in the cold load which is warmer than $77^{o}K$. This calibration puts the output reading of the test set in a one-to-one correspondence with the input (I.F.) temperature or equivalent power. Next the I.F. cable is disconnected from the cold load and temperature readings of the open circuit are taken with the noise diode off and on to determine the noise contribution from reflected I.F. noise and the effective temperature of the noise diode at the cable end.

T_{nbx} = temperature reading of open circuit cable with noise diode off

T_r = temperature reading of open circuit cable with noise diode on

$\Delta T_r = T_{or} - T_{nbx}$

Mixer noise data are taken, after calibration, by recording three temperature readings plus backshort positions, bias current and voltage, etc. The three temperature readings are as follows:

T_{co} = I.F. temperature with a cold absorber ($T_c = 77^{o}K$) placed in front of feed horn

T_{ho} = I.F. temperature with a room temperature ($T_h = 297^{o}K$) absorber placed in front of feed horn

T_{ro} = I.F. temperature with noise diode on

The uncorrected mixer conversion loss is calculated from

$$L = \frac{2\ \Delta T}{T_{ho} - T_{co}}$$

where $\Delta T = T_h - T_c = 297 - 77^oK = 220^oK$ and the factor of two is included because of the implicit assumption of DSB operation. In instances where operation is verified not to be DSD, a more complex calculation is required. The reflection coefficient of the I.F. mismatch is determined from

$$R = (T_{ro} - T_{ho})/\Delta T_r \tag{8}$$

The corrected conversion loss is then

$$L_c = (1 - R)\ L \tag{9}$$

The corrected mixer noise temperature is

$$T_{mxr} = (T_{co} - R\ T_{nbx})\ L - 2\ T_c \tag{10}$$

and the diode noise temperature is

$$T_d = T_{mxr}/(L_c - 2)$$

Test set errors are related primarily to gain and offset stability and to the inaccuracy of obtaining a relatively small ΔT from two larger numbers each of which tend to fluctuate on the order of $\pm 1^oK$ rms.

Measured noise temperature and conversion loss data for room temperature mixer operation are displayed in Figure 9. These numbers have been corrected for I.F. mismatch. The mode of operation is DSB for these data but the numbers are given as SSB values. Between 200 and 230 GHz the mixer loss is between 13 and 14.5 dB and the noise temperature is between 4700 and 5800^oK. The conversion loss and noise temperature are quite flat across this frequency band indicating the capability for broad band operation of this mixer. The diode noise temperature T_D corresponding to the data in Figure 9 is in the range of 200 to $260^{\circ}K$ which is about 25% below the typical value for a single diode fundamental mixer. The I.F. match was very good, typically less than 0.5 dB mismatch loss, without an I.F. matching transformer. This is in agreement with the developed theory and is a unique characteristic of subharmonic (antiparallel diode) mixers.

5. Discussion of Results and Conclusion

In order to put the measured performance of the present mixer in some perspective it is worthwhile to compare the data presented in Figure 9 with results quoted in the literature for the same frequency range. A fundamental mixer at 230 GHz reported in 1974 by Schneider and Wrixon[12] gave a DSB conversion loss of 12.6 dB ($T_{DSB} \cong 5000^{o}K$). A second harmonic mixer at 230 GHz reported in 1976 by Goldsmith and Plambeck[13] gave a DSB noise temperature of $6000^{o}K$ (NF = 13 dB) with L_c (DSB) $\approx$ 10 dB. And, in 1977 was reported a subharmonically pumped mixer at 230 GHz by Schneider giving SSB performance of L_c = 9 dB and $T_{mxr} = 2200^{o}K$ operating at room temperature[14]. Thus the measured performance of the first mixer of this type to be assembled and tested at NRAO is quite encouraging.

This mixer was originally designed with the intention of mounting the GaAs Schottky diodes on the conductive surface of the stripline circuit at a location just at the end of the signal choke a distance of 15.5 mils. from the top wall of the signal waveguide. After fabrication, the signal waveguide height was measured to be 23 mils. rather than 21.5 mils. placing this position a distance of 14 mils. from the waveguide edge. Also, tests of the scaled stripline circuit indicated that the diodes could not be mounted on the stripline as originally intended since the physical size of the chips caused several undesirable resonances for this configuration. It was decided to move the antiparallel chips to the location described earlier with one chip and whisker in the top wall of the signal waveguide and the other chip and whisker in the bottom wall of the signal waveguide (on the miniature whisker pins).

