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**Implementation Of A Microstrip Based “Magic-T”
Hybrid Coupler In The Design Of A
Single-Balanced Diode Mixer**

Carlos Almeida, B. Eng.

CARLETON UNIVERSITY

Ottawa, Canada

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acceptance of the thesis

“Implementation Of A Microstrip Based Magic-T
Hybrid Coupler In The Design Of A
Single-Balanced Diode Mixer”

submitted by

Carlos Almeida, B. Eng.

in partial fulfillment of the requirements for
the degree of Master of Engineering

Thesis Supervisor

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ABSTRACT

Mixer circuits find use in numerous applications and come in a variety of topologies. They consist of one or more nonlinear devices which are typically diodes or field effect transistors. Many circuits consist of an even number of nonlinear devices arranged in a symmetrical manner referred to as balanced circuits. Balanced circuits employing two nonlinear devices generally make use of hybrid couplers. These structures serve to inject RF and LO signals into the mixer circuit such that their sum and difference appear across the nonlinear mixing elements.

One type of hybrid coupler commonly found in the microwave field is the waveguide "Magic-T". Circuits based on waveguide technology are not characteristically small, compact, or light weight. The use of the microstrip medium has made possible the development of many high performance, miniature, light weight, and planar circuits which have found use in many areas. With the ever increasing attention being given to microstrip based integrated circuits it would be desirable to implement this structure in microstrip form. This provided the motivation for researchers to eventually develop a microstrip prototype circuit that illustrated the properties of the "Magic-T".

The microstrip-based "Magic-T" is a multilayered structure. The upper layer implements a T-branch circuit and the lower layer consists of a feedline terminated in an open-circuit. A common ground plane separates both layers. By utilizing an aperture in the ground plane, the bottom and top metallization can be used to realize an aperture-coupled parallel line circuit. The combination of T-branch and aperture-coupled lines is used to realize the properties associated with the "Magic-T".

This particular work investigates the development of a microstrip-based "Magic-T" that effectively operates as a 180-degree hybrid coupler. This structure is then integrated with other components to realize a working single-balanced diode mixer.

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CHAPTER 1: INTRODUCTION

1.1 General

At present, mixer circuits of many different topologies and complexities are used in a multitude of applications. In each one, the mixer is used to shift the frequency of an incoming signal either upwards or downwards to an intermediate frequency where the signal can be more readily used. Typical applications involve using mixers in transmitters and receivers. However, over the years an increasing number of applications require the use of receivers and consequently more attention has been directed towards the implementation of mixers in receivers.

Major Edwin Armstrong is generally accepted to have been the first to develop the concept of mixing and use it. In the 1930's he concentrated his efforts in the field of radio communications with particular emphasis on frequency modulation [32], [33]. During this time he developed a receiver that used vacuum tube mixers to shift incoming signals to an intermediate frequency where they could be more easily amplified with good selectivity, high gain, and low noise, before being finally demodulated. His superheterodyne receiver, as it came to be known, has found wide spread use and continues to form the basis of many receiver designs today.

Figure 1.1 provides one example of how mixers can be used in a typical receiver. In this case two mixers are used to facilitate the implementation of a dual-conversion UHF/VHF communications receiver. In this receiver the desired input RF signal is first filtered from all others and its strength increased to overcome the noise in the later stages. The first mixer stage uses the first LO to shift the signal's frequency to a much higher intermediate frequency. At this higher frequency, the image frequency is moved well away from the RF, thus allowing it to be rejected effectively by the following filter. The second mixer stage uses the second LO to shift the first intermediate frequency downwards to a much lower frequency where the demodulator works better. At this lower fre-

quency narrower bandwidths are more easily realized and thus better selectivity can be obtained through filtering. In this receiver two mixer stages were used to address two specific concerns. Other receivers might use mixers differently to meet other considerations.

Figure 1.1 Dual conversion VHF/UHF communications receiver [11]

Besides finding use in the modulation and demodulation of FM signals, vacuum tube mixers were also used as the first stage of early UHF/VHF receivers. However, as radar systems began operating at microwave frequencies during World War II, the sensitivity (low noise) of vacuum tube mixers was unsuitable for operation in radar receivers. Sensitivity is a limiting factor in determining a radar's maximum detection range, hence mixers with improved sensitivity were needed. Research in the 1940's and 1950's turned to diode mixers. As a result, diode mixer theory has become well established and diode mixers are the highest frequency low-noise components available. With the development of low-noise amplifiers and their use in receivers, the emphasis on low noise mixers is not as paramount as it once was. However, diode mixers still offer supe-

rior performance over LNA's in receivers that operate above 100 GHz. In applications in the terahertz range such as in the fields of plasma diagnostics, radio astronomy, and radar imaging diode mixers are the only option available.

