AN ALL-SOLID-STATE - 1 THz - RADIOMETER FOR SPACE APPLICATIONS

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SUMMARY

A heterodyne receiver at 1 THz is described, whose components are waveguide mixer and frequency multipliers with Schottky-diodes as mixing and multiplying elements.

Most important is the breakthrough in local-oscillator development:

The power at the fundamental frequency is generated by an InP-Gunn-Oscillator which delivers 70 mW at 111.2 2 GHz. The output power from two cascade triplers (at the 9th-harmonic) was 60 μW (min.)

Coupling of signal- and L.O.-power is achieved by a Martin-Puplett Interferometer. System noise temperature was determined with absorbers at room temperature and 77 K yielding Tsys = 8800 K (DSB) and total conversion loss = 13dB. Atmospheric losses appear to account for half of the noise temperature and 3dB conversion loss. The mixer is prepared for cooling to liquid nitrogen from which we hope for a further reduction in system noise of approximately a factor of two.

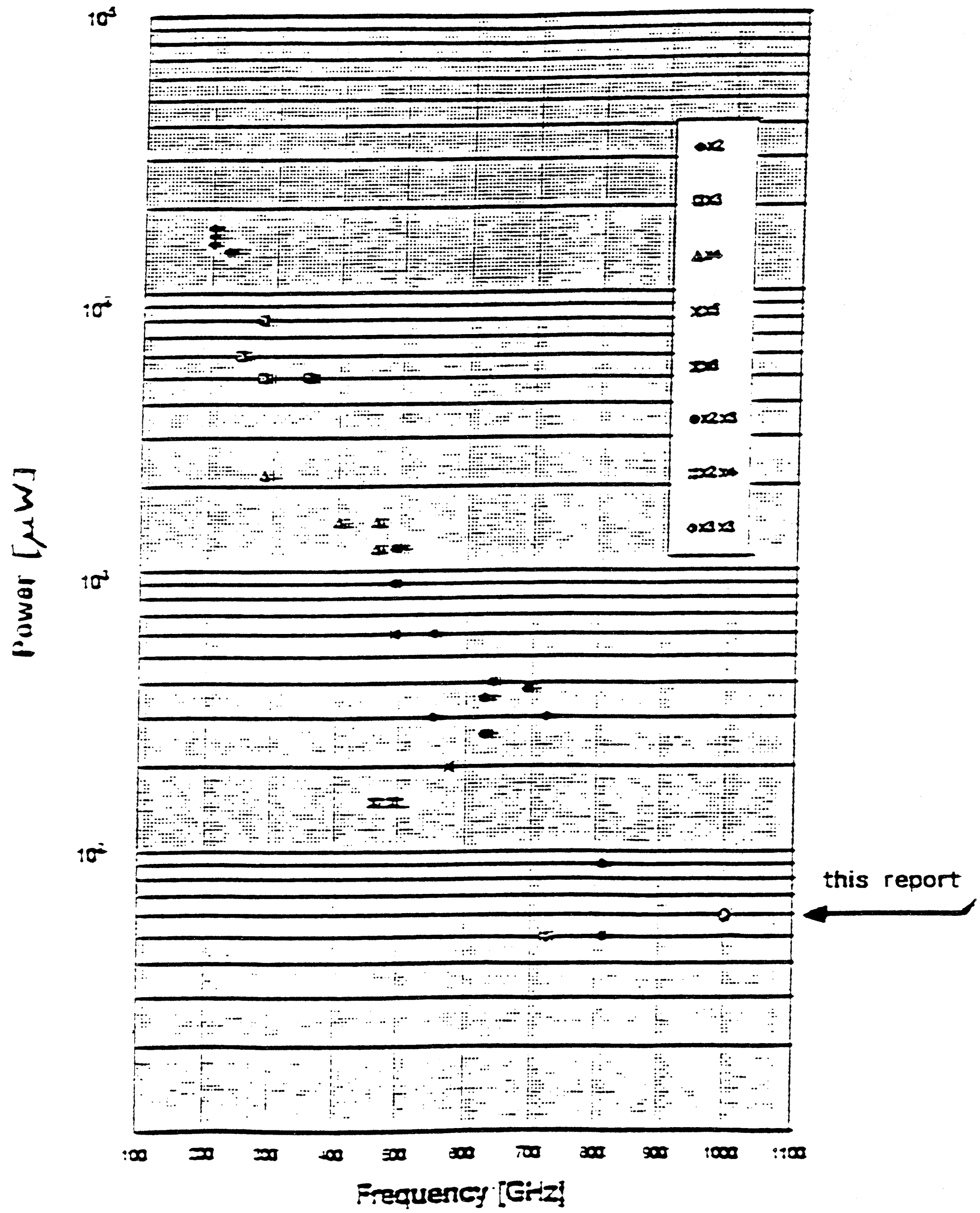
INTRODUCTION

A study for the FIRST - project of the European Space Agency requires submm-receivers up to 1200 GHz for a spaceborne mission, with low weight, low power consumption and high receiver sensitvity. Despite considerable improvements in SIS-receivers over the past years, their performance at and above 1000 GHz is

SIS-receivers over the past years, their performance at and above 1000 GHz is unknown and their liquid helium requirement could be a drawback for space operation.

Schottky-receivers could be operated in space with closed-cycle refrigerators at 80 K, but up to now only carcinotrons or lasers have been suitable as LO-sources. Recent improvements in solid-state oscillators up to 800 GHz (see Figure1) have indicated that all-solid-state LO's will be usable to above 1 THz.

Figure 1: Output Power of Local Oscillator sources



LOCAL OSCILLATOR DESIGN CONSIDERATIONS

From practical experience with receivers at 550 GHz and 650 GHz [1,2] we estimate that a cooled Schottky diode requires approximiately 100 μW output power, in order to achieve lowest conversion loss. Since this is the most critical step in building an all Solid-state-Schottky-receiver, we describe how the final L.O. design was chosen by employing a varactor-diode multiplier driven by a solid-state fundamental oscillator.

