

**A QUASI-OPTICAL Ka BAND SUBHARMONIC MIXER
WITH SEPARATELY BIASED DIODES ON A PLANAR
ANTENNA**

A Thesis

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Abstract

The design and development of inexpensive, high power LO sources is one of the major challenges for researchers today. Low available power from solid-state sources, poor efficiency of present-day multipliers and lack of tuning capability of far-infrared lasers are the major motivations for looking into receiver components whose LO power requirements are more easily achieved. Harmonic mixing is certainly one such avenue.

A subharmonic mixer with anti-parallel diodes requires a lower frequency LO signal and has the added benefit of reduced LO noise and suppression of fundamental and other harmonic mixing products. The LO power requirement for anti-parallel diode subharmonic mixers can be further reduced by biasing the diodes separately. However, it is not very easy to individually bias diodes that have been integrated on conventional quasi-optical coupling structures. In this research, a new coupling structure using a coplanar transmission line is proposed. The structure allows separate biasing of the diodes, a matching transformer between the diodes and antenna impedance, and coplanar lowpass filtering for the IF. A Ka band proof-of principle mixer is designed and developed which shows lowering of LO power requirement with biasing.

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List of Symbols

Symbol	Definition
C	Capacitance
C_j	Junction Capacitance
E	Electric field
E_c	Bottom of conduction band energy level
E_f	Fermi energy level
E_v	Top of valance band energy level
f_l	Local Oscillator frequency
f_p	Pump frequency
f_s	Signal frequency
g	Diode conductance
GHz	Gigahertz
H	Magnetic field
I_s	Diode saturation current
k_d	Propagation constant in the dielectric
k_z	Propagation constant in the guide
L	Inductance
m	Integer
n	Integer
Nb	Niobium

Symbol	Definition
pF	Picofarad
q	Unit charge - 1.6×10^{-19} C
R_s	Series resistance of Schottky diode
SIS	Superconductor Insulator Superconductor
TE_{xx}	Transverse electric field
THz	Terahertz
V_{bi}	Diode built-in voltage
V_{LO}	Local Oscillator voltage
V_{TO}	Diode threshold voltage
W	Coplanar transmission line width
w_0	Beam waist of a Gaussian beam
Z_{ant}	Antenna impedance
α_c	Conductor loss in coplanar waveguide
α_d	Dielectric loss in coplanar waveguide
α_r	Radiation loss in coplanar waveguide
ϵ_{eff}	Effective permittivity
ϵ_r	Relative permittivity
η	Diode ideality factor
$\phi_{barrier}$	Schottky barrier height
Φ	Divergence angle of a Gaussian beam

Symbol	Definition
ω_L	Local Oscillator frequency in radian
ω_s	Signal frequency in radian

Chapter 1

Introduction

The tremendous improvement in the field of device fabrication technology at millimeter and submillimeter-wave frequencies in recent years has created the opening for low noise, high sensitivity receiver design. Below 500 GHz, the most sensitive receivers incorporate SIS devices which have shown near quantum limited performance. However, the performance of SIS receivers degrades above the gap frequency of the superconducting elements (700 GHz for Nb). Until now, whisker contacted Schottky barrier diodes mounted in corner cube reflectors have been the most sensitive receivers in the terahertz range. Unlike SIS receivers, Schottky diode receivers need no cooling and DSB mixer noise temperatures of about 4000 K at 2.5 THz have been obtained [1]. However, the LO power requirement of Schottky mixers is much higher than for SIS devices. Also the arduous task of whisker contacting and the fragile nature of whiskers has been a major concern towards the reliability and ruggedness of these receivers. At present, the lack of reliable, high power solid-state local oscillators is a major limitation in the development of millimeter and submillimeter-wave Schottky

diode receivers. Although electron tube sources and far-infrared lasers can provide ample RF power in this frequency range, size, power requirements, lack of tuning capability and difficulty in maintenance have limited their usefulness and make them unsuitable for many applications. The alternative is to use a solid-state source in conjunction with a multiplier. The poor efficiencies of present day multipliers, coupled with their small size, results in solid-state sources having relatively low output power. The problem can also be approached from the detector's point of view by building receiver components whose LO power requirements are more easily achieved.

1.1 Subharmonic Mixers

Design and development of inexpensive, high power LO sources is one of the major challenges for researchers. The power available from solid-state sources drops off with the inverse square of frequency due to electronic limitations in the material, and hence, at higher frequencies, higher LO powers come at a much higher cost. Therefore, one of the main goals of terahertz mixer design has been the reduction of LO power requirements, with emphasis towards receiver configurations that permit harmonic mixing [2, 3, 4]. The advantages of harmonic mixing surpass the disadvantages (higher conversion loss compared to fundamental mixing) when a pair of anti-parallel diodes are used as mixer element. This has the added benefit of reduced LO noise, suppression of fundamental and other odd harmonic mixing products, and also the suppression of the even harmonics of the LO. For the anti-parallel diode pair shown in Figure 1.1(a), the pump signal must have sufficient power to turn on each diode

once in a single RF cycle, i.e., the V_{LO} must swing from $-V_{TO}$ to V_{TO} . It is clear that a subharmonic mixer employing anti-parallel diodes requires more power than an optimally biased single diode mixer. One possible improvement would be to design the mixer and associated coupling structure in a way which permits separate biasing for each of the diodes in the anti-parallel pair (Figure 1.1(b)). Since each diode is biased near V_{TO} , the V_{LO} does not need to swing all the way from $-V_{TO}$ to V_{TO} .

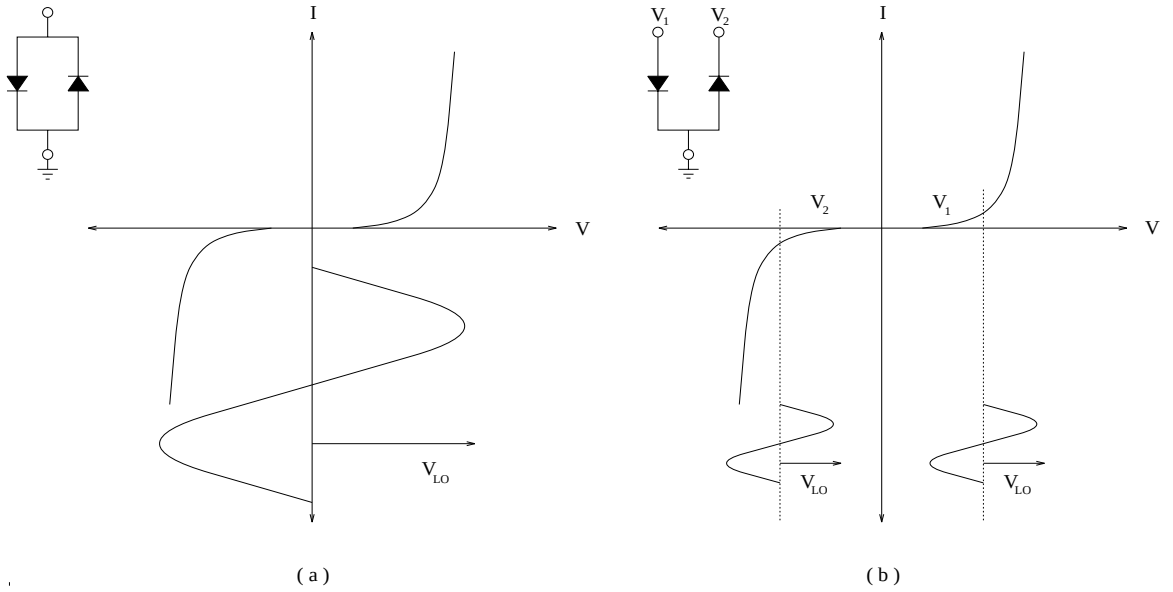


Figure 1.1: I-V curves for unbiased and biased diodes

Efficient coupling of LO and RF signals to the diodes is one of the essential requirements for achieving the lowest possible conversion loss and highest receiver sensitivity. At submillimeter wavelengths, integrated circuit antennas are, perhaps, the most convenient coupling structures. Lens coupled log-periodic and spiral antennas are extensively used in quasi-optical mixer designs. Because of the self-complementary nature of these broadband planar antennas and the diode geometry, separate biasing

of the individual diodes is not straight forward. Care must be taken not to disturb the symmetry in the system which permits subharmonic pumping. One circuit that can be used is shown in Figure 1.2. The terahertz research group at the University of Michigan (Rebeiz *et al.* [4, 3]) has successfully used this structure for subharmonic mixing at 90 GHz with individual biasing of the diodes. A split is made in the planar antenna to facilitate individual biasing of the diodes. An overlay capacitor on the split was used by Lee *et al.* [4] to maintain RF continuity.

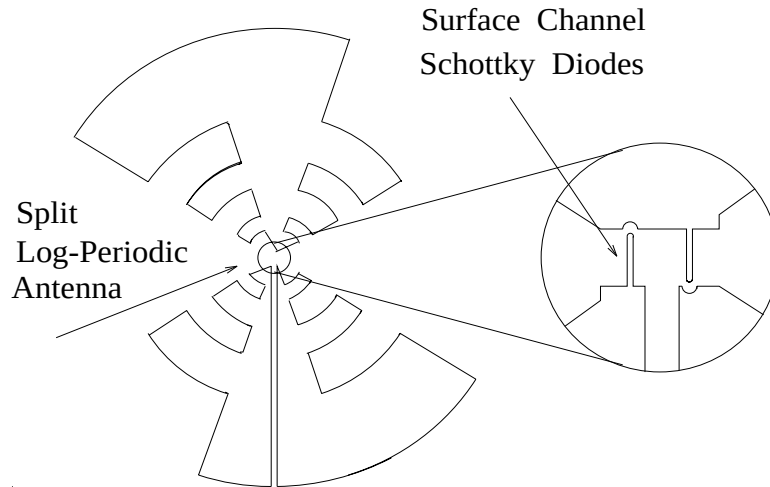


Figure 1.2: Schematic of anti-parallel diodes with a split log-periodic antenna

One of the limitations of this particular structure is that an impedance transformer can not be incorporated to match the diode pair to the antenna impedance. Also, there is no IF filter integrated in the structure, which may result in the LO and RF traveling with the IF all the way upto the IF connector. Reflections from the connector end may degrade the mixer performance. A new structure is proposed here which attempts to eliminate the above shortcomings. The schematic of the proposed design is shown in Figure 1.3. Instead of placing the diodes directly at the

antenna apex, a coplanar transmission line is used as a feed. This design allows more flexibility because the coplanar transmission line can be used as an impedance transformer between the antenna and the diode pair. The IF signal is extracted through a coplanar lowpass filter. When illuminated with radiation, the antenna launches an antisymmetric quasi-TEM mode on the coplanar line which, by symmetry, subharmonically pumps the diode pair. The presence of three separate metallizations in coplanar waveguide is also advantageous with respect to individually biasing the diodes. Two ground planes and the center conductor could each be kept at different voltages. Thus, this configuration is natural for subharmonic receiver applications.