Presently, the existing mixer is being prepared for cryogenic tests. The scalar feed design is a 7° taper half angle to match the 36-ft. antenna and existing lens. It appears that a mixer noise temperature of less than 2500°K will be quite useful for initial observations at 230 GHz, therefore, a cooled version of the present subharmonically pumped mixer is a good candidate for a new 200 to 230 GHz receiver for the 36-ft. antenna if the usual factor of 2 to 3 reduction in noise is obtained through cooling.

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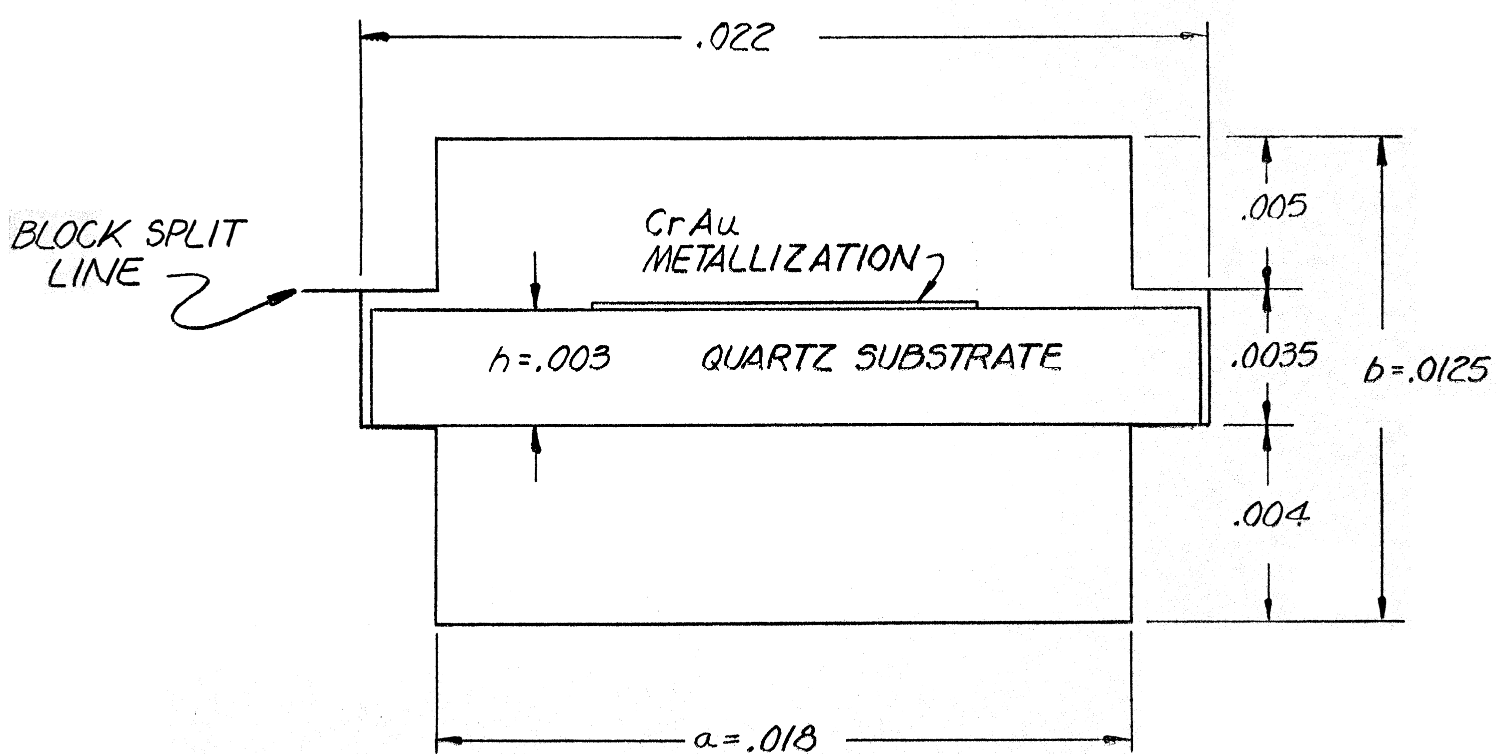
LIST OF FIGURES