As fundamental oscillators we considered gunn-oscillators in the range 100-125 GHz, due to their excellent spectral properties and space compatibility. In 1991, 50 mW was obtainable at 111 GHz and 25 mW at 125 GHz, today the figures are (respectively) 70 mW and 50 mW.

The basic theory of frequency multiplication by varactor-diodes is given by Penfield and Rafuse [3] and has been used to create a multiplier analysis program by Siegel, Kerr and Hwang [4]. On the basis of this program we evaluate which multiplication factor might produce the highest efficiency.

The selection of suitable varactor diodes commenced by taking the parameters given by the manufacturer. Table 1 lists the principal parameters: series resistance (R_s),junction capacitance (C_j)and breakdown voltage (V_{Br}) in the reverse direction.

Unfortunately the key parameter in varactor applications ,the capacitance modulation ,is not given. It determines the figure-of -merit or the dynamic cut-off-frequency (f_C)

$$f_c = \frac{S_{max} - S_{min}}{2\,\pi R_s}$$

where Smax and Smin are maximum and minimum values of the junction ellastance (= capacitance^{-1}) over one cycle of the pump waveform. Measurements at RPG yield capacitance modulation results also shown in Table 1.

TABLE 1

Diode	R_S [ ]	C_j [fF]	V_{Br} [V]	f_C [GHz]
VD 011*	10.6	12.3	16.5	1370
VD 012*	14.7	9.3	15.0	1440
6P4 **	8	19	20	2200
2T2 **	12.4	5.6	10.8	4800
2T8 **	14.7	4.0	10	n.m.

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Generally the diode with the highest cutoff frequency leads to the highest efficiency , but the input power is restricted by the breakdown voltage in the reverse direction and the maximum safe current in the forward direction. Since high input power and reliability is required for effective operation, the 6P4 was chosen for the first multiplier stage. For the second stage, and generally speaking for higher input frequencies, the 2T2 was the diode of choice due to the high-cutoff frequency <u>and</u> low capacitance.