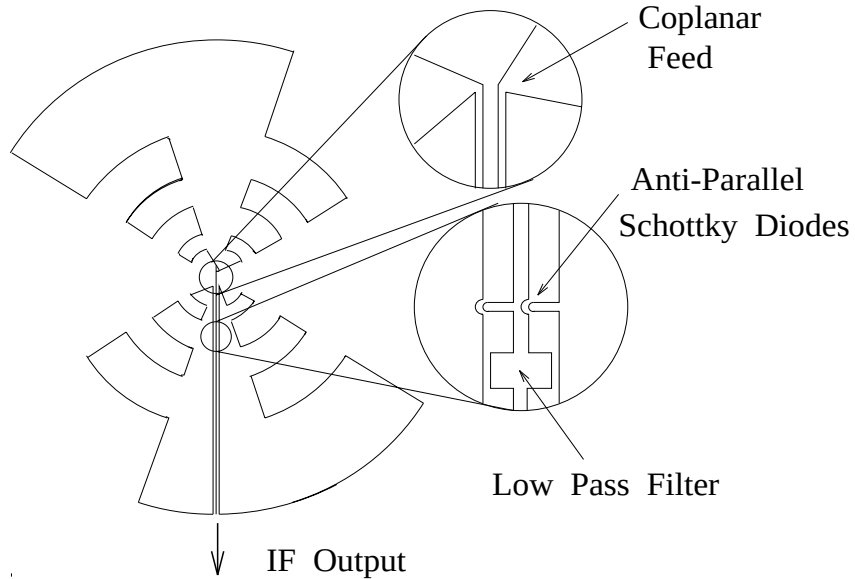


Figure 1.3: Coplanar coupled log-periodic antenna

A Ka band subharmonic mixer with separately biased anti-parallel diodes is designed and developed as a proof-of principle demonstration. It has been shown that with the biasing of the diodes, the LO power requirement is reduced. This design may be

frequency up-scaled to terahertz frequencies at a later date.

1.2 Organization of the Thesis

Chapter 2 gives a brief overview of subharmonic mixer theory and quasi-optical coupling structures. Chapter 3 presents the design, fabrication and the results of the subharmonic mixer. Chapter 4 describes the design of a 585 GHz diagonal horn to be used for SIS and planar Schottky diode receiver system. Chapter 5 summarizes the research and presents suggestions for further research.

Chapter 2

Theory of Harmonic Mixing

2.1 Mixer Overview

A mixer is fundamentally a multiplier. Figure 2.1 shows a multiplier block whose output consists of the sum and the difference frequencies of the input signals. Any nonlinear device can be used as a multiplier, and hence as a mixer.

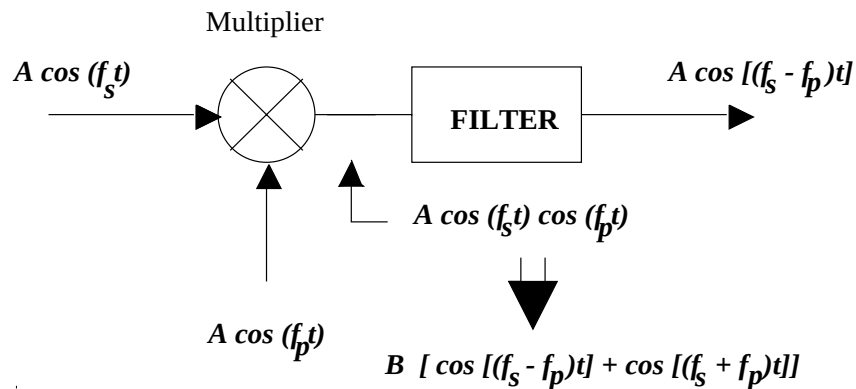


Figure 2.1: Multiplier Block Diagram

The I-V characteristics of a nonlinear device can be written using a power series,

$$I = a_0 + a_1V + a_2V^2 + a_3V^3 + \dots \quad (2.1)$$

If V is made equal to the sum of two different signals, after some trigonometric manipulations, it can be shown that the current contains components at frequencies $f_n = f_0 + nf_l$, where f_0 is the difference frequency $f_s - f_l$. The current also contains the harmonics of the LO, but it is easy to filter out the undesired frequencies and process the desired difference frequency.

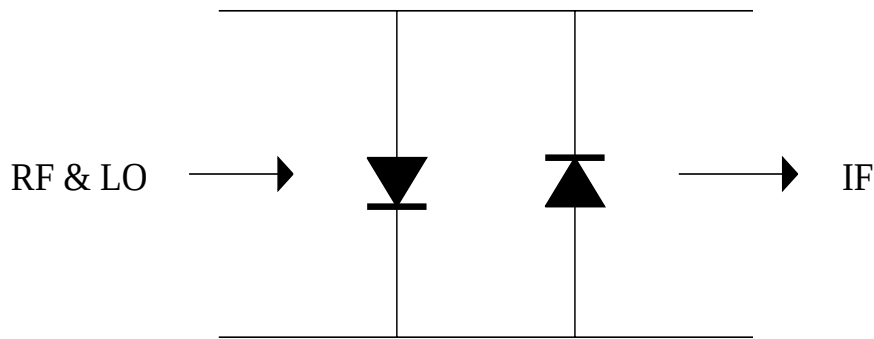


Figure 2.2: Anti-Parallel Diode Mixer

Schottky barrier diodes are the most popular nonlinear mixing element at millimeter and sub-millimeter-wave frequencies. They can be incorporated in waveguide or quasi-optical designs, have instantaneous bandwidths of several gigahertz and can cover the entire spectral range to 0.1 mm. As pointed out earlier, for many applications, it is extremely difficult, expensive and inconvenient to generate a fundamental frequency local oscillator signal at sub-millimeter wavelengths. To overcome this problem, quite often, a nonlinear mixing element is pumped with half the LO fre-

quency and the RF is mixed with the second harmonic of the LO generated in the nonlinear device. Though it is possible to have subharmonic mixing using a single diode, the fundamental mixing response is greater than the second harmonic response in such mixers [5]. As a result, the conversion loss in such mixers is greater. Instead, two diode mixers (anti-parallel configuration), as shown in Figure 2.2, give better performance in terms of conversion loss and noise performance. If the diodes used are identical, this configuration suppresses fundamental and other harmonic mixing products as well as even harmonics of the LO [6].

2.2 Schottky Barrier Diode

This section presents a brief overview of the Schottky barrier diode and describes its equivalent circuit model. Equivalent circuit models are very useful for the harmonic balance analysis of mixers.

Schottky barrier diodes are made by a metal contact to a semiconductor - the metal contact end acting as anode and the semiconductor end as cathode. The difference in work function between the metal contact and the semiconductor gives rise to the rectification property in the Schottky barrier diodes. It is a majority carrier device, because the conduction is due to the thermionic emission of the majority carriers over the barrier formed by the unequal work functions of the metal and the semiconductor. Figure 2.3 shows the band structure of the Schottky junction. Figure 2.4 and Figure 2.5 show the band structure for the forward and reverse biased Schottky junction.

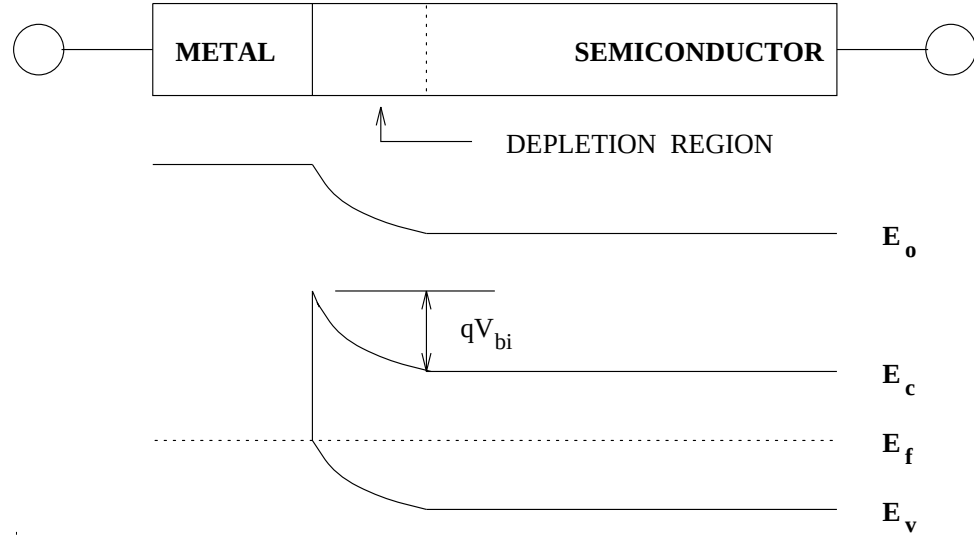


Figure 2.3: Band Structure of Schottky Diode

The current voltage characteristics of diode is given by :

$$I(V) = I_0(e^{qV/\eta KT} - 1) \quad (2.2)$$

where V is the applied voltage, q is the unit charge, T is the absolute temperature, K is Boltzmann constant, η is the diode ideality factor - which identifies the strength of the diode nonlinearity; and

$$I_0 = A^{**}WT^2e^{-q\phi_b/KT} \quad (2.3)$$

where A^{**} is the modified Richardson constant and W is the junction area. Figure 2.6 shows the equivalent circuit model of the Schottky diode. This intrinsic diode model has a nonlinear resistance and capacitance, and a linear series resistance. The series

resistance also varies with the junction voltage, but the variation is not significantly large and for most practical purposes can be neglected. This model does not show the parasitic capacitances and inductances which arise from diode metallization or lead geometry.