FIGURE 1. SUSPENDED STRIPLINE CROSS-SECTION



DRD 1/28/77
SHPM choke

FIGURE 2: CHOKE STRUCTURE

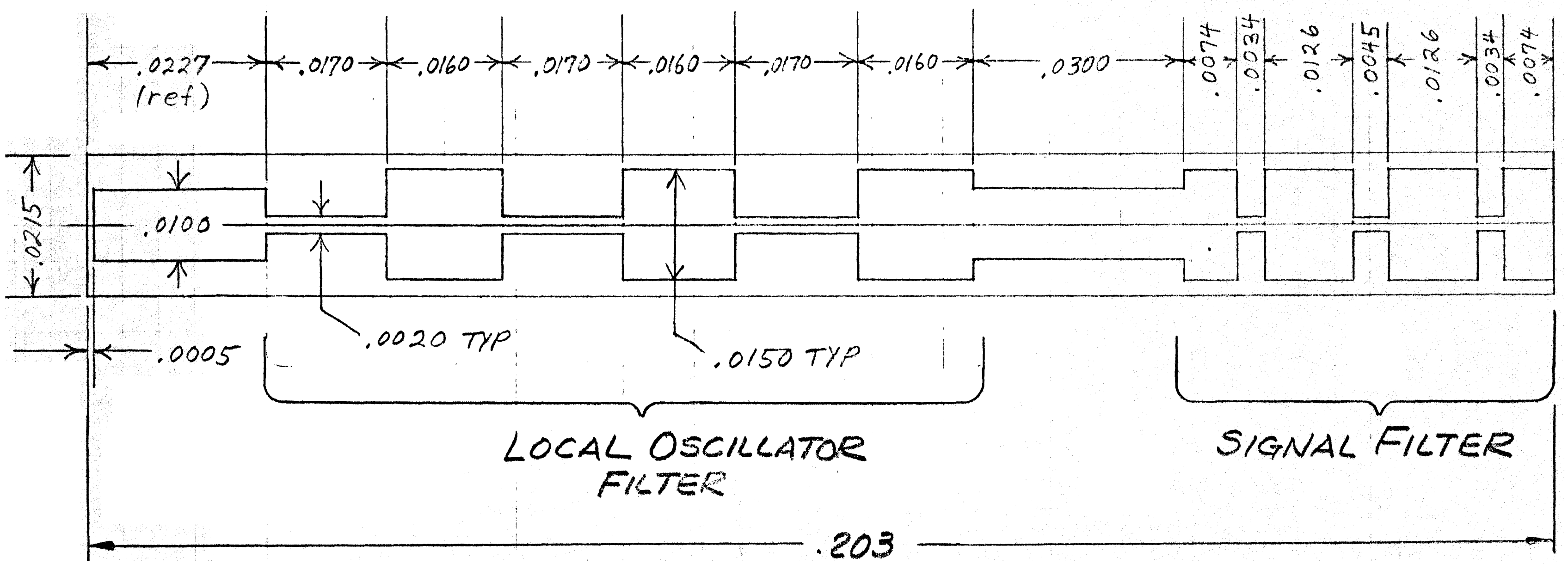
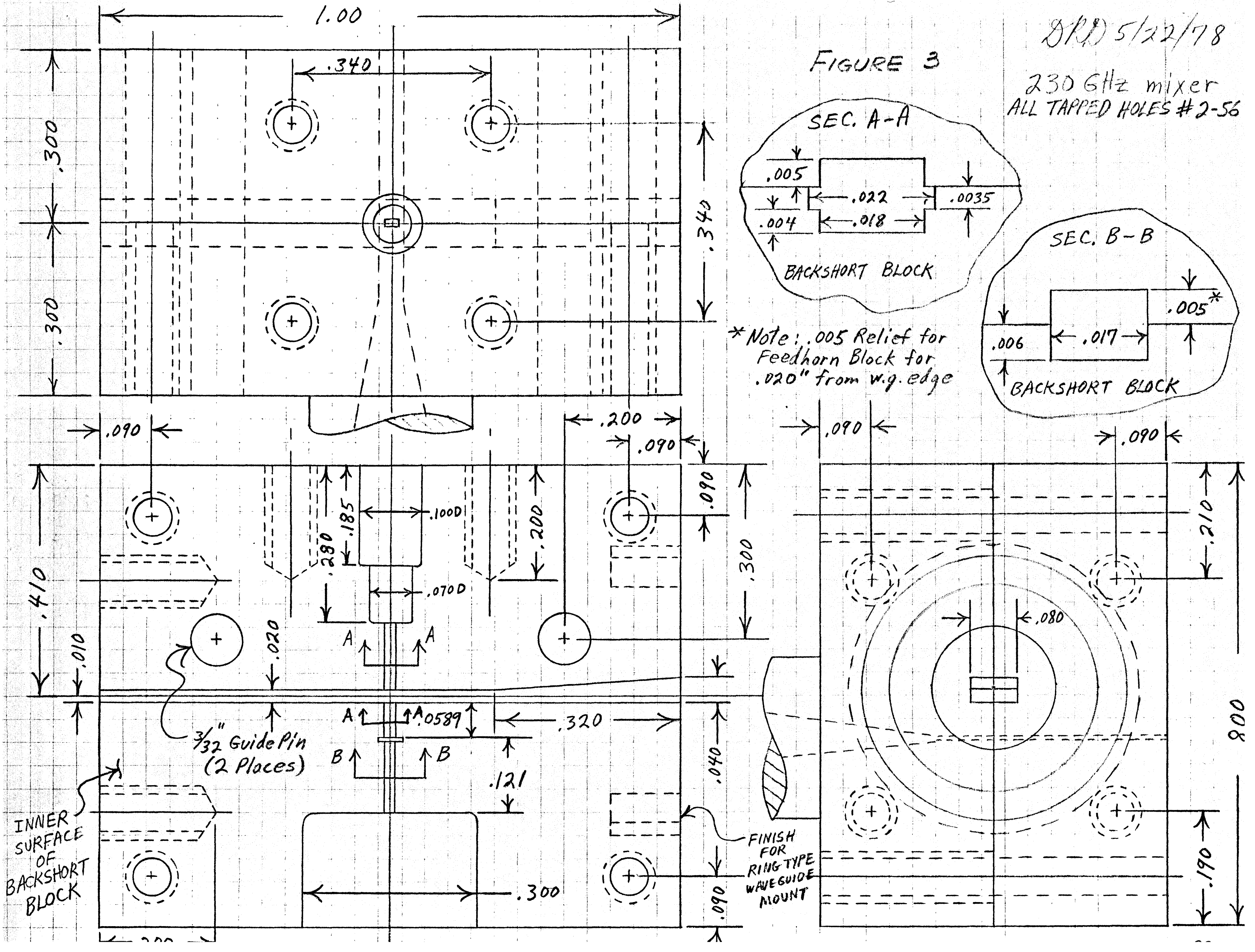