In principle multiplication can be achieved in a single diode, but the efficiency for x8, x9 or x10 is very low. In addition there would be the difficulty of implementation due to the many idlers which have to be correctly terminated. Hence we considered cascaded multipliers of x2x2x2, x2x4, x2x5 and x3x3, using the diagrams for efficiency in [3] and the multiplier program [4].

The results are shown in Table 2.

TABLE 2

Multiplicat. factor	Input frequ. (GHz)	Cutoff frequ. (GHz)	Input power (mW)	Efficiency (%)	Output power (mW)	
1.) x2	125	1400	25	25	5.5	(2.75)
x4	250	2000	2.75	0.5	<0.002	
2.) x2	125	1400	25	22	5.5	(2.75)
x2	250	2000	2.75	15	0.4	(0.2)
x2	500	2000	0.2	1	0.002	
3.) x3	100	1400	50	30	15	(7.5)
x5	200	2000	7.5	0.2	<0.015	
4.) x3	111	2200	50	26	13	(6.5)
x3	333	2000	6.5	5,5	0.36	(0.18)

Values in parantheses are for estimated 3 dB mount losses

It can be seen:

a.) the octupler combinations x2x4 and x2x2x2 deliver output power of only 2μW. This does not change significantly even if the fundamental power is raised to 50 mW.

b.) The x2x5 suffers from the problem that the efficiency of the second stage (quintupler) is too low.

c.) The x3x3 combination produces the most output power, whereby the first stage can deliver as much as 6.5 mW and the second stage up to 180 μW.

The calculations do not consider the velocity saturation effects as theoretically described in [5] ,and thus it is expected that the actual output powers will be significantly lower than predicted. However, it is hoped that the analyses are accurate enough to compare the various multiplier designs and justify our choice of a x3x3 multiplier chain.

Table 3 shows the predicted performance of a tripler to 990 GHz with a 2T2 when velocity saturation effects are neglected and when the diode current is clipped at 23 mA to simulate these effects. With 10mW of input power assumed, the result is a drop in output power from 1.2 mW (n = 11.6 %) to 0.06 mW (n = 0.6 %).

TABLE 3 from [5]

Table 3: *Calculated performance of 990 GHz Schottky Barrier Diode Varactor (2T2) Multiplier.*

	P_{in} (mW)	i_{RF} (mA)	dW/dt 10^5 (m/s)	η	R_s (Ω)	a
3x330	2	24	2.3	4.1	12	-
GHz	10	53	4.6	11.6	12	-
i_{sat}=	2	23	2.0	2.3	R(i)	$1.53 \cdot 10^9$
23 mA	10	31	2.9	0.6	R(i)	$1.53 \cdot 10^9$

An accurate analysis of a submillimeter wavelength multiplier circuit, including all of the embedded impedances at the fundamental and important harmonics, and an accurate model of the electron transport in the Schottky diode has not yet been developed. This is partially due to the extreme nonlinearity of the circuit. However, the complex nature of the electron transport in the diode also complicates such an analysis. Traditional multiplier analyses use the measured low frequency series resistance, junction capacitance and breakdown characteristics to generate a diode equivalent circuit that is assumed to be accurate at very high frequencies. As was initially pointed out by Kollberg , this is generally not a valid assumption. Charge carrier inertia and velocity saturation certainly affect multiplier performance and must be accurately incorporated into the analyses. Furthermore, the difficulty of modeling the diode's reverse breakdown due to avalanche multiplication in such a non-dc situation is also considerable. Attempts are now being made to develop more complete analyses of high frequency multipliers which will take these effects into account [6,7].