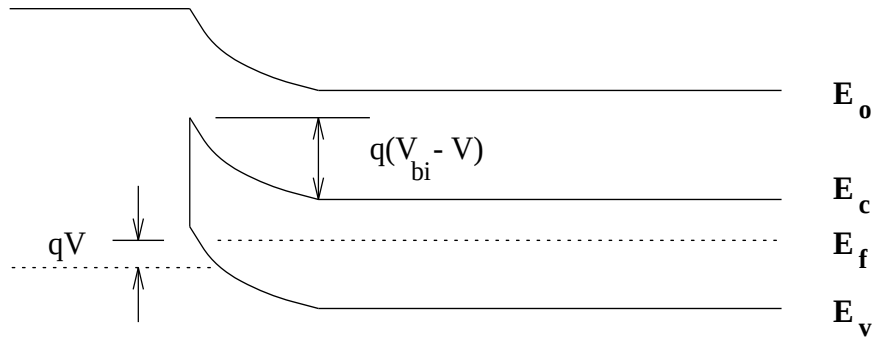


Figure 2.4: Forward Biased Schottky Diode Band Structure

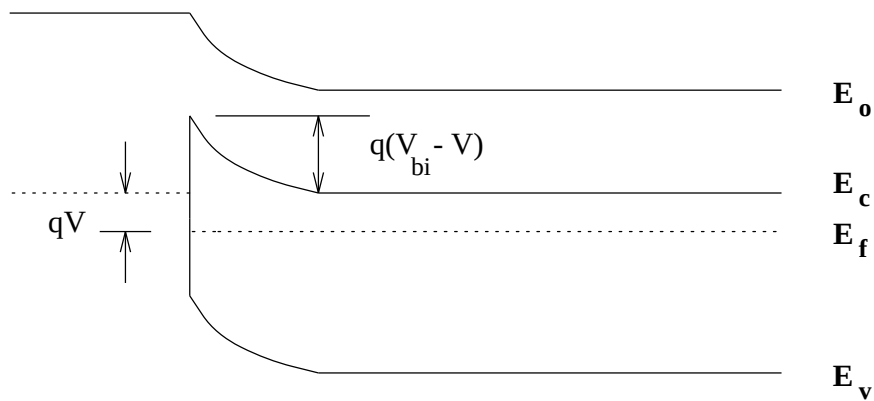


Figure 2.5: Reverse Biased Schottky Diode Band Structure

Schottky diodes can be of different kinds, depending on the fabrication methodology. For high frequency applications, whisker contacted diodes have historically been the most widely used. The whisker contacted diode has the advantage of minimum parasitics, and the usefulness of the whisker as a tuning element. However, the whisker

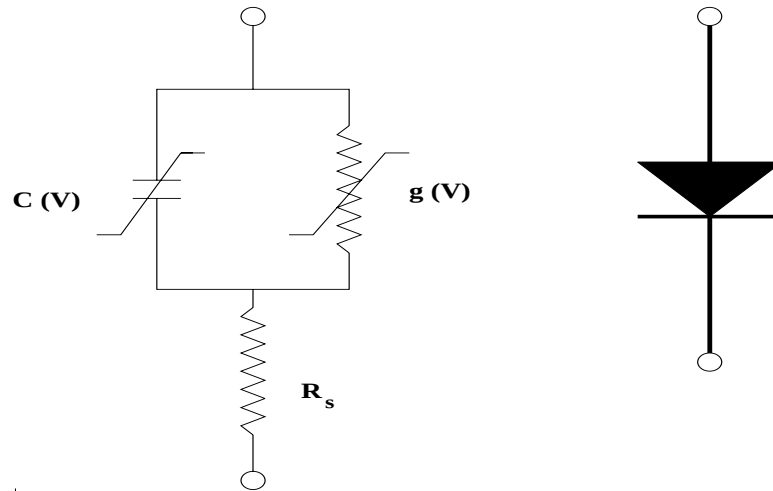


Figure 2.6: Schottky Diode Equivalent Circuit

is very fragile and the loss of contact is common under vibration and shock. On the contrary, planar Schottky diodes are rugged and can easily be integrated into arrays. The disadvantage of the planar Schottky diode is the added parasitics. Researchers at the University of Virginia have developed surface-channel planar Schottky diodes with very low pad to pad capacitance [7]. Figure 2.7 shows a planar Schottky diode structure.

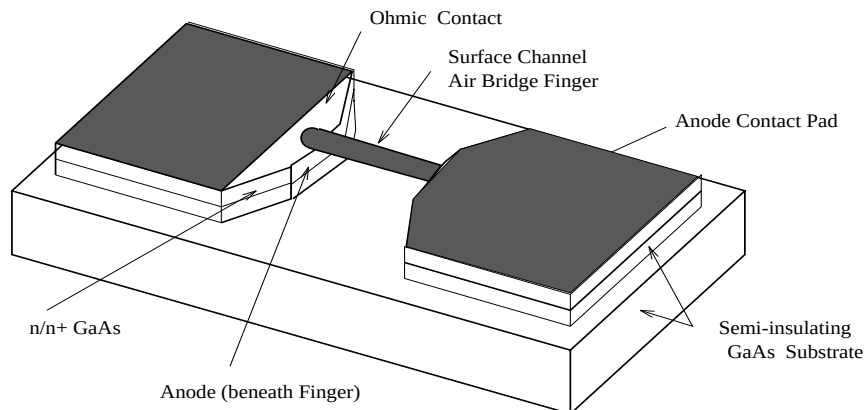


Figure 2.7: Planar Schottky Diode Structure

The planar surface-channel Schottky diode is well-suited for integration with planar

structures, such as integrated circuit antennas.

2.3 Subharmonic mixing with anti-parallel diodes

The balanced mixers have been one of the main building blocks of microwave engineering for many years. The symmetrical structure of the balanced mixers have two major advantages over single diode mixers - i) the down converted AM noise from the local oscillator (LO) does not appear at the IF output and ii) the signal and the LO power enter the mixer through separate ports, eliminating any external diplexer. Although two-diode subharmonic mixers have properties similar to balanced mixers (like AM noise suppression), the basic operating principle is different. In this section, the theory of the two-diode subharmonic mixer will be described with some mathematical details.

In a conventional single diode mixer, as shown in Figure 2.8, application of a voltage waveform

$$V = V_L \sin \omega_L t + V_s \sin \omega_s t \quad (2.4)$$

to the asymmetric diode I-V characteristic results in the diode current having all the frequencies $m f_L \pm n f_s$. However, in the case of an anti-parallel diode pair, as shown in Figure 2.9, the diode current contains frequencies for which $m + n$ is an odd integer. The terms for which $m + n$ is even, (i.e., even harmonics, fundamental mixing products $(\omega_s - \omega_L)$ and $(\omega_s + \omega_L)$, and the dc term), flow only within the diode loop. From Figure 2.9, the instantaneous current through the diodes can be written as :

$$i_1 = -i_s(e^{-\alpha V} - 1) \quad (2.5)$$

$$i_2 = i_s(e^{\alpha V} - 1) \quad (2.6)$$

where α is the diode slope parameter $q/\eta KT$. Similarly the differential conductance for each diode can be written as :

$$g_1 = \frac{di_1}{dV} = \alpha i_s e^{-\alpha V} \quad (2.7)$$

and

$$g_2 = \frac{di_2}{dV} = \alpha i_s e^{\alpha V} \quad (2.8)$$

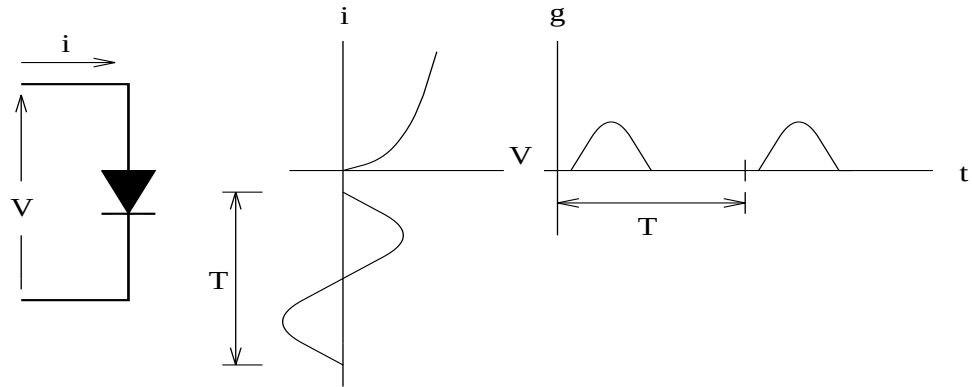


Figure 2.8: Single diode Mixer

The composite time varying differential conductance is given by the sum of these two

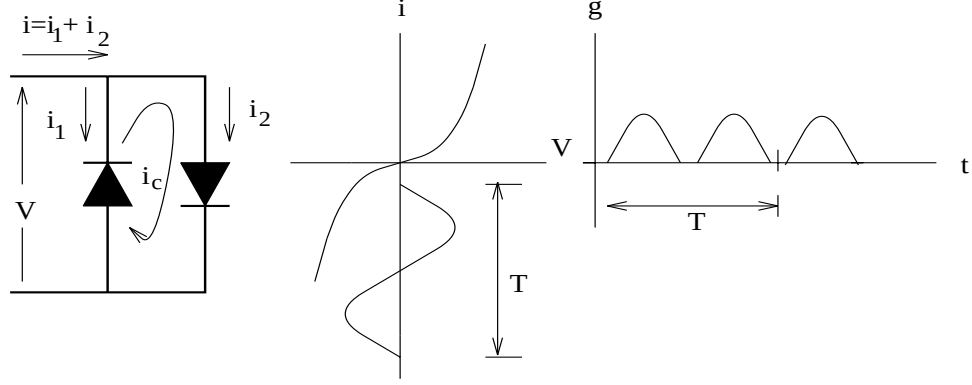


Figure 2.9: Anti-parallel Diode Mixer

individual conductances.