Over the years other nonlinear devices have been developed and used in the design of mixer circuits. Bipolar transistors are used occasionally below 1 GHz but are uncommon at microwave frequencies. However with the development of heterojunction bipolar transistors a resurgence in their use could be forthcoming. Another common device found in mixer circuits, especially in the millimeter wave range, are field-effect transistors (FETs). In particular, metal-semiconductor field-effect transistors (MESFETs) and high-electron-mobility transistors (HEMTs) have paved the way for active mixer circuits in the microwave and millimeter frequency ranges. Compared to diodes they offer conversion gain instead of loss, lower noise, lower levels of intermodulation distortion, and require less LO power. Dual-gate FETs are also available for mixers. These devices offer inherent LO-RF isolation since the LO and RF signals can be applied to different gates. However, noise figure and conversion gain are not as good as those of single FET mixers.

Presently many mixers are implemented as monolithic circuits. FETs are very easily integrated into monolithic circuits which are increasingly finding use in more and more applications. Diodes can be fabricated on monolithic circuits but the existing technology does not lead to as good quality diodes as found in discrete components. Monolithic circuits offer advantages of small size, low cost for fabrication in large quantities, and uniform performance. However, development costs are still high and the technology is appropriate in a limited range of applications. With great attention currently focused on their development it is expected that fabrication costs will decrease in the future, and the use of monolithic FET mixers will become more widespread.

Single FETs or diodes can be used as mixers. In fact most high performance millimeter wave mixers are single-device designs. However, common mixers

employ more than one device and are classified as either singly balanced or doubly balanced mixers. Single-balanced mixers consist of two mixing devices connected together through a hybrid coupler. Doubly-balanced mixers usually consist of four or more untuned mixing devices that are interconnected by multiple hybrids, transformers, or baluns. They are too complicated for individual tuning of devices so they have higher conversion loss than single devices or singly balanced mixers. However, lack of tuning elements combined with broadband hybrids often give these mixers very wide bandwidths. Doubly balanced mixers are more difficult to realize with FETs than with diodes. The parasitics associated with the interconnection of several FETs limit the mixer performance. Balanced FET mixers are somewhat more practical in MMIC's but their high frequency performance is often unspectacular. Improvements in high frequency performance is expected in the future as device and circuit technologies mature.

Fundamental to the operation of many balanced mixer circuits are the hybrid couplers. They provide the means to combine the LO and RF input signals into the mixer. Hybrid couplers are four port passive devices (2 inputs and 2 outputs) and come in two varieties, the 180-degree hybrid coupler and the 90-degree coupler. In a 180-degree coupler, the input signals divide evenly and appear as a summation across one output arm and as difference across the other. In a 90-degree hybrid the input signals divide evenly and appear across each output arm 90-degrees out of phase from one another. Figure 1.2 depicts a common 180-degree hybrid implemented in waveguide. This is commonly referred to as a "Magic-T".

With more attention focused on the development of microstrip and monolithic microwave integrated circuits, it is not surprising that many waveguide components and devices have been converted to these newer technologies. The waveguide "Magic-T" is one such device that could lend itself to many applications in the microwave field if a microstrip-based counterpart could be developed. The aperture-coupled microstrip structure shown in its constituent parts in Figure 1.3 is one microstrip representation that has

been proposed to function in the same manner as the waveguide “Magic-T”.

Figure 1.2 Waveguide “Magic-T”

Investigation into an aperture-coupled microstrip “Magic-T” was first conducted by Katsube [12]. His objective was to demonstrate in theory and experimentally that an aperture-coupled microstrip structure could exhibit the same properties typically associated with the waveguide “Magic-T”. The microstrip structure of Figure 1.3 is the microstrip implementation of the “Magic-T” that will be used in the design of the single-balanced diode mixer. It consists of two back-to-back substrates separated by a ground plane. The upper substrate consists of a T-branch circuit. The lower layer is a feed line which is essentially a microstrip transmission line terminated in an open circuit. The slot in the ground plane allows electromagnetic energy to travel from the lower substrate to the upper substrate while at the same time allowing very little to flow in the opposite direction.

Figure 1.3 Microstrip “Magic-T” realization

1.2 Thesis Objectives

The first objective is to design and fabricate a working 180-degree hybrid coupler using a “Magic-T” implementation in microstrip. The T-branch and the aperture-coupled line sections will operate at 18.5 GHz and 20 GHz respectively. Both components will be designed with minimum insertion loss and return loss. Once the “Magic-T” has been realized its performance must be evaluated by measuring various parameters such as isolation, insertion loss, return loss, and phase difference between ports. The second objective is to design and fabricate a single-balanced diode mixer which implements the microstrip “Magic-T” coupler. The single-balanced diode mixer will operate as a

downconverter for use in a communications system. The RF signal, LO signal, and IF signal will be at 20 GHz, 18.5 GHz, and 1.5 GHz respectively. The mixer will be designed for minimum conversion loss over as wide an IF bandwidth as possible. Measurement of the final circuit will serve to grade the overall performance of the mixer circuit.