FIGURE 3

DRD 5/22/78

230 GHz mixer
ALL TAPPED HOLES #2-56



DRD 9/19/77

FIGURE 4 — 230 GHz Scalar Feed Horn

Detail of Mandrel for Electroforming

Mat'l: Al   Req'd: 4

Note: Fabricate 4 sections of .043 x .0055 Al waveguide for backshort to be electroformed simultaneously

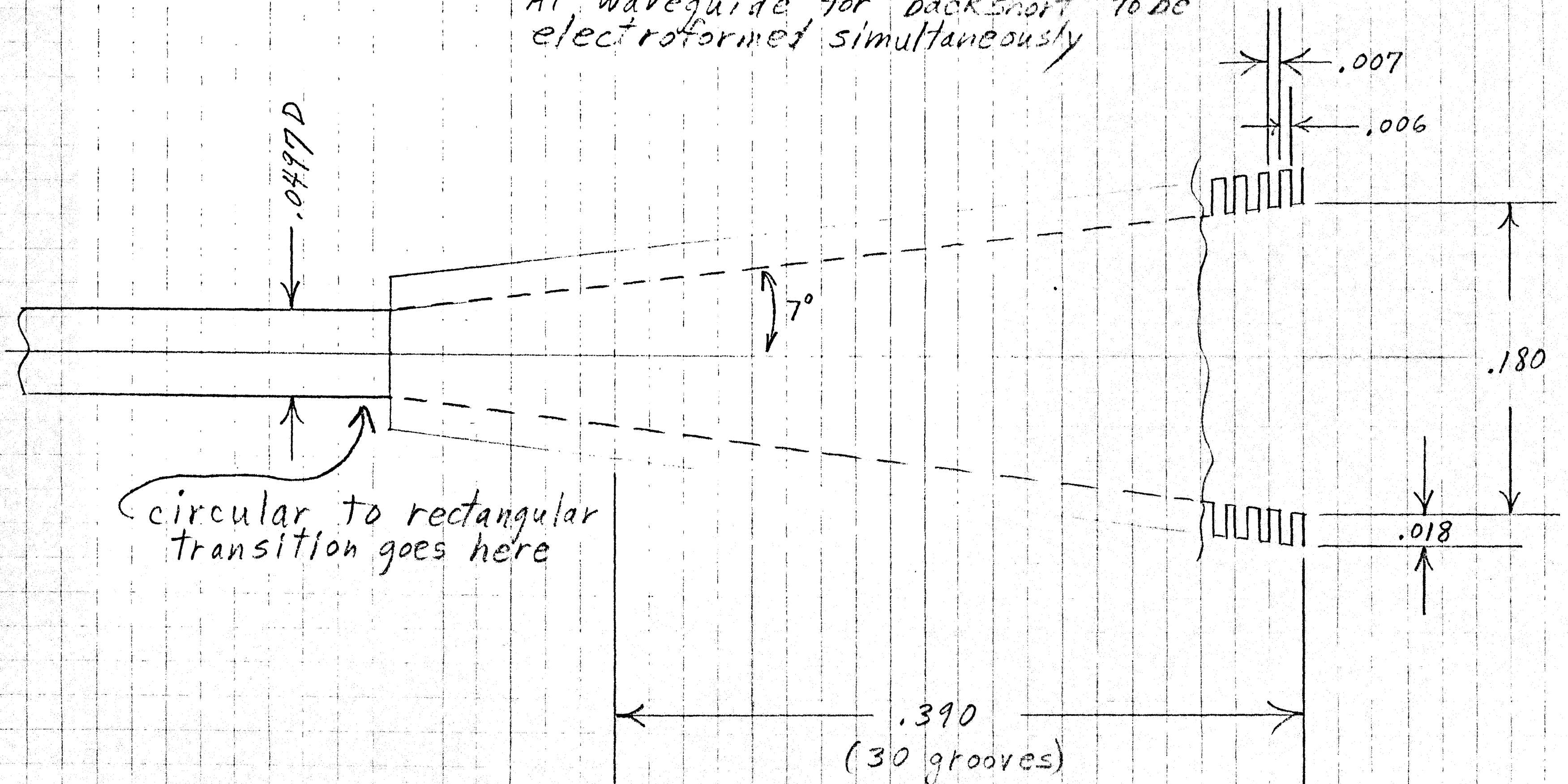