Due to these theoretical limitations, we chose simultaneously to pursue an experimental path toward varactor diode optimization. For example, it seems clear that diodes with thinner and more heavily doped epitaxial layers should be less likely to suffer from velocity saturation effects.This appears to have been confirmed in a recent publicattion [8] which compared two diodes, UVa 6P4 and 5T1, in a 100-200 GHz doubler. Both of these diodes are described in Table 4 and the newer diode (5T1) was also used in the first stage tripler for this paper.

Since velocity saturation effects become more critical at higher frequencies, the second stage tripler is even more challenging. Up to the present, two diodes, UVa 2T2 and 2N1, have been used in this system. However, only the 2N1 diode, with its higher doping density and thinner epitaxial layer, has yieldied significant output power.

Table 4 also presents the characteristics of a variety of newer varactor diodes that have been fabricated with doping densities as high as $4x10^{17}$ cm^{-3} and zero-bias capacitances in the range from 3 to 6 fF. The trade-offs are clearly evident in this data: As the doping is increased the series resistance is reduced for a given zero-bias capacitance level, but the breakdown voltage is reduced. Thus, greater efficiency is expected but at the cost of reduced power handling ability. However, this appears to be a necessary trade-off for THz frequency multipliers.

TABLE 4

Nominal Parameters for Some Schottky Varactor Diodes

Diode Batch	Epitaxial Doping Density (cm^{-3})	Epilaxial Thickness (μm)	Anode Diameter (μm)	Series Resistance (Ω)	Breakdown Voltage (V)	Zero-bias Capacitance (fF)	Capacitance at Breakdown (fF)
6P4	$3.5x10^{16}$	1.0	6.3	9.5	20	20	5
5T1	$1x10^{17}$	0.6	4.7	5.3	10	22	7.5
2T2	$1x10^{17}$	0.6	2.5	11.5	10.7	5.5	1.9
2T8	$1x10^{17}$	0.6	2.5	11.5	10.7	5.5	1.9
2N1	$2x10^{17}$	0.26	1.8	9	9.8	5.3	2.4
2T12	$2.3x10^{17}$	0.24	2.1	7.2	8.2	5.5	—
2T13	$4x10^{17}$	0.16	1.8	5	5.9	6.2	—
2T14	$2.3x10^{17}$	0.24	1.7	8.5	7	4	—
2T15	$4x10^{17}$	0.16	1.4	8.5	6	3.3	1.6

LOCAL OSCILLATOR CONSTRUCTION

The local Oscillator comprises of a Gunn-oscillator, coupler, the tripler-tripler cascade and an output feedhorn (s. fig. 2).