1.3 Thesis Outline

In Chapter 2, we will first look at the concept of mixing and what the requirements are at the device level for effective mixing. We will then look at the Schottky diode and examine how it can be used to implement mixing. This will require understanding its physical structure and electrical properties such as I/V characteristic, junction resistance, junction capacitance, and series resistance and how they impact on its operating performance. A model that represents the diode under both large and small-signal operating conditions will also be necessary. A diode model will be presented and the required measurements for parameter extraction and the modelling technique will be discussed. With this in mind a commercially available Schottky-barrier diode will be chosen as the mixing element, measured, and modelled for use in the final mixer circuit.

Chapter 3 will focus on the proposed 180-degree hybrid coupler referred to as the microstrip “Magic-T”. First the ideal properties of the hybrid coupler will be reviewed and the expected deviations from ideality will be discussed. Next the waveguide “Magic-T” illustrated in Figure 1.2 will be broken up into its constituent components and microstrip counterparts for each will be presented. The operating theory for these microstrip structures, the T-branch and aperture-coupled microstrip lines, will be examined. Both of these structures will then be designed using a commercially available microwave CAD package, LibraTM [13], and an electromagnetic simulator based on the transmission line method (TLM) [14]. Once designed both structures will be fabricated and assembled to realize the microstrip “Magic-T”. Finally, this 180-degree hybrid coupler will be mea-

sured and its performance results presented.

The filter circuits needed for the single-balanced mixer circuit will be examined in Chapter 4. First the general theory behind filtering circuits will be presented. Then a design technique for realizing the required filters will be discussed. Once designed, the filter structures will be simulated and optimized in LibraTM. Finally, the filters will be fabricated and their performance measured.

Chapter 5 will look into the single-balanced mixer circuit. A general introduction to mixer circuits followed by the operating theory behind single-balanced diode mixers will be outlined. The basis and method of implementing harmonic balance techniques in the analysis of mixer circuits will also be covered. The function and design of components necessary for the realization of the single-balanced diode mixer, such as matching circuitry, DC-IF returns, etc. will also be discussed. Next, the performance parameters typically associated with mixer circuits and their measurement will be reviewed. Finally the measured results of the single-balanced mixer circuits will be presented.

The final chapter will summarize and discuss the results obtained for the microstrip “Magic-T” coupler and the single-balanced diode mixer. It will also evaluate how well the objectives were met, discuss any shortcomings, possible alternatives and recommendations, as well as possibilities for future work.

CHAPTER 2: THE SCHOTTKY-BARRIER DIODE

2.1 The Mixing Operation

In theory a mixer operates similarly to a multiplier [11]. Figure 2.1 shows an ideal multiplier with two different signals applied to its input ports. An amplitude modulated RF (radio frequency) sinusoidal signal, $A(t) \cos(\omega_{RF}t)$, is applied to one port and an unmodulated LO (local oscillator) sinusoidal signal, $\cos(\omega_{LO}t)$, is applied to the other. The same analysis would apply if other forms of RF modulation were used.

Figure 2.1 Typical multiplier circuit [11]

At the output of the multiplier the signal would be $A(t) \cos(\omega_{RF}t) \cos(\omega_{LO}t)$. Making use of the following trigonometric identity

$$\cos(\omega_{RF}t) \cos(\omega_{LO}t) = \frac{[\cos(\omega_{RF} + \omega_{LO})t + \cos(\omega_{RF} - \omega_{LO})t]}{2} \quad (2.1)$$

the output signal can be expressed as

$$A(t) \frac{[\cos(\omega_{RF} + \omega_{LO})t + \cos(\omega_{RF} - \omega_{LO})t]}{2} \quad (2.2)$$

Equation (2.2) shows that components consisting of the sum and difference of RF and LO frequencies exist at the output of the multiplier. Either the sum or differ-

ence component can then be selected by means of a filter. Mixing is not only realized by an ideal multiplier, but nonlinear devices can also perform the multiplying function. However, nonlinear devices are not ideal multipliers. Their use in multiplication results in the generation of LO harmonics and in mixing products other than the desired ones. The desired output frequency component must be filtered out from all the rest.

2.2 The Diode As A Mixing Element

The diode is one particular nonlinear device that has been readily used in mixers. By utilizing the diode's nonlinearity we can implement the multiplying or the mixing function. Figure 2.2 illustrates how the diode's nonlinear I-V transfer function produces an output current, (I_{OUT}) , upon application of an input signal, $(V_p \cos(\omega t))$. The nonlinear relationship between current (I) and junction voltage (V_j) across the diode can be described ideally by the following exponential relationship

$$I = I_S \left(e^{\beta V_j} - 1 \right) \quad (2.3)$$

where I_S is the output current at $V_j = 0$ and β is a temperature factor.