$$\begin{aligned}
 g &= g_1 + g_2 \\
 &= \alpha i_s (e^{\alpha V} + e^{-\alpha V}) \\
 &= 2\alpha i_s \cosh \alpha V
 \end{aligned} \tag{2.9}$$

From the above expression, it is clear that g has even symmetry with V , shown in Figure 2.9, and the number of conductance pulses per LO cycle in the anti-parallel diode circuit is twice that for a single diode circuit. When this diode pair is pumped with LO, it modulates the conductance of the diode and substituting $V = V_L \cos \omega_L t$ in equation 2.9, we get :

$$g = 2\alpha i_s \cosh(\alpha V_L \cos \omega_L t) \tag{2.10}$$

which, upon expansion, gives :

$$g = 2\alpha i_s [I_0(\alpha V_L) + 2I_2(\alpha V_L) \cos 2\omega_L t + 2I_4(\alpha V_L) \cos 4\omega_L t + \dots] \tag{2.11}$$

where $I_n(\alpha V_L)$ are modified Bessel functions of the second kind. It is clear from the above equations that the conductance terms consist of a dc term and the even harmonics of the LO frequency, ω_L . When the applied voltage is

$$V = V_L \cos \omega_L t + V_s \cos \omega_s t \quad (2.12)$$

the current will be :

$$i = g(V_L \cos \omega_L t + V_s \cos \omega_s t) \quad (2.13)$$

$$\begin{aligned} i = & A \cos \omega_L t + B \cos \omega_s t + C \cos 3\omega_L t + D \cos 5\omega_L t \\ & E \cos(2\omega_L + \omega_s)t + F \cos(2\omega_L - \omega_s)t + G \cos(4\omega_L + \omega_s)t \\ & H \cos(4\omega_L - \omega_s)t + \dots \end{aligned} \quad (2.14)$$

It can be seen from the above that the total current contains only frequency terms $m f_L \pm n f_s$, where $m + n$ is an odd integer, i.e., $m + n = 1, 3, 5, \dots$

There is one more component of current i_c , as can be seen in Figure 2.9. This circulating current arises because the Fourier expansion of individual currents i_1 and i_2 gives rise to components of current which are opposite in phase. Because of the opposite polarity, they cancel each other at the output terminal but circulate within the diode loop. The circulating current can be written as (from Figure 2.9) :

$$\begin{aligned} i_c &= \frac{(i_2 - i_1)}{2} \\ &= i_s [\cosh \alpha V - 1] \end{aligned} \quad (2.15)$$

Substituting

$$V = V_L \cos \omega_L t + V_s \cos \omega_s t$$

$$\begin{aligned}
 i_c &= i_s \left[1 + \frac{(V_L \cos \omega_L t + V_s \cos \omega_s t)^2}{2!} + \dots - 1 \right] \\
 &= \frac{i_s}{2} [V_L^2 \cos^2 \omega_L t + V_s^2 \cos^2 \omega_s t + 2V_L V_s \cos \omega_L t \cos \omega_s t + \dots] \\
 &= \frac{i_s}{2} \left\{ \frac{V_L^2 + V_s^2}{2} + \frac{V_L^2}{2} \cos 2\omega_L t + \frac{V_s^2}{2} \cos 2\omega_s t + V_L V_s [\cos(\omega_L - \omega_s)t \right. \\
 &\quad \left. + \cos(\omega_L + \omega_s)t] + \dots \right\} \tag{2.16}
 \end{aligned}$$

From the above equations it can be seen that the circulating current only contains frequencies $m f_L \pm n f_s$, where $m + n$ is an even integer.

Thus, the anti-parallel diode pair has the advantage of suppressing fundamental and other odd harmonic mixing products and also the even harmonics of the LO. However, it should be kept in mind that the degree of suppression degrades with the imbalance in the diode pair. It should also be noted that the degradation of receiver noise figure due to LO noise sidebands (which is the case in single diode mixers) is also reduced in even harmonic mixing ($m = \text{even}, n = 1$) with anti-parallel diodes. This is because the LO noise sidebands whose separation from the LO (f_L) equals IF (f_{IF}), generate IF noise which only circulates within the diode loop when they mix fundamentally with the LO; but second harmonic mixing of these noise sidebands with the virtual LO ($2f_L$) produces noise which are not within the IF amplifier pass

band (Figure 2.10).

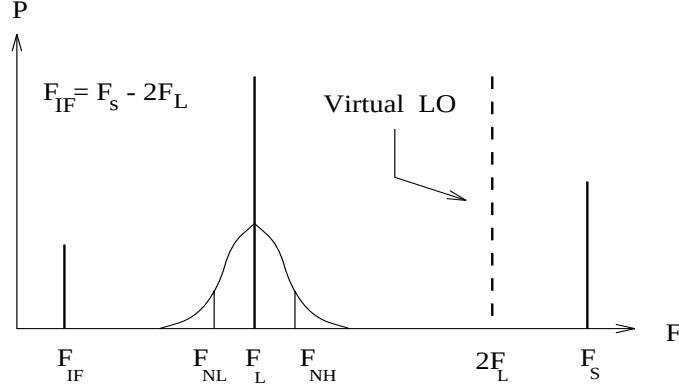


Figure 2.10: Noise sideband Mixing Products

Finally, the anti-parallel diode circuit has inherent self protection against large peak inverse voltage, because a reverse biased diode is always in parallel with a forward biased diode, which limits the reverse bias swing less than the breakdown voltage of the diodes.

2.4 Quasi-optical Mixers

At millimeter and sub-millimeter-wave frequencies, coupling of RF and LO signals to the diode is often done quasi-optically. This is because, at these short wavelengths, waveguide dimensions and tolerances are very difficult to realize in practice, reproducible electrical characteristics are a major hurdle, and waveguide losses increase with frequency. For quasi-optical mixers, the same coupling structure is used for LO and RF frequencies. This does not pose any problem in fundamental mixing as the frequency separating the RF and LO is small and hence, the bandwidth of the cou-

pling structure is not a major concern. However, in the case of subharmonic mixers, the RF is mixed with the second harmonic of the LO. So, the coupling structure should have at least one octave bandwidth. This section highlights the major aspects of broadband antennas for subharmonic mixing. Excellent references are available in the antenna literature about planar, frequency independent antennas and their radiation properties [8, 9, 10, 11].

It is observed that the impedance and radiation properties of antenna are dependent on the shape and dimensions expressed in wavelengths. If an antenna is scaled in wavelengths and transformed to another structure identical to the original, except for a possible rotation about the vertex, then its properties will be the same at both frequencies. The form of the antenna, in such cases, can be specified entirely by angles only and not by any other dimensions. This is one of the concepts behind frequency independent antennas.

The second concept is that if a structure becomes equal to itself when scaled by a factor $\frac{1}{\tau}$, it will have the same properties at frequencies f and τf . As a result, the antenna characteristics become a periodic function, with a period of $\log|\tau|$, of the logarithm of the frequency. This kind of structure is known as a *log-periodic* antenna. By making τ close to 1, the variation of an antenna properties over the band f and τf can be made very small. In practice, even with τ not very close to 1, good frequency independent antenna characteristics are observed.

A third kind of antenna, which is used in this research, is the self complementary pla-

nar conducting strip antenna which has a frequency independent input impedance. A self complementary strip antenna is obtained by interchanging the conducting and non-conducting planar surfaces. The self complementary condition requires that the complementary structure is identical to the original structure except for a rotation about the vertex point. It then follows from *Booker's* relation that the product of the input impedances of the original and the complementary structures is a constant and is independent of frequency, since the two impedances are identical.

Figure 2.11 shows a typical log-periodic antenna. The dimensions $R_{n+1}, R_n, r_{n+1}, r_n$ and the angles a and b determine the lowest and highest operating frequency and the impedance of the antenna.

When this kind of antenna is realized on a substrate (microstrip, slotline etc.), it is found that the radiation pattern is not symmetric, but is stronger into the dielectric than to free space [10]. As a result, coupling is primarily done from the substrate side. Unfortunately, antennas on dielectric slabs couple strongly to substrate modes. Radiation incident at an angle greater than the critical angle is reflected from the dielectric-free space boundary and is trapped as a surface wave. Hemispherical and hyper-hemispherical lenses are used to eliminate these substrate modes by making the substrate appear as a semi-infinite medium.

Kormanyos *et al.* [3], Lee *et al.* [4] and Siegel [12] have successfully used this planar broadband log-periodic antenna for both subharmonic and fundamental mixers. Be-

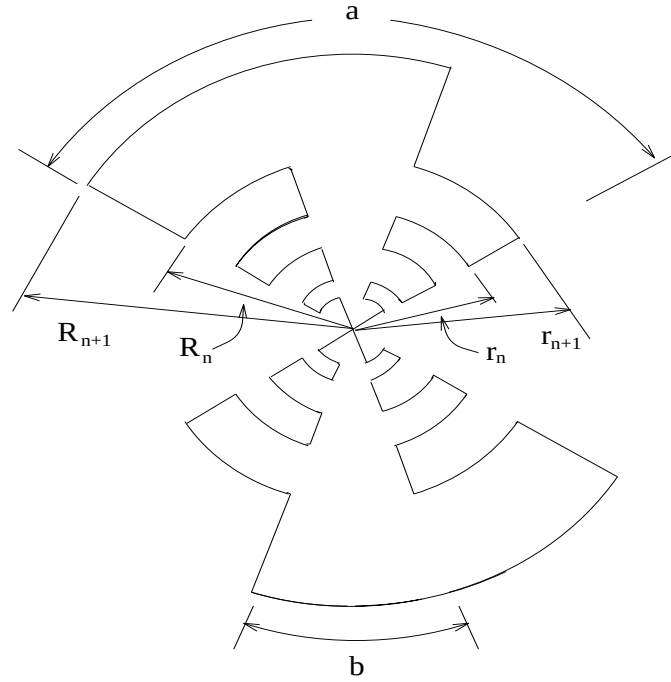


Figure 2.11: Log-Periodic Antenna

cause of the planar structure, the non-linear mixer element can easily be integrated with this log-periodic antenna, meaning the mixer may be fabricated monolithically.