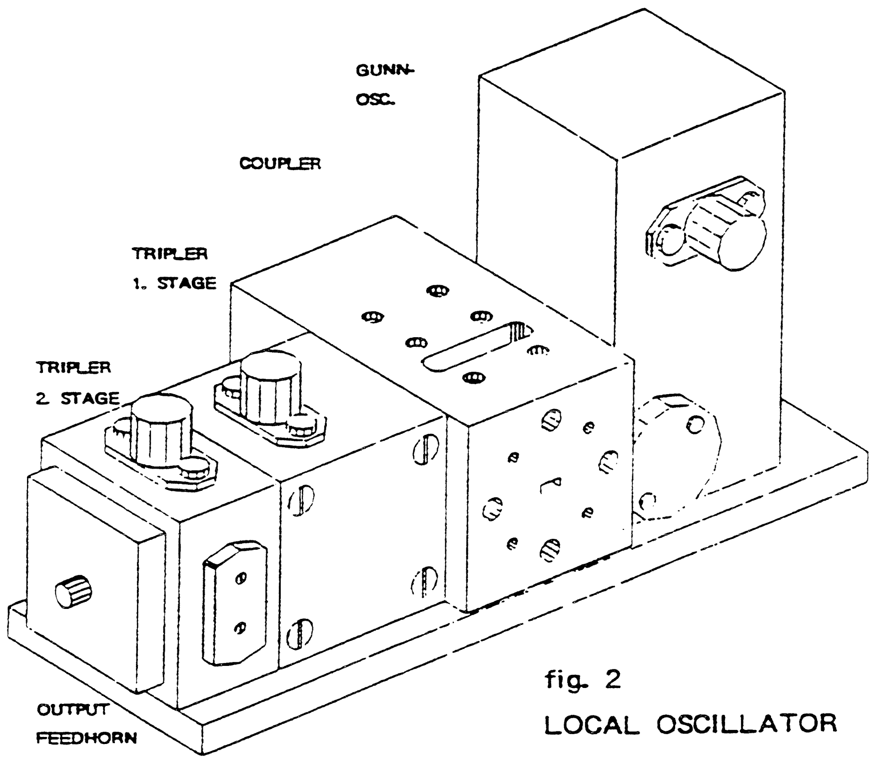

fig. 2
LOCAL OSCILLATOR

Contrary to the usual multiplier design, whereby input- and output-waveguide are at 90^{0} to each other (crossguide), we have developed an in-line-construction. This considerably simplifies construction when two or more multipliers have to be cascaded.

The principal details of the tripler mounts are shown in Fig.3.

These mounts consists of three parts:

The UPPER BLOCK contains the bias connector , a rejection filter to prevent leakage of the fundamental frequency and the input waveguide. The MIDDLE BLOCK contains the coupling structure for the fundamental frequency to the diode and is simultaneously a rejection filter for the harmonics. The LOWER BLOCK contains the output waveguide, idler cavity and the whisker post.

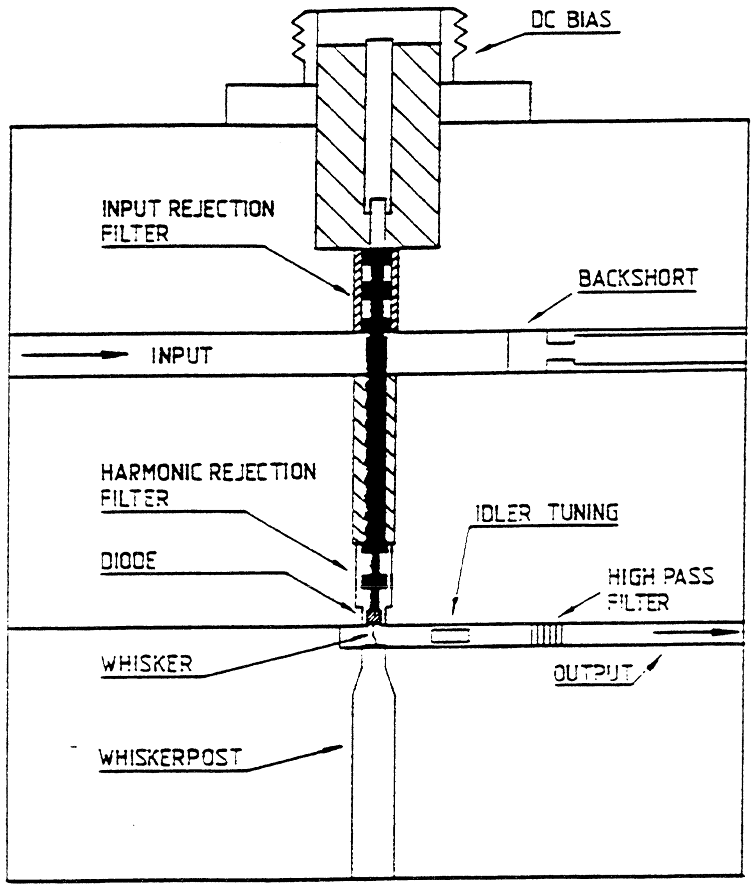

fig. 3 MULTIPLIER SCHEMATIC

Critical features of the design are:

a.) Input power coupling structure

This includes the waveguide and a low pass-filter with low losses at the fundamental frequency and high rejection at all higher harmonics. The waveguide is single moded and reduced in height to provide efficient coupling to the coaxial probe, which is part of the filter. The waveguide bends 90° around this probe. The backshort is of the contacting loop-type, as this results in best performance up to the highest frequencies.

b.) Harmonic rejection-filter and diode

This filter is in-line with the coaxial probe, and is constructed in coaxial airline technique being held by a macor spacer.