Examining the relationship between current and voltage more closely it becomes evident that the current $I(V)$ is a function differentiable over any arbitrary voltage interval $[a, b]$. As a result a Taylor expansion [15] of the function in equation (2.3) can be performed to obtain

$$I(V_o + V_j) = I(V_o) + \left(V_j \frac{dI}{dV} \Big|_{V_o} \right) + \left(\frac{V_j^2}{2!} \frac{d^2 I}{dV^2} \Big|_{V_o} \right) + \dots \quad (2.4)$$

In this case, the output current $I(V)$ is due to the junction voltage which consists of both

DC bias (V_o) and AC signal (V_j) components. Evaluating the derivatives and ignoring the second term in equation (2.3), which is insignificant, leads to

$$I(V) = I(V_o) + V_j \beta I + \frac{V_j^2}{2!} \beta^2 I + \frac{V_j^3}{3!} \beta^3 I + \dots \quad (2.5)$$

Figure 2.2 Diode I-V transfer function [17]

In order for mixing to occur the input signal, which in this case is the junction voltage (V_j), must contain both the LO and RF voltage waveforms. Typically the junction voltage is composed of a summation of these waveforms. Therefore looking at equation (2.5), the generated current will consist of a summation of sinusoidal products at the RF and LO frequencies. Following the same reasoning as was presented for the ideal multiplier these sinusoidal products can be simplified to sinusoidal components containing the summation and difference of the LO and RF frequencies and their multiples. The following equation describes the generated frequency components (ω_{OUT}) defined as

$$\omega_{OUT} = m\omega_S + n\omega_P \quad (2.6)$$

where ω_P is the LO signal frequency, ω_S is the RF signal frequency, and m, n are integers from $-\infty$ to ∞ . An input and output spectrum illustrating the generated frequency components as a result of applied LO and RF signals is displayed in the figure below.

Figure 2.3 The generated mixing spectrum for applied RF and LO signals [17]

2.3 Operating The Diode As A Switch

Looking more closely at Figure 2.2, the signal is applied at a bias voltage equal to zero volts. The portion displaying the most nonlinearity is the area in the vicinity of (V_{th}) , the junction voltage which turns on the diode. Consequently, the most efficient mixing can be achieved by operating in this neighborhood, by applying a bias equal to the turn on voltage. At this point the diode turns on and begins to conduct much current. Below this point the diode is considered to be turned off and thus very little current is generated. The diode can be thought of as approaching the behaviour of a short circuit when turned on and an open circuit when turned off. Ideally the diode's I-V transfer function should approach that of the switching function, $S(t)$, as shown in Figure 2.4. Application of an input signal, V_{IN} , would then generate an output signal, V_{OUT} as shown.

Figure 2.4 Ideal transfer function for mixing [11]

The ideal nonlinear mixer would then be a switch that toggles between a short circuit and an open circuit. Since the diode cannot realistically behave in this manner, it must be ensured that operation is at the most nonlinear operating point possible. This requires that the junction voltage be large enough to fluctuate below and above the threshold voltage, the most nonlinear point in a diode's I-V curve. Since RF signals tend to be small, it is more feasible to make the LO signal large enough to effectively pump the diode between the on and off states.

2.4 Available Diode Structures

Besides having strong nonlinearity, the mixing device should also have low noise, low distortion, and adequate frequency response. Furthermore these electrical properties must be consistent between individual devices. The PN junction, point-contact, and Schottky-barrier diodes are typical diode structures commonly used as mixing devices.

The PN junction diode [18], [19] consists of a semiconductor crystal which has been doped with two concentrations of different charge carriers. One region consists of n-type (high electron concentration) material and the other p-type (high hole concentration) material. Current generation and control is achieved through the injection of both minority and majority charge carriers from one region to the other under the influence of applied voltage across the diode terminals. The fact that operation relies on minority carriers leads to charge storage effects which limit the switching speed of PN junctions. For this reason PN junction diodes have limited use in high frequency mixers.

The point-contact diode [16] consists of a pointed metal wire contact to a bulk semiconductor. The majority of point contact diodes are considered to be primitive forms of Schottky-barrier devices. These devices are majority carrier devices and as such do not suffer from minority carrier effects that limit their switching speed like PN junction diodes. However their fabrication process is not precise, involves comparatively more

manual labor than the other two diodes, and is more expensive. The point-contact diode was the dominant device for mixers before the advent of the PN junction and Schottky-barrier diodes.