FIGURE 5- CIRCULAR TO RECTANGULAR TRANSITION
AND STEP IMPEDANCE TRANSFORMER

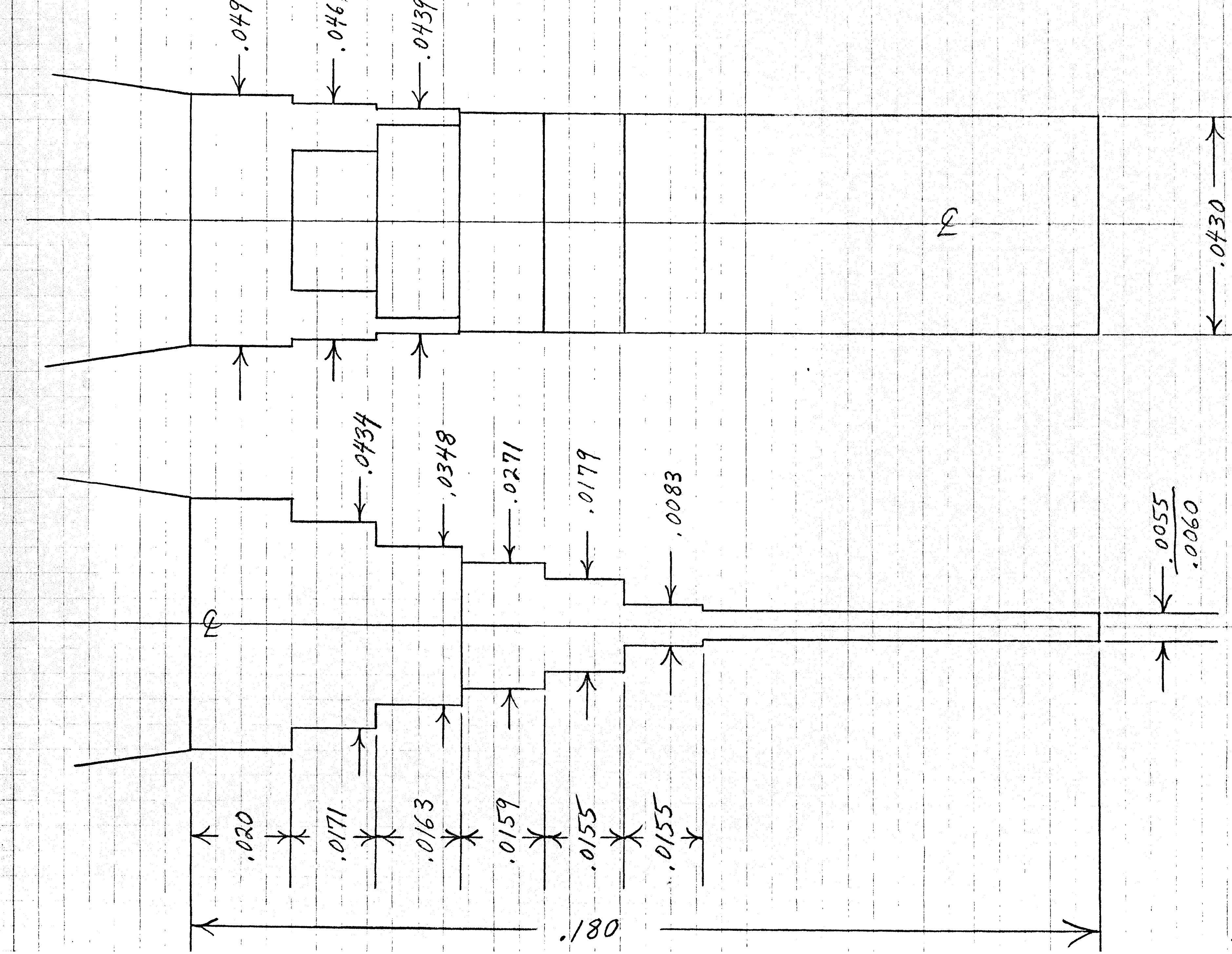
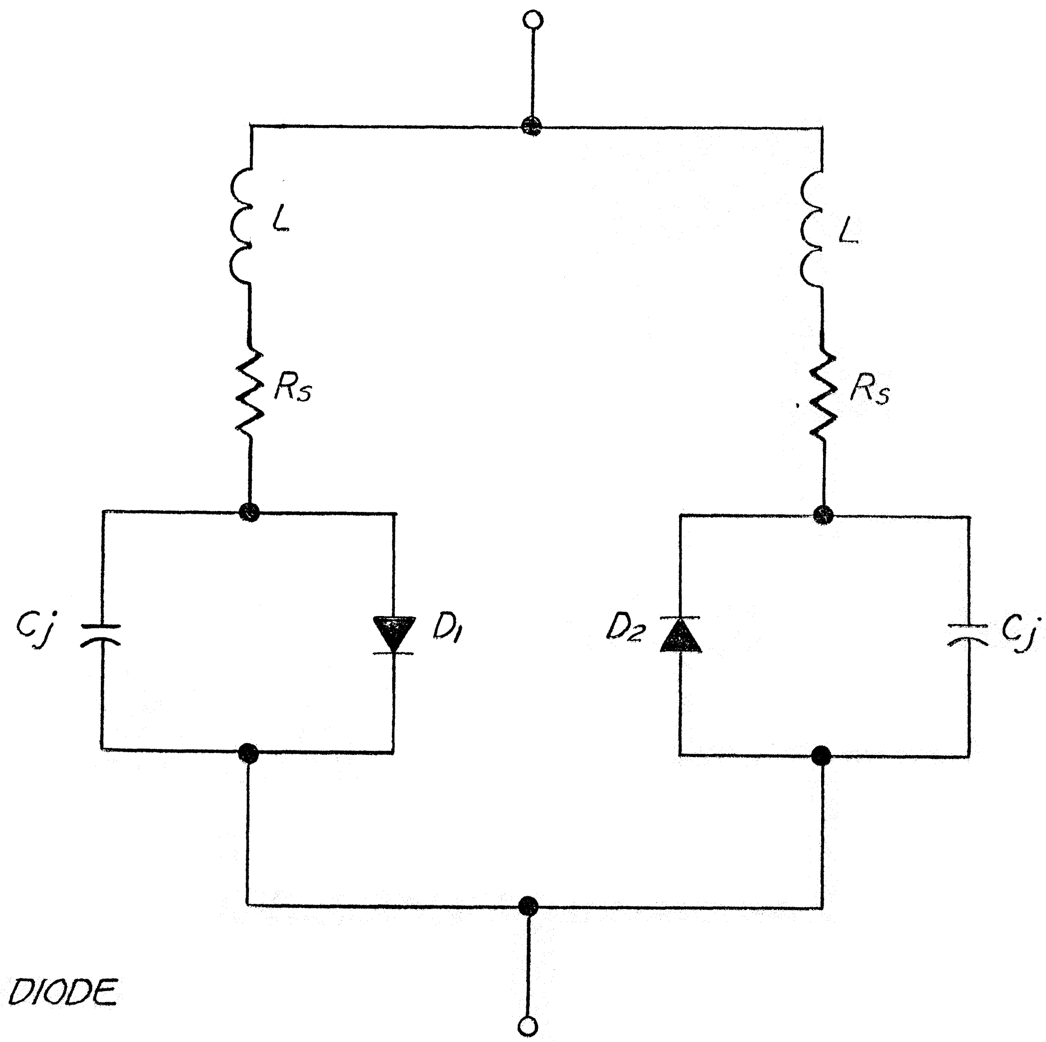