This has the following advantages over microstrip:

1.) Lower dissipative losses

2.) If changes are required during multiplier development these can be achieved more rapidly.

The dimensions have to be kept very small, to prevent the excitation of the first coaxial waveguide mode, whose cutoff-wavelength is given by

$$\lambda = \pi (a + b)/2$$

(a , b = inner and outer diameter of coaxial section)

At 1000 GHz the first section $a + b$ should be ≤ 0.19 mm, which can be achieved with $a = 0.07$ mm, $b = 0.12$ mm. In addition, excellent coaxial alignment is essential, particularly when the inner section becomes wider.
The filter was originally from a Tchebycheff design and was then modified for best performance by modelling between approx. 2 and 20 GHz.

c.) output section

This consists of the whisker, idler tuning circuits, high-pass filter and output waveguide.
The whisker is a further key element for good multiplier performance. Firstly it provides a DC-contact to the diode and should provide an ohmic contact without adding any series-resistance. Secondly the whisker length and diameter are critical since they determine the inductance in series with the diode.
The output waveguide was chosen to be cutoff for idler and fundamental frequencies. It was matched to the diode-whisker idler circuit by a moving backshort.
The dimensions of the output-waveguide became as small as 0.27x0.05mm to prevent losses into higher modes.

LOCAL OSCILLATOR RESULTS

The output power of the gunn-oscillator at 111.22 GHz was measured with a Hughes Thermistor Mount. The result was 70 mW $\pm$ 5%.
With 70 mW input power the output power of the first tripler was 7 mW. This was measured with a TK-powermeter * in a quasioptical setup, no corrections for possible inaccuracy of beam-matching or atmospheric losses are applied. We estimate the accuracy of this measurement in the range $\pm$ 10%.
Though the result in itself is quite acceptable, the achieved efficiency of 10% is low compared to the prediction of 26%. At present it is not clear whether the discrepancy is due to dissipative or mismatch losses or to the velocity saturation effect.
The second tripler, delivered a minimum output power of 60 µW. This was again measured with a TK-power meter. The accuracy of this result is difficult to estimate due to the problem of atmospheric absorption . The variations during the measurement indicate higher power.
An indication about the atmospheric losses at 1000 GHz is given by a measurement of rectified power at a mixer diode: when the mixer-feedhorn was placed against the multiplier output feedhorn, the increase in mixer current was three times more than when the mixer feedhorn was illuminated through the quasioptical path in the receiver. From our experience in building receivers with low loss quasioptical components we estimate that most of the loss was due to atmosphere: we estimate 3 dB.

QUASIOPTICS AND MIXER

Signal and local oscillator were coupled into the waveguide mixer through a Martin Puplett polarisation diplexer [9] and a dual-mode horn [10]. The coupler consisted of 10µm tungsten wires with 60µm spacing.

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Two elliptical mirrors served as beam matching devices to mixer and multiplier feedhorn respectively.

The mixer is a scaled version as described in [11] with a waveguide cross-section 270x50μm . The mixer diode chip, a 1T15 from Semiconductor Device Lab. of Univ. of Virginia was mounted on a four-section coaxial filter protruding through the middle of the broad wall of the waveguide. This diode is described in detail in [12] and was specially designed for terahertz use. The epitaxial layer doping density was $1\text{x}\ 10^{18}\ \text{cm}^{-3}$ and the anode diameter was 0.25 microns. The measured DC series resistance was roughly 25 Ohm and the zero-bias capacitance was 0.25fF, yielding a figure-of-merit cut-off frequency of of 25 THz ($1/2\pi R_S.C_j$). The whisker-wire was eight microns in diameter with a sharply pointed tip. This diode has been specifically developed for THz mixing applications and has already yielded excellent results in corner-cube mixers at 2.5 THz [13]. At 1 THz it is not expected to be the most sensitive mixer diode available, but was chosen for this work because of its low local oscillator power requirement.