The Schottky-barrier diode [16], [17], and [18] consists of a metal semiconductor interface and like the point-contact diode is a majority carrier device and thus does not face the same limitations as PN diodes. Unlike the point-contact diode, the Schottky-barrier device is fabricated photolithographically on epitaxial substrates. This leads to very good electrical characteristics and uniformity between devices. Lower junction capacitance, lower series resistance, and a better I-V transfer function (i. e. more non-linear, smaller turn on voltage, and higher breakdown voltage, etc.) are typical. For these reasons the Schottky-barrier diode is the predominant type of diode found in high frequency mixers.

2.5 The Structure Of The Schottky-Barrier Diode

At the heart of the Schottky-barrier diode is the metal-semiconductor junction. The nonlinear I-V characteristic inherent in this device is due to the movement of majority carriers (electrons typically) from the semiconductor to the metal. Only certain materials can be used to effectively perform this operation. Figure 2.5 provides an illustration of a typical Schottky-barrier diode, in particular a beam-lead diode (the one used), other types exist [11] but most have the same general form.

In this structure the diode chip consists of a lightly-doped thin epitaxial layer of semiconductor material which is used for the junction and is made thick enough to contain the depletion region over its entire range of operating bias voltages. This layer is highly resistive and is grown on a semiconductor substrate that is much thicker, is heavily doped, and has a lower resistance. Besides minimizing series resistance, this layer also makes it easier to fabricate a good quality ohmic contact between the substrate and the cathode. N-type GaAs or Si is the material exclusively used to realize the semiconductor.

The carrier mobility in n-type material is much higher than in p-type material, thereby offering lower series resistance and higher cutoff frequencies. As GaAs becomes less and less expensive and continues to offer higher and higher performance (less noise, higher electron mobility, higher breakdown voltages, and lower series resistance), it will continue to be the most popular semiconductor used.

The junction metal can affect the performance, electrical characteristics, and reliability of the diode. Typically copper, platinum, silver, aluminum, titanium, and gold are common metals used for the junction metal. Platinum and titanium are the most commonly used metals for GaAs based semiconductors. More information on the different types of metals and their implications can be found in [11]. Gold-germanium is almost exclusively used for ohmic contacts. The anode and ohmic contacts have integral gold ribbons for attaching the device to a circuit.

The figure below depicts a diode structure containing a thick oxide layer ($2 - 3 \mu\text{ms}$) that extends well beyond the edge of the diode chip. The support provided by the oxide layer allows the anode ribbon to be made very narrow. The combination of narrow ribbon and thick oxide layer results in low junction capacitance. Other alternative configurations exist in order to minimize the capacitance across the diode terminals [11].

a)

Figure 2.5 The beam-lead Schottky diode a) overall structure and b) layer profile [11].

2.6 The Energy Band Structure Of The Schottky Diode

Fundamental to the operation of the Schottky-barrier diode is the metal-semiconductor junction. In light of this, it is necessary to examine the energy band structure to gain insight into its operation. Rectification or diode-like behaviour occurs as a result of differences in the work functions of the metal (Ψ_M) and semiconductor (Ψ_S) rather than a nonuniform doping profile as in the case of PN diodes [18]. The work function is the energy required to move an electron from the surface of the material in question to free space. By having unequal work functions an electrostatic barrier is created between the metal and the semiconductor in the path of current flow. By varying the height of this barrier through applied voltage, the thermodynamic emission of majority carriers over the barrier can be controlled and thus so can the current.

Figure 2.6 Energy band structures for a) metal, b) semiconductor, c) and the metal-semiconductor junction [16] at zero bias

Figure 2.6 shows the energy band structure for the metal and semiconductor when they are isolated from one another and when they are joined through an ohmic contact. In the above diagram (W_F) is the Fermi energy level, a reference energy level where approximately half the electrons are at. In equilibrium the Fermi level is at a constant level throughout the material in question. (W_C) and (W_V) represent the energy levels in the conduction and valence bands of the semiconductor respectively. The conduction band is occupied by free electrons and the valence band is occupied by bound electrons. The Fermi energy level position in the semiconductor depends on the type and amount of doping. Figure 2.6 depicts a Fermi level that is closer to the valence band and is thus representative of a typical n-type semiconductor. In metal the conduction and

valence bands are not quite as distinct as in the semiconductor but overlap and thus the electrons can all be considered to be at the Fermi level.

The electrons of concern are in the conduction band of the semiconductor. A term known as the electron affinity (χ) is the energy required to remove an electron from the conduction band to free space. Looking at the conduction and valence bands of the isolated semiconductor, it is noted that the conduction and valence bands bend upwards upon approaching the surface of the semiconductor. In the ideal case there would be no band bending in the semiconductor, however due to the discontinuous crystal structure at the surface of certain semiconductors (GaAs among them), surface states are present [17]. These surface states acquire a net positive charge which leads to the observed upward bending of the semiconductor energy band edges.