Chapter 3

Design, Fabrication And Results

3.1 Antenna Design

Two different log-periodic antennas were designed, with the same design parameters but different coplanar transmission line feed structures. The first antenna, which is split in two halves to incorporate a 50Ω coplanar transmission line feed, is used for antenna pattern measurements. The second antenna has an RF matching network and an IF filtering structure in the coplanar transmission line integrated to the antenna. The antenna is a planar self complementary log-periodic antenna with 45° bow angle, $\sigma = 0.707$ and $\tau = 0.5$ and is designed to cover the 10 GHz to 40 GHz frequency range. These values of σ and τ yield a wide band antenna that maps onto itself every octave. There are a total of six teeth, three on each side of the log-periodic antenna. The following equations show the details of the design parameters σ and τ (refer to

Figure 2.11).

$$\frac{R_n}{R_{n+1}} = \tau = 0.5$$

$$\sigma = \sqrt{\tau} = 0.707 = \frac{r_{n+1}}{R_{n+1}}$$

The length of the largest tooth is calculated from the quarter-wave length at the lowest frequency of the design. The antenna was fabricated on a TMM 4¹ substrate, because it was decided to use an already existing 1 inch diameter quartz lens for the quasi-optical coupling. Also, TMM 4 has an ϵ_r of 4.5, which is close to ϵ_r (3.8) of quartz. The log-periodic antenna was placed at the back of the quartz lens to eliminate loss due to substrate modes. The dielectric lens also enhances the radiation pattern in the direction of the dielectric and increases the gain. The 1 inch diameter quartz hemispherical lens yields nearly a 1λ aperture at the lowest frequency of operation (10 GHz), and a 3λ aperture at 40 GHz.

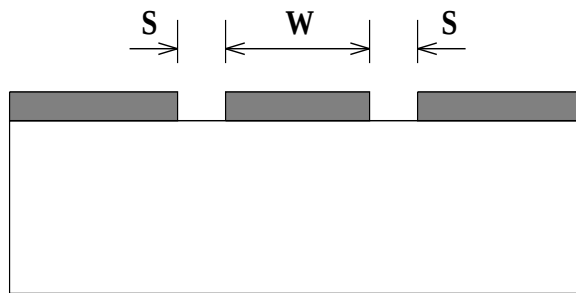


Figure 3.1: Coplanar Transmission Line

Coplanar transmission line parameters for a 50Ω characteristic impedance were cal-

¹TMM 4 substrates are manufactured by the Rogers Corporation

culated from the equations given by Ghione *et al.* [13]. The parameters are (as shown in Figure 3.1) $W = 0.85mm$ and $S = 0.087mm$. The antenna input impedance is independent of frequency and is given by :

$$Z_{ant} = \frac{189\Omega}{\sqrt{0.5(1 + \epsilon_r)}} = 114\Omega$$

189Ω is the impedance of any self complementary structure in free space and ϵ_r is the relative dielectric constant of the substrate. The antenna with 50Ω transmission line is shown in Figure 3.2.

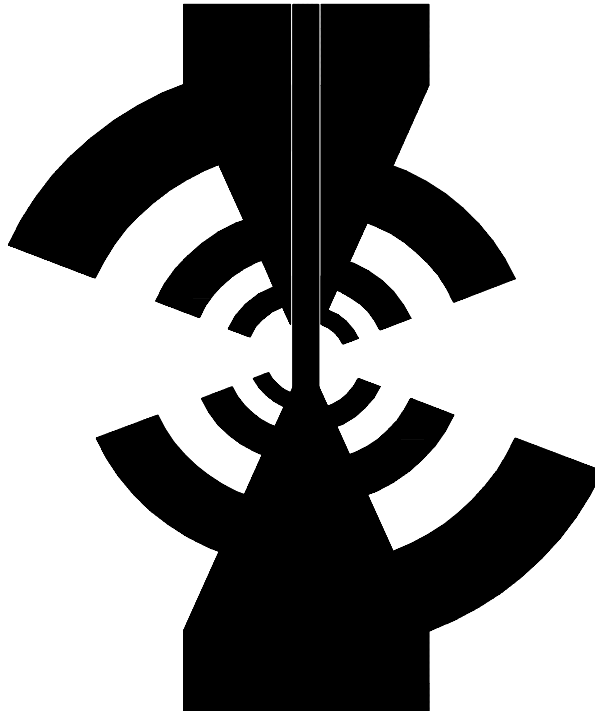


Figure 3.2: Log-Periodic Antenna with Coplanar Transmission Line

A positive mask was made from the above layout. Photolithography was done using

the positive mask and negative photoresist. A small mounting structure for the antenna was designed and fabricated. Since the width of the coplanar line is small, and also the maximum operating frequency of the antenna is 40 GHz, it was decided to use a K-connector (manufactured by Wiltron), instead of a standard SMA connector. The silicon lens was attached to the back of the substrate using g-wax, heating the substrate to about 150° C.

3.2 Antenna Measurement

The antenna with the 50Ω transmission line was used to measure the antenna radiation pattern. The measurement set-up is shown in Figure 3.3. One port of the HP8510 was connected to the log-periodic antenna to feed 15 GHz and 31.5 GHz signals and Ku and Ka band horns were connected to the other port to receive the radiated power. This set-up was also used to measure the insertion loss (S_{21}) of the system at 15 GHz and 31.5 GHz, from which the actual power coupled to the diodes at these frequencies for the mixer measurement were calculated. The radiation pattern in the E and H plane at 15 GHz and 31.5 GHz are shown in Figure 3.4 and Figure 3.5 respectively.

It is clear that at both the frequencies, the antenna pattern is not symmetric and the maximum gain point of the antenna is off-centered, by about 50° at 31.5 GHz.

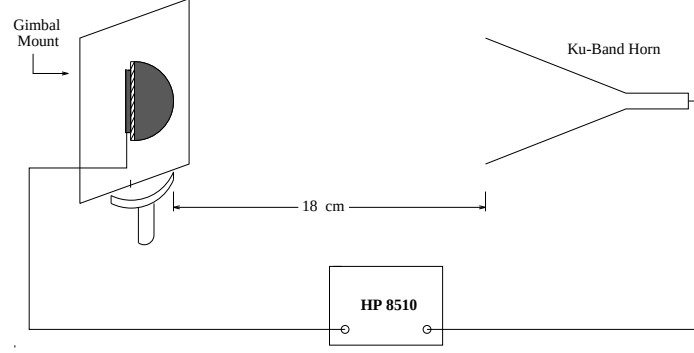


Figure 3.3: Antenna Measurement Set-up

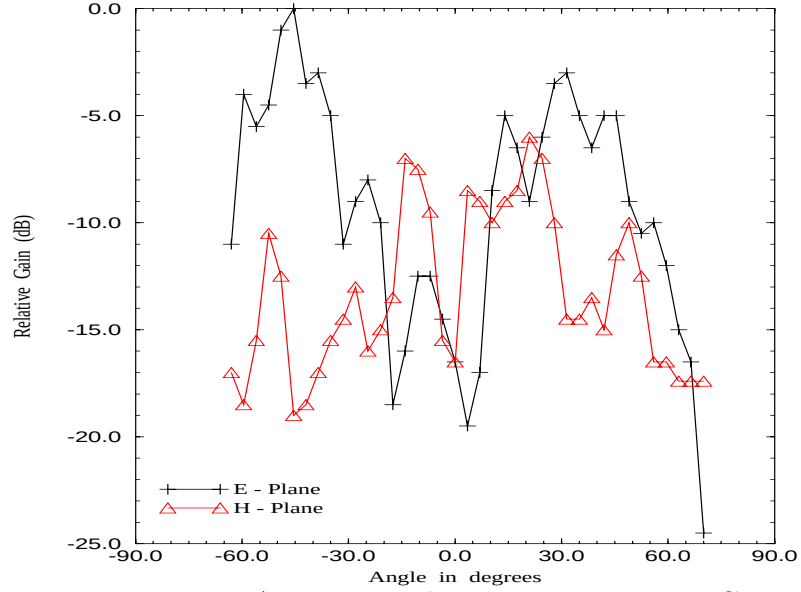


Figure 3.4: Antenna Radiation Pattern at 15 GHz

The reason for this off-centered radiation pattern appears to be the radiation from the coplanar transmission line. Radiation occurs in coplanar transmission lines on thick substrates because the waves on the line propagate faster than the waves in the dielectric [10]. There are three types of losses associated with a coplanar transmission line, namely, i) conductor loss α_c , ii) dielectric loss α_d and iii) radiation loss α_r . Using the equations given in Collin [14], the conductor and dielectric losses were calculated at 31.5 GHz and were found to be 0.08 dB and 0.2 dB respectively for 12.7mm length of line. The radiation loss was calculated from the equations given by Rutledge *et*

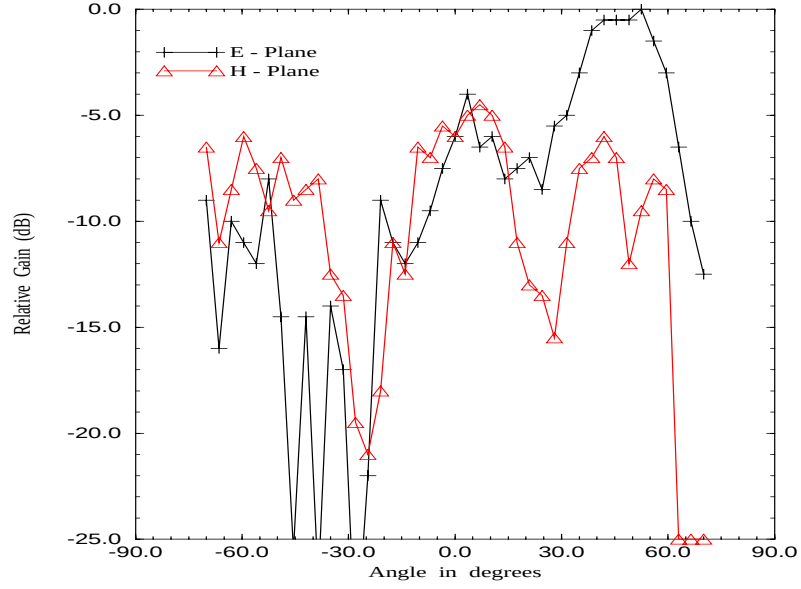


Figure 3.5: Antenna Radiation Pattern at 31.5 GHz

al. [10], and was found to be 1.3 dB, which undoubtedly is quite substantial.