FIGURE 6 - EQUIVALENT CIRCUIT OF ANTIPARALLEL DIODE PAIR



C_j = 4 TO 8 fF
D_1, D_2 = EXPONENTIAL DIODE
R_S = 6 TO 13 Ω
L_W = 100 TO 140 pH

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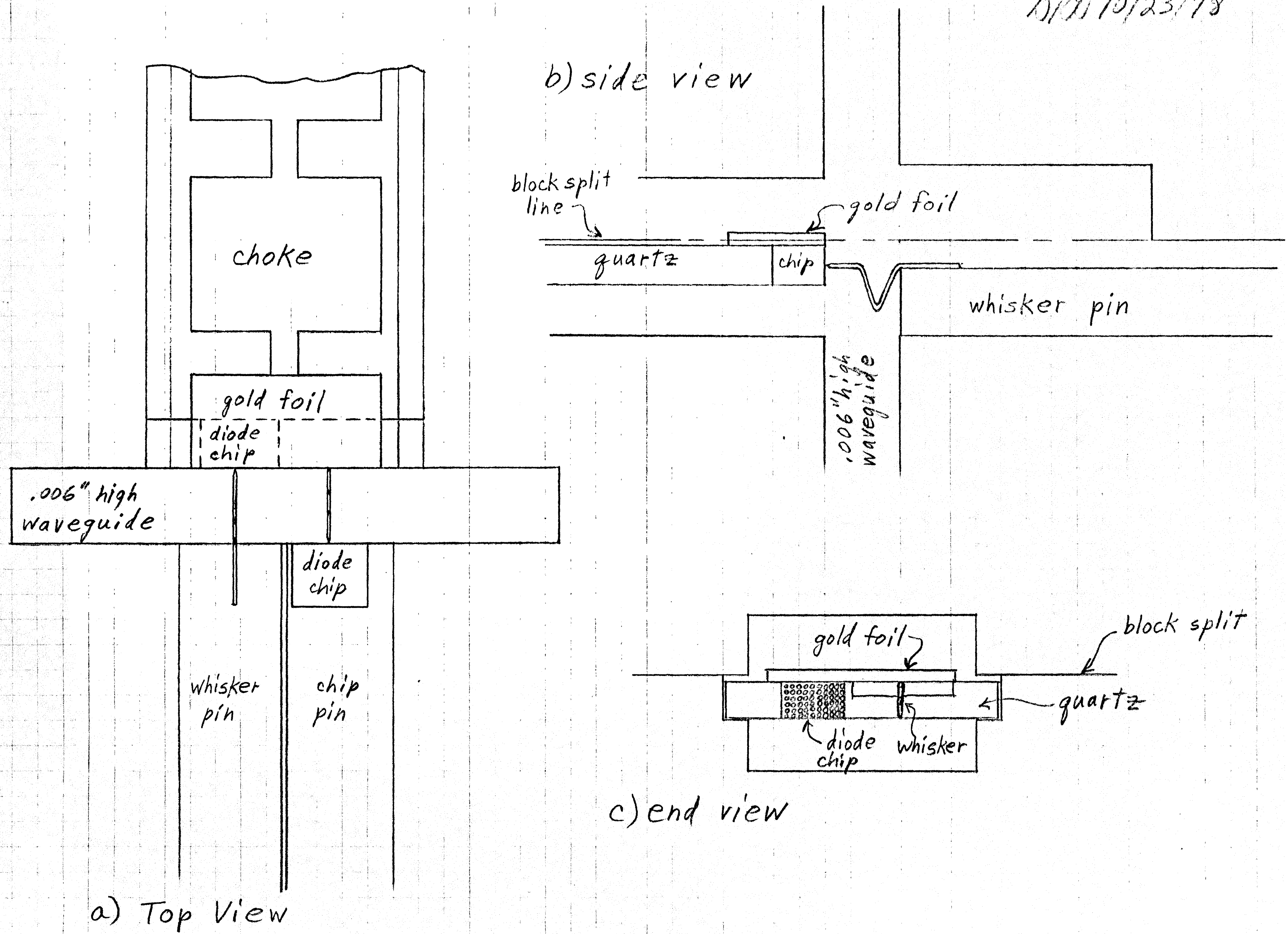