Referring to Figure 2.6 a) and b), the Fermi level in the isolated metal is lower than that of the isolated semiconductor. Subsequently the work function for the metal has to be greater than that of the semiconductor. This signifies that the electrons in the semiconductor are at a higher energy state than those in the metal. Therefore when the semiconductor is brought into contact with the metal, electrons will flow spontaneously from the semiconductor to the metal. The combined structure will eventually reach a new equilibrium where energy levels shift to make the Fermi level the same in both the metal and semiconductor as shown in Figure 2.6 c). The potential across the contact is referred to as the barrier or contact potential (Δ) and is the difference between the work functions of the semiconductor and the metal ($\Psi_M - \Psi_S$).

The transfer of electrons from the semiconductor to the metal leaves behind a positively charged depletion region (void of mobile electrons) that reaches into the semiconductor by a distance termed the depletion width (d) and a negative charge on the surface of the metal. This positive charge in the semiconductor leads to further distortion (upward bending) of the energy band edges leading to the creation of an electrostatic barrier between the edge of the depletion region and the metal surface. As the contact dis-

tance (δ) is made smaller, the electric field will increase further and there will be more upward band bending. This electrostatic barrier will have a built-in potential denoted (Φ_{BI}) which is a measure of the degree of band bending. Any electrons in the conduction band of the semiconductor side wanting to pass into the metal must acquire a potential greater than this value. Also a metal-semiconductor barrier height denoted (Ψ_{MS}) and equal to $\Psi_{MS} = \Psi_M - \chi - \Delta$ will also be present to impede the flow of electrons from the metal to the semiconductor.

2.7 The I-V Characteristic Of The Schottky-Barrier Diode

At equilibrium the net current flow across the junction will be zero. The flow of electrons with energy greater than ($q\Psi_{MS}$) crossing the junction from the metal side will be balanced by an equal flow of electrons from the semiconductor side having energy levels greater than ($q\Phi_{BI}$). The current density flowing from the metal to the semiconductor ($J_{M \rightarrow C}$) and vice versa ($J_{C \rightarrow M}$) can be derived as in [18] to give

$$J_{M \rightarrow C} = J_{C \rightarrow M} = q\overline{v}_M N_D e^{-\left(\frac{q\Psi_{MS}}{kT}\right)} \quad (2.7)$$

where $\overline{v}_M$ is the average electron velocity and N_D is the electron doping density.

Upon application of a voltage signal, the metal-semiconductor structure will no longer be in equilibrium. The following figure shows what happens when a positive voltage (reverse bias) and negative voltage (forward bias) is applied to the semiconductor. The arrows indicate the direction of current flow.

Figure 2.7 The Schottky diode under forward (a) and reverse bias (b) [18]

The application of a negative voltage (Va) to the semiconductor raises the Fermi level on the semiconductor side and reduces the depletion width from its equilibrium value of d_0 to d . This in turn leads to a higher concentration of electrons at the semiconductor surface and a reduction in the barrier height from $(q\Phi_{BI})$ to $q(\Phi_{BI} - Va)$. Therefore a dramatic increase in the flow of electrons from the semiconductor to the metal occurs leading to the net current flow in the direction shown. The current density due to this increased electron flow is given by

$$J_{C \rightarrow M} = q \bar{v}_M N_D e^{-\left(\frac{q\psi_{MS}}{kT}\right)} e^{\left(\frac{qVa}{kT}\right)} . \quad (2.8)$$

The flow of electrons from the metal to the semiconductor remains unhindered since the barrier in this path remains unaffected. Thus the current density in this direction is the same as in the equilibrium condition.

Conversely the opposite situation occurs when a positive voltage is applied to the semiconductor. In this case the reduction in electron concentration and the increase in barrier height in the semiconductor is to such an extent that very little electron flow occurs from the semiconductor to the metal. The current flow in the device is then due solely to the flow of electrons from the metal to the semiconductor. Again the barrier height on the metal side remains unaffected and the current density in this direction is the same as in equilibrium.

In summary, the net current in reverse or forward bias can be expressed by the following expression

$$I_N = I_S \left(e^{\left(\frac{qV}{kT}\right)} - 1 \right) \quad (2.9)$$

as outlined in [17] where the saturation current density I_S is given by

$$I_S = A^* T^2 e^{-(q\psi_{MS})/(kT)} . \quad (2.10)$$

The factor A^* is a material constant termed the effective Richardson constant.