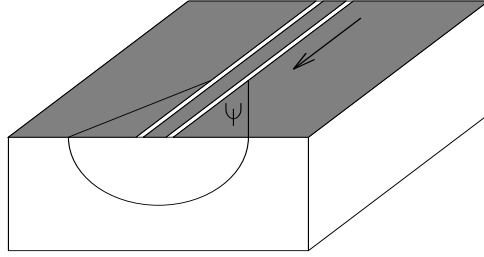


Figure 3.6: Radiation from coplanar Transmission line

As pointed out earlier, the radiation in a coplanar line is due to the unequal phase velocities of the waves on the line and in the dielectric. The radiation is similar to that of a leaky-wave antenna and is emitted in a semi-cone (as shown in Figure 3.6) near an angle ψ given by [10]

$$\cos \psi = \frac{k_z}{k_d}$$

where k_z is the guide propagation constant and k_d the dielectric propagation constant.

The above equation simplifies to

$$\cos \psi = \sqrt{\frac{\epsilon_{eff}}{\epsilon_r}}$$

where ϵ_{eff} is the effective dielectric constant of the medium ($\epsilon_{eff} = \frac{\epsilon_r+1}{2}$), and ϵ_r is the substrate dielectric constant. With ϵ_{eff} as 2.75 and ϵ_r as 4.5, ψ is found to be 38° . This is in reasonable agreement with the measured radiation patterns and indicates that the poor antenna performance is likely due to the presence of the coplanar line.

3.3 Mixer Design

The non-linear harmonic balance analysis program of Hewlett Packard's Microwave Design System (MDS) was used to design and analyze the subharmonic mixer. It was decided to use GaAs Schottky-diode chips SC2T3, designed and developed at the University of Virginia, in anti-parallel configuration, for the mixer design. The diode parameters are shown in Table 3.1.

Table 3.1: Schottky Diode Parameters

Epi Doping	Epi Thickness	Buffer Doping	Buffer Thickness	Anode Dia.	Finger Length	Chip Dimension
2×10^{17}	800Å	5×10^{18}	$5 \mu m$	$2 \mu m$	$50 \mu m$	5x15x1.5 mils

Both the diodes have a barrier height close to 0.85V. The DC parameters (shown

Table 3.2: Diode DC Parameters

DC Parameter	R_s	η	I_s	$\phi_{barrier}$	C_j	C_{total}
Diode 1	4.3 Ω	1.17	1.2x10 ⁻¹⁶ A	0.85 V	5.5 fF	14 fF
Diode 2	4.2 Ω	1.19	2.4x10 ⁻¹⁶ A	0.83 V	5.5 fF	14 fF

in Table 3.2), except the capacitances, are obtained from a least-square fitting of the experimental I-V curves. The harmonic balance analysis was performed with a 15 GHz LO, 31.5 GHz RF and 1.5 GHz IF. In the analysis, an ideal low-pass filter was used for IF filtering, and it was replaced by a transmission line stepped-impedance filter in the actual circuit. A single transmission line section was used to match the diode impedance to the 114 Ω antenna impedance.

The MDS simulation set-up is shown in Figure 3.7. Since the diodes are to be biased separately, a split in the antenna was needed to isolate the ground plane and the bias point. In a coplanar transmission line, the current flows through that part of the ground strip which is close to the center conductor. Also, by the time the current reaches the connector end, the high frequency components of the current will be filtered out because of the coplanar transmission line filter. Keeping these facts in mind, the split in the antenna was made near the connector end of the antenna, as shown in Figure 3.8. A 680pF (0.15 Ω at 1.5 GHz) chip capacitor was soldered at the split to have RF continuity. A flip-chip mounting technique is used to solder the single devices down to the antenna using low temperature Indium solder. Again, the quartz lens was attached to the back of the antenna using g-wax.

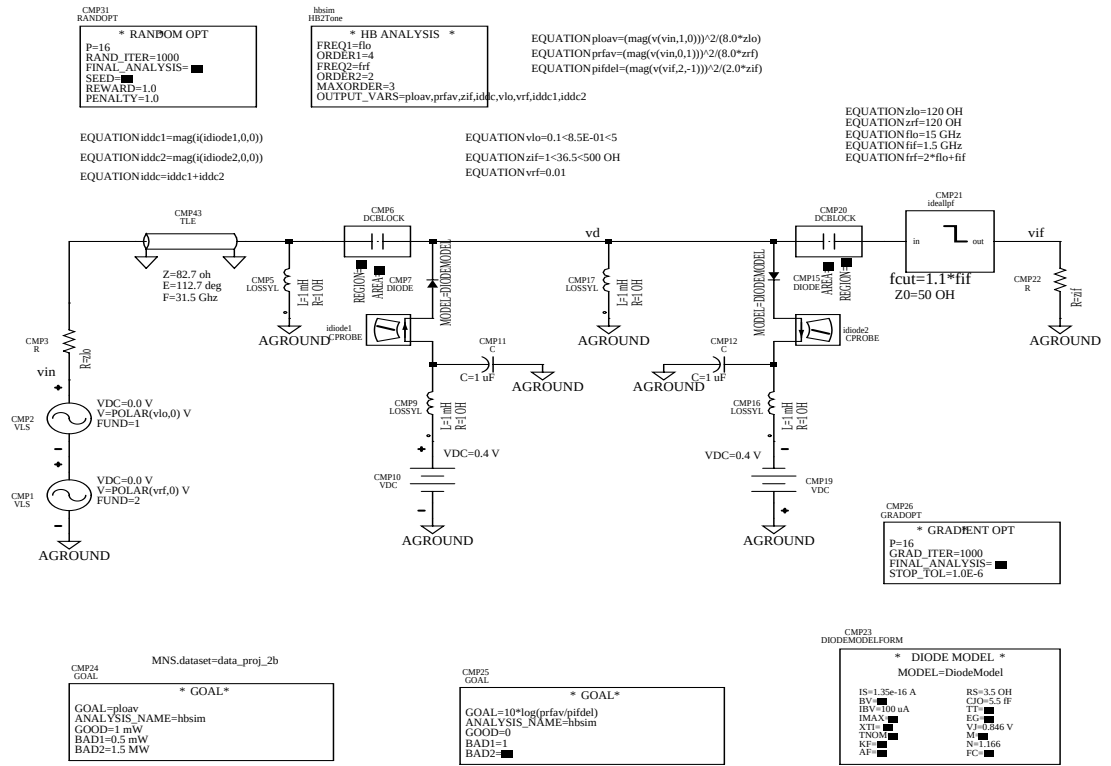


Figure 3.7: MDS Simulation Circuit Page



Figure 3.8: Antenna with IF and RF Matching Circuit

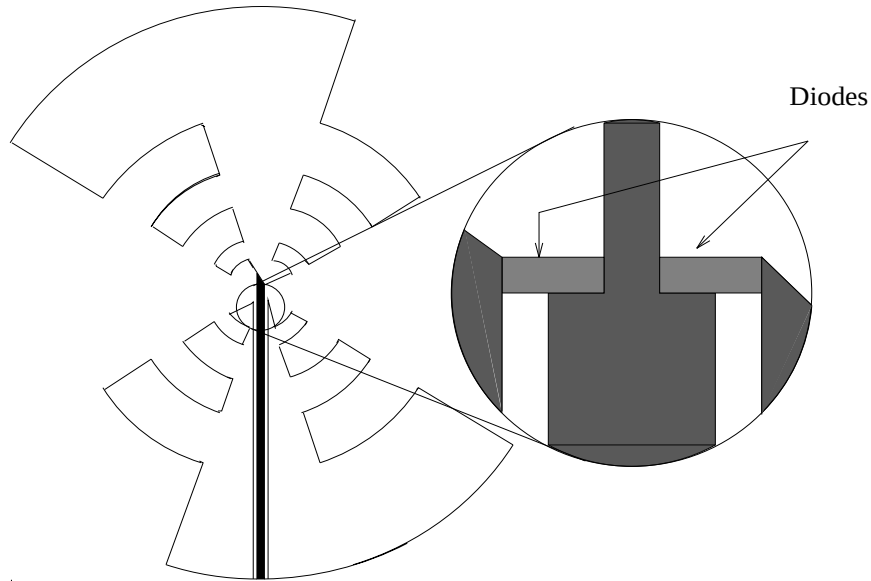


Figure 3.9: Details of the Diode Mounting on the Antenna

Figure 3.9 shows the position of the diodes soldered on the planar antenna.

3.4 Mixer Measurement

The mixer measurement set-up is shown in Figure 3.10. It should be noted that in this measurement set-up, a diplexer is not used to couple LO and RF to the diodes. The reason is non-availability of a suitable dichroic plate and the poor performance of the antenna. Instead, the diodes were pumped by the LO from the back side of the antenna while the RF radiation was coupled from the front side. The power coupled to the diodes at the RF and the LO were measured using the antenna fed by a 50Ω line. The system gain was corrected by including the conductor, dielectric and radiation losses.