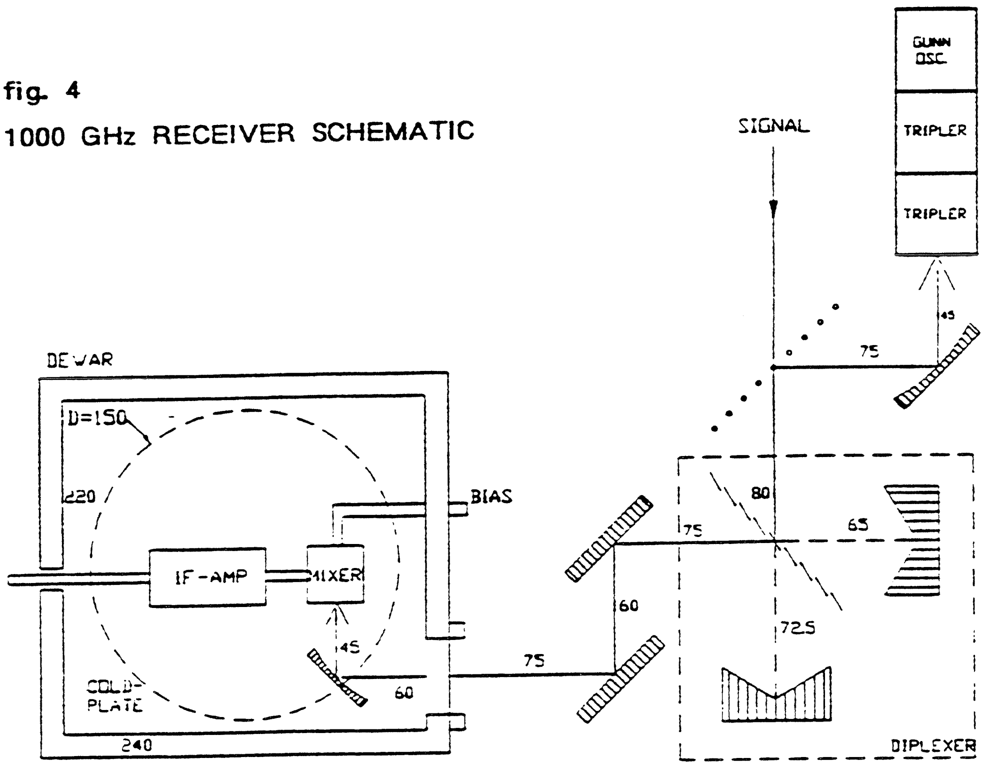
The output of the coaxial filter was matched to the IF amplifier by a transformer centred at 1.4 GHz.

RECEIVER RESULTS

A schematic of the receiver is shown in fig. 4 . The mixer is mounted in a dewar in preparation for later cooling to liquid nitrogen. The IF amplifier was centred at 1.4 GHz with 100 MHz bandwidth, and had a noise temperature of 35 K. System noise temperature was measured by the Y-factor method, using absorbing material (Emerson and Cumings AN-72) at 77 K and ambient temperature. No corrections are made for non-ideal absorbtivity (exact value not known) and errors due to assuming the Rayleigh-Jeans approximation (about 10 K, at this frequency).

The measured double-sideband noise temperature and conversion loss were respectively 8800 K and 13 dB.
Since, an improvement of about 3 dB could be obtained by correcting for atmospheric (water vapour) losses in the quasioptical path, we would obtain a noise temperature of 4400 K and conversion loss of 10 dB; we hope for an additional factor 2 improvement by cooling to 77 K.

fig. 4
1000 GHz RECEIVER SCHEMATIC



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