2.8 Junction Capacitance

Another important property of the Schottky-barrier diode is the junction capacitance. The movement of electrons from the semiconductor to the metal surface leads to a negative surface charge on the metal and a positively charged depletion region in the semiconductor. The resulting potential difference and charge storage across the metal-semiconductor interface sets up the observed junction capacitance. The nonlinear capacitance (C) across the depletion region can be expressed by the following relation between charge and junction voltage

$$\frac{dQ}{dV_J} = C(V_J). \quad (2.11)$$

The determination of the junction capacitance requires that the total charge (Q) that has been moved from the semiconductor to the metal be known. As an approximation it can be assumed that all donor atoms in the depletion region are ionized when the semiconductor and metal come into contact. Therefore the charge density of positive carriers left in the depletion region is equivalent to that of the original electron doping density in the semiconductor material (N_D) before contact. Knowing the junction area (A), the depletion width (d) and the charge per carrier (q), the total charge in the depletion region can be expressed as

$$Q = qdAN_D. \quad (2.12)$$

In the above expression, the depletion width (d) is the parameter dependent on the junction voltage. Therefore the depletion width must be expressed in terms of the junction voltage before the junction capacitance can be evaluated using equation (2.11). The general relationship between voltage and the electric field in one dimension is as follows

$$E = -\left(\frac{dV}{dx}\right) \quad . \quad (2.13)$$

The electric field is directed from the semiconductor to the metal, attains a maximum value of (qN_D/ϵ_S) at the metal-semiconductor junction (application of Gauss's law in one dimension [34]), and a minimum value of zero at the edge of the depletion region. Therefore the electric field at any point in the depletion region can be written as

$$E(x) = \frac{qN_D}{\epsilon_S} (x - d) \quad (2.14)$$

where (ϵ_S) is the permittivity or the dielectric constant of the semiconductor.

Consequently the voltage at any point in the depletion region using equation (2.13) is

$$V(x) = \frac{qN_D}{\epsilon_S} \left(\frac{x^2}{2} - xd \right) + \psi_{MS} \quad (2.15)$$

The junction voltage, also termed the built in potential, is found to be

$$\Phi_{bi} = V(x=d) - V(d=0) = \frac{qN_D d^2}{2\epsilon_S} \quad . \quad (2.16)$$

Solving this equation for the depletion width and using equations (2.12) and (2.11) to solve for the total charge and the junction capacitance respectively results in

$$C_j(V) = A \sqrt{\frac{qN_D \epsilon_S}{2\Phi_{bi}}} = \frac{A\epsilon_S}{d} \quad . \quad (2.17)$$

Under bias (V_O) all the above equations would hold true but Φ_{bi} would be replaced by $(\Phi_{bi} - V_O)$. The capacitance is usually expressed in the form

$$C_j(V) = \frac{C_{j0}}{\sqrt{1 - V/\Phi_{bi}}} \quad (2.18)$$

where C_{j0} is the junction capacitance at zero bias.

2.9 Series Resistance

Another important property of the Schottky-barrier diode that warrants discussion is the series resistance. The applied voltage does not appear entirely across the junction, but a portion of it falls across the series resistance. Therefore if the series resistance is not taken into account the diode current will be overstated under forward bias and understated under reverse bias.

The greatest contributor to the series resistance is the resistance associated with the undepleted portion of the epitaxial layer under the junction. The greater the thickness of the undepleted region the higher is the resistance. The epitaxial layer is made as thick as required to contain the depletion region when subjected to high reverse voltages. However, typical diode operation is at voltages in which the thickness of the undepleted portion of the semiconductor in the epitaxial layers remains fairly significant. The substrate is heavily doped to minimize its contribution to series resistance. However, it and the ohmic contact also add to the series resistance. Also at high frequencies the connecting wires also contribute to the series resistance. It is difficult to describe a general procedure for estimating series resistance since it depends on the diode structure. However, it can be approximated from the I-V characteristic of the diode as will be shown in a later section.

2.10 The Ideality Constant

The ideality factor η is a parameter introduced to take into account the non-ideal behaviour that is not accounted for in the I-V characteristic relation defined in (2.9). As η increases, the strength of the diode's nonlinearity decreases. This decrease in nonlinearity will lead to reduced efficiency in mixing. To take into account this nonideal behaviour the ideality factor is introduced into equation (2.9) as

$$I_N = I_S \left(e^{\frac{qV_j}{\eta(kT)}} - 1 \right). \quad (2.19)$$

There are a variety of reasons why physical diodes do not completely follow the ideal diode equation as defined by equation (2.9). The most well known ones are quantum mechanical tunnelling, surface imperfections, image forces, and edge effects [11], [17]. Quantum mechanical tunnelling is a mechanism whereby electrons with insufficient energy to overcome the potential barrier can cross the junction by burrowing through it. The surface imperfections that occur are a result of semiconductor contamination when processing the semiconductor and adding the junction metal. Image forces or Schottky-barrier lowering is a tendency for the barrier height (Ψ_{MS}) to vary with applied bias rather than remaining constant. Finally, edge effects is the term given to describe the effect of having fringing electric fields on the metal anode which are greater than the fields perpendicular to the junction.