Figure 7 - Diode contacting geometry

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FIGURE 8 — 230 GHz Backshort

Mat'l : Be Cu

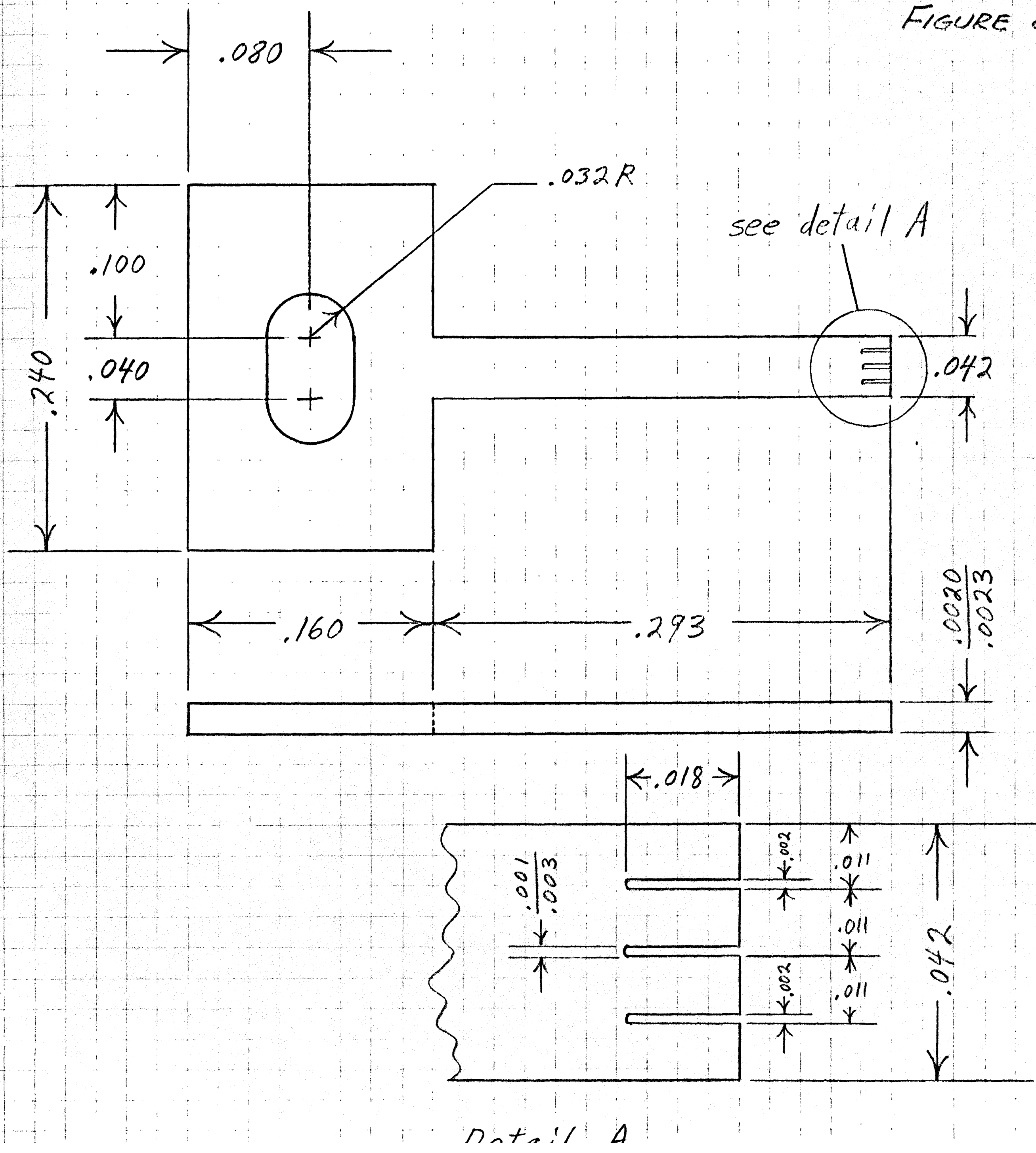

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FIGURE 9 - NOISE TEMPERATURE AND CONVERSION LOSS OF 230 GHz SHPM (DSB OPERATION)