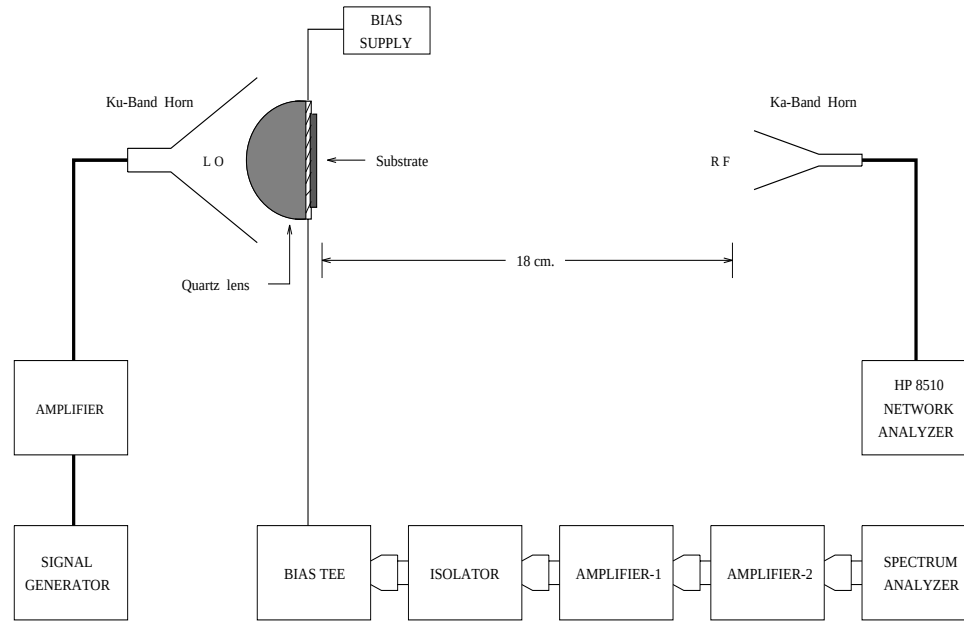


Figure 3.10: Mixer Measurement Set-up

The maximum LO power coupled to the diodes was measured to be -5.5 dBm . With this LO power and 0.5 Volt of biasing, the conversion loss was measured to be 20 dB . This conversion loss is quite high. To check whether the mixer was pumped with enough LO power, the IF power was measured as a function of LO power.

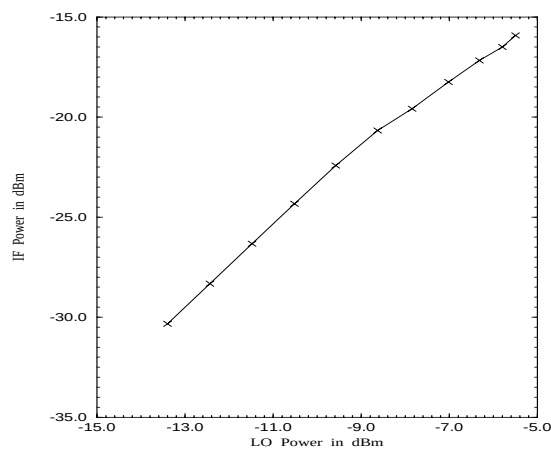


Figure 3.11: IF output Vs. LO Power

It can be seen from Figure 3.11 that the output power increases linearly with the

local oscillator power, indicating that the diodes are not pumped with the optimum local oscillator power. However, the 20 *dB* conversion loss is consistent with the MDS simulation, which predicts a conversion loss of 18 *dB* with -5.5 *dBm* LO power and 0.5 Volt of biasing.

A broadband (6-18 GHz)amplifier was used to get more LO power at 15 GHz. The Table 3.3 shows the mixer performance.

Table 3.3: Mixer Performance

Biasing	Conversion Loss	LO Power
± 0.4 V	10.0 dB	1.3 mW
± 0.2 V	10.5 dB	2.1 mW

In summary, the mixer performance is as predicted by the MDS simulation. The quasi-optical coupling structure is not performing as expected and needs improvement.

Chapter 4

Diagonal Horn Antenna

This chapter describes the design of a feed horn for a 585 GHz SIS receiver system. Commonly used horns, like the corrugated pyramidal and conical horns, radiate a near-perfect *Gaussian* beam, but are very difficult to realize in practice because of the small dimensions and tolerances involved at sub-millimeter-wave frequencies. Single mode horns, like the conical (TE_{11}) or pyramidal (TE_{10}) horns, exhibit a lack of symmetry in the cardinal plane of the radiation pattern, which makes them less suitable for launching a *Gaussian* beam [15]. The pyramidal horn also suffers from what is known as *astigmatism* i.e., the phase centers for the E - and H - planes do not coincide [16]. The solution to this is to introduce appropriate additional modes into the horn which will propagate to the horn aperture along with the dominant mode. In most practical cases, only one additional mode is sufficient. Such horns are called dual-mode horns, and the diagonal horn is the simplest form of a dual-mode horn.

In a diagonal horn, two spatially orthogonal modes TE_{10} and TE_{01} are excited with

equal amplitude and phase in a square waveguide which flares into a pyramidal horn (Figure 4.1).

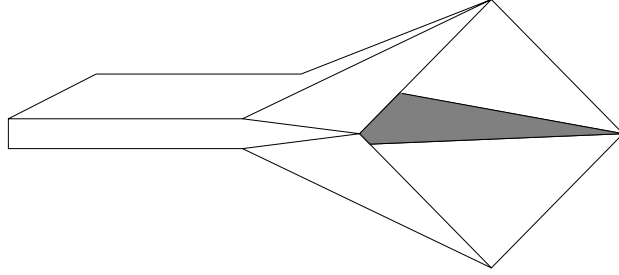


Figure 4.1: Diagonal Horn

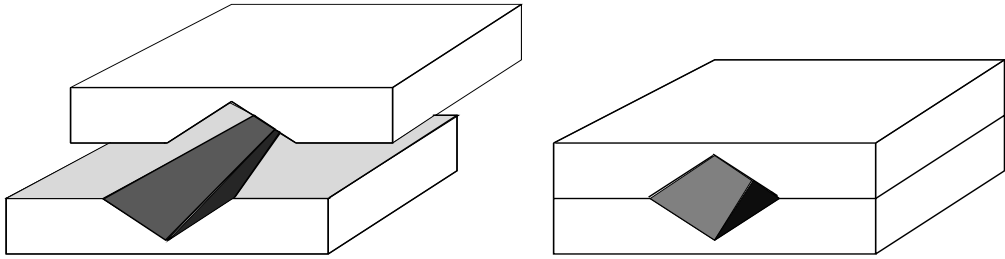


Figure 4.2: Split-Block Technique

One of the major advantages of the diagonal horn is the ease with which it can be fabricated. When using waveguide technology at millimeter and sub-millimeter wavelengths, it is quite common to design the mixer using split block technique. The block is fabricated in two pieces and the waveguide is formed by milling a square cross-section channel in both halves. The losses are small for TE_{10} mode since the split occurs along the center of the broad walls of the waveguide. It is clear from Figure 4.2 that the diagonal horns are well-suited for split block technology.

Figure 4.3(a) shows the two equal, co-existing modes (TE_{10} and TE_{01}) in the horn

and Figure 4.3(b) shows the resulting electric field pattern at any particular cross-section and at a particular instant of time [17].

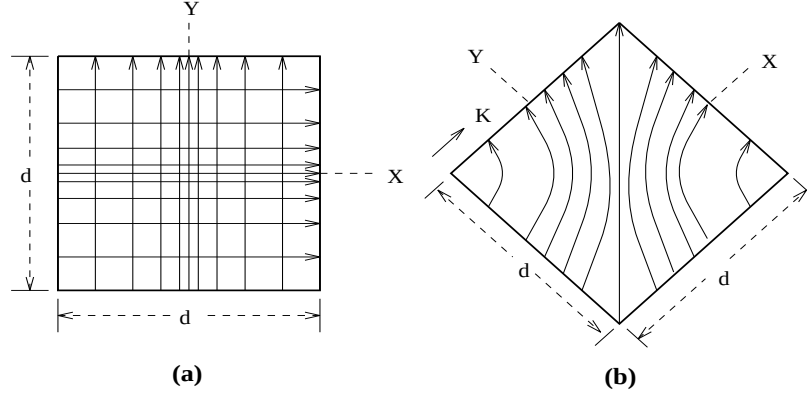


Figure 4.3: Electric field configuration inside square horn

From the electric field vector, it is clear that only E_x exists for one mode and only E_y exists for the orthogonal mode. The spatial variation of E_x and E_y is given by

$$\begin{aligned} E_x &= \cos \frac{\pi y}{d} \\ E_y &= \cos \frac{\pi x}{d} \end{aligned} \quad (4.1)$$

(The common propagating wave function $e^{j(\omega t - \beta z)}$ is not shown here.)

Thus, at any point within the cross-section of Figure 4.3(a), the resultant electric field is

$$E = \sqrt{\cos^2\left(\frac{\pi x}{d}\right) + \cos^2\left(\frac{\pi y}{d}\right)} \quad (4.2)$$

and its direction is inclined at an angle α to the x-axis, where

$$\tan \alpha = \frac{E_y}{E_x}$$

$$= \frac{\cos(\frac{\pi x}{d})}{\cos(\frac{\pi y}{d})} \quad (4.3)$$

The differential equation for the lines of electric force is

$$\cos(\frac{\pi y}{d}).dy = \cos(\frac{\pi x}{d}).dx \quad (4.4)$$

The equation of the lines of force is obtained by integrating Equation (4.4) and is given by

$$\begin{aligned} \sin(\frac{\pi y}{d}) &= \sin(\frac{\pi x}{d}) + K \\ (-2 \leq K \leq 2) \end{aligned}$$

Where K is a constant for any line of force [17]. The resultant field pattern, as shown in Figure 4.3(b), resembles the dominant TE_{11} mode in circular waveguide, which suggests a circular transition for launching such a wave. However, Johansson *et al.* [15] have shown that the transition is not critical and a *direct* transition from rectangular waveguide works well for most purposes (Figure 4.4). They have also shown that the aperture field of the horn has the desired symmetry property for such a transition.