2.11 Equivalent Diode Circuit

In previous sections the properties associated with the Schottky-barrier diode, namely the junction current, junction capacitance, series resistance, and ideality constant were examined. Equivalent circuit models are available. The one shown in the following figure takes into account the behaviour of these properties under both large (LO) and small (RF and IF) signal conditions.

Figure 2.8 Large and small-signal diode models [35]

In the large-signal (LO signal) model, the junction current and capacitance can be assumed to be nonlinear functions of the junction voltage alone and change instantaneously with it. Since the Schottky diode is largely immune to minority carrier effects this assumption can be made and equations (2.18) and (2.19) are valid in determining $C(V_j)$ and $I(V_j)$. Under large-signal conditions the capacitance is termed the incremental capacitance and the large-signal current across it is defined as

$$I_C(t) = \frac{dQ}{dt} = \left(\frac{dQ}{dV} \bigg|_{V=V_j(t)} \right) \cdot \left(\frac{d}{dt} V_j(t) \right) \quad (2.20)$$

The series resistance is considered to be a constant value and can be found from the DC I-

V characteristic as will be shown later. Other elements may be added to describe package capacitance, bond wire inductance, and other parasitics.

In the small-signal model, the I-V and C-V characteristics are linearized around the instantaneous large-signal voltage (LO signal) and the junction conductance $g(t)$ and capacitance $C(t)$ are treated as linear, time-varying elements. This is justified by the fact that the small-signal (RF) is a small perturbation of the large-signal (LO) junction voltage. Hence the diode conductance and capacitance presented to a small signal (RF) for any instant of time during the LO cycle can be found. The small-signal junction conductance is defined as the rate of change in junction current with respect to junction voltage

$$g(t) = \frac{d}{dV_j} I(V_j) = \frac{q}{\eta k T} I_o e^{\frac{q V_j(t)}{\eta k T}} = \frac{q}{\eta k T} (I(V_j(t)) + I_S) . \quad (2.21)$$

which gives the result that the junction conductance is proportional to current. Because it is small compared to $I(V_j)$ the I_S term in the above equation is often ignored. The linear, time-varying capacitance is defined as

$$C(t) = C(V_j(t)) \quad (2.22)$$

and the small-signal current across this capacitance is as follows

$$i_C(t) = \frac{d}{dt} [C(t) v(t)] = C(t) \frac{d}{dt} v(t) + v(t) \frac{d}{dt} C(t) \quad (2.23)$$

where $v(t)$ is the small-signal voltage. The series resistance and external elements that take into account parasitics are the same as in the large-signal case.

2.12 Diode Modelling

The GaAs Schottky-barrier diode is the nonlinear element chosen for the single-balanced diode mixer; specifically, the DMK2791-000 beam lead Schottky-barrier diode as provided by Alpha. It was selected since it was readily available and it could operate at 20 GHz, the highest operating frequency. Manufacturer's specifications can be found in Appendix A. A model that accurately portrays its operation is required to aid in the design of the overall mixer. Once a diode model has been chosen, its element parameters must be evaluated. This requires assigning them initial values, performing simulations, comparing simulation to measurement, adjustment of model parameters as necessary, and repeating the process until the parameter values are such that simulation approaches measurement.

2.12.1 The Diode Model

The basic diode model has been outlined in an earlier section. However, to more accurately represent the Schottky-barrier diode that will be used, a modified version will be implemented. Figure 2.9 illustrates the model that has been chosen to represent the Schottky-barrier diode.

Figure 2.9 Schottky-barrier diode model

The portion of the model contained within the dotted area incorporates the basic diode model. It is the diode model as implemented in OSA90/HarpeTM[20], a simulation and optimization software package that has been chosen for the diode modelling. The model is used in HarpeTM to represent both PN and Schottky-barrier diodes. The elements that constitute this subsection of the model are the forward current (I_D), the reverse current (I_B), the junction capacitance (C_J), and the diffusion capacitance (C_D). They are defined as follows:

$$I_D = I_S \cdot [e^{(qV_D)/(\eta \cdot k \cdot TEMP)} - 1] \quad (2.24)$$

$$I_B = I_{B0} \cdot (V_B - V_D)^{EB} \quad \text{for } V_D < V_B \quad (2.25)$$

$$I_B = 0 \quad \text{for } V_D \geq V_B \quad (2.26)$$

$$C_J = C_{J0} \cdot (1 - V_D/V_J)^{-EJ} \quad \text{for } V_D < FC \cdot V_J \quad (2.27)$$

$$C_J = C_{J0} \cdot (1 - FC)^{-EJ} \quad \text{for } V_D \geq FC \cdot V_J \quad (2.28)$$

$$C_D = C_{D0} \cdot e^{(qV_D)/(\eta \cdot k \cdot TEMP)} \quad (2