The horn for this project was designed from the *Gaussian* mode model equations given by Johansson *et al.* [15]. The beam parameters are given by

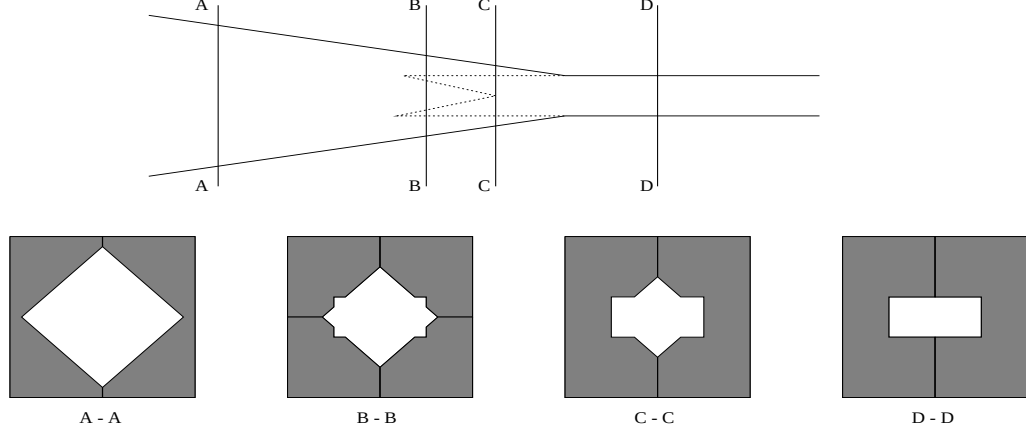


Figure 4.4: Transition from rectangular waveguide to diagonal horn

$$\begin{aligned}
 w(z) &= w_0 \sqrt{1 + \left(\frac{z}{z_c}\right)^2} \\
 R(z) &= z \left[1 + \left(\frac{z_c}{z}\right)^2\right] \\
 z_c &= \frac{\pi w_0^2}{\lambda} \\
 \Phi(z) &= \arctan \frac{z}{z_c}
 \end{aligned} \tag{4.5}$$

where w denotes the beam waist radius, R the phase radius of curvature, z_c the confocal distance, and Φ the so-called phase slip. It should be noted that w and R are common to all modes but the phase slip Φ is progressively multiplied for higher order modes. The geometry of the equivalent *Gaussian* beam is shown in Figure 4.5. Johansson *et al.* [15] have shown that for maximum *Gaussian* beam coupling ($\eta_{Gauss} \approx 0.843025$), w_A/a should be equal to 0.863191.

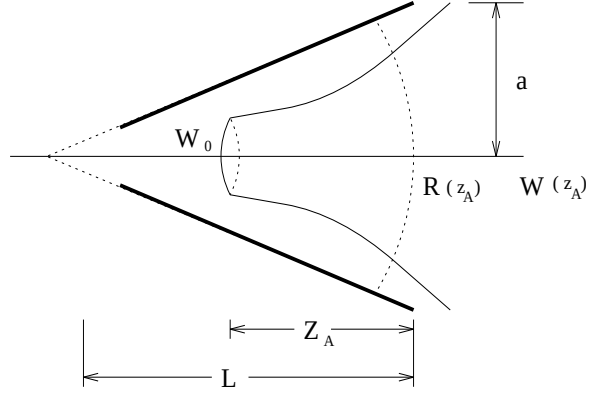


Figure 4.5: Geometry of the equivalent *Gaussian* beam

The equivalent *Gaussian* beam parameters are calculated from the equations given below.

$$\begin{aligned}
 w_A &= w_0 \sqrt{1 + \left(\frac{z_A}{z_c}\right)^2} = \kappa a \\
 R_A &= z_A \left[1 + \left(\frac{z_c}{z_A}\right)^2\right] = L \\
 z_c &= \frac{\pi w_0^2}{\lambda} \\
 \Phi_A &= \arctan \frac{z_A}{z_c} \\
 &= \arctan \kappa^2 M \\
 M &= \frac{\pi a^2}{\lambda L}
 \end{aligned} \tag{4.6}$$

By algebraic manipulation of the above equations, we get,

$$\begin{aligned}
w_0 &= \frac{\kappa a}{\sqrt{1 + \tan^2 \Phi_A}} \\
z_A &= \frac{L}{1 + \cos^2 \Phi_A}
\end{aligned} \tag{4.7}$$

And the above two parameters (w_0 and z_A) are mainly needed for the design of the horn. Since the horn will be used inside a cryogenic dewar, some restriction was imposed on the maximum length that the horn can have. Equations (4.6) and (4.7) show clearly that the antenna length is a function of the beam waist of the *Gaussian* beam. Since some external optics can be used to change the beam parameters, the beam waist was decided to be 0.7 mm and that gave an antenna length of 13.13 mm (517 mils). Once the length of the antenna was obtained, the aperture length a was calculated from the above equations. The final antenna drawing with the flange is shown in Figure 4.6.

A special cutting tool was needed to get the flare of the horn right and keep the aperture a perfect square. The horn was fabricated with gold plated brass in the NRAO¹ workshop. The horn has not been tested yet because of the non-availability of a detector on a WR-2 waveguide mount. Jeffrey Hesler of the SDL² has designed a Schottky-diode receiver at 585 GHz on waveguide mount and the horn will be

¹National Radio Astronomy Observatory, Charlottesville, Virginia.

²Semiconductor Device Lab. Dept. of Electrical Engineering, UVa

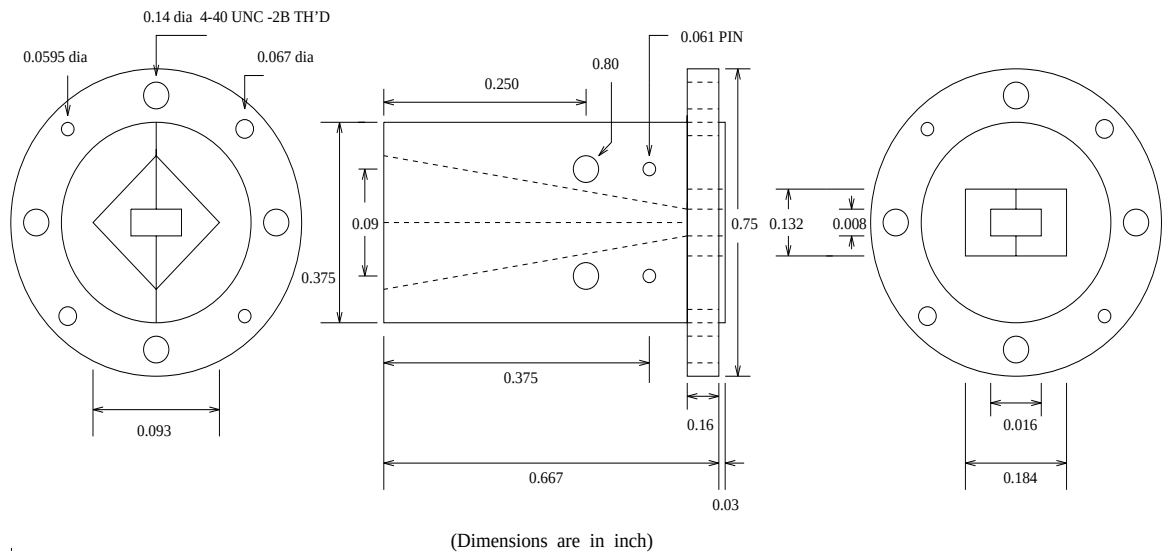


Figure 4.6: Diagonal horn antenna with flange

characterized when that system becomes available.

Chapter 5

Conclusions

The two main objectives of this research were, i) to investigate the log-periodic antenna with coplanar transmission line feed, and ii) to demonstrate that a coplanar IF filtering network and RF impedance transformer can be incorporated in the antenna structure to improve the mixer performance.

The antenna radiation pattern shows that the coplanar transmission line has grossly affected the radiation pattern. The reason for this is the discontinuity of the RF currents in the antenna due to the split. The current in a log-periodic antenna flows through the edges of the teeth, as shown in Figure 5.1. Due to the presence of the coplanar transmission line, RF currents can no longer flow all the way to the other side. A possible solution to this would be to use an overlay capacitor which will allow RF currents to flow and still permit the diodes to be biased separately.

The mixer result shows that the LO power requirement goes down almost by a fac-

tor of two for change of bias from $\pm 0.2\text{V}$ to $\pm 0.4\text{V}$. The conversion loss also shows that the mixer performance is close to that predicted by the MDS harmonic balance analysis.

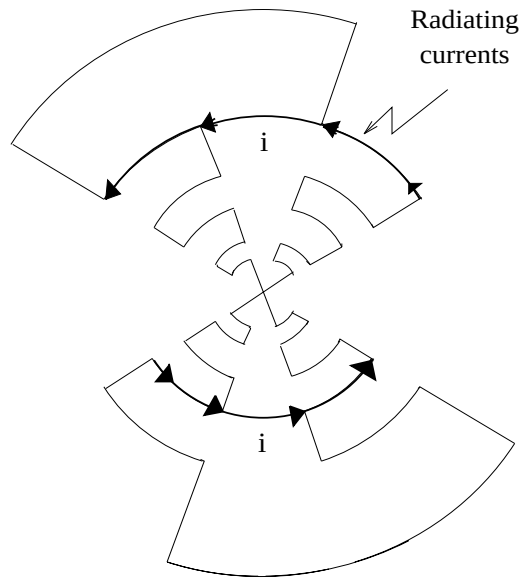


Figure 5.1: Current Flow in Log-Periodic Antenna

At millimeter and submillimeter-wave lengths the antenna and the mixer elements should be fabricated monolithically. Monolithic structure will also allow to incorporate overly capacitor for RF continuity. Further study in characterizing planar integrated antennas that may accommodate separate biasing should be carried out to find better quasi-optical coupling structures for coupling free space radiation to the planar diodes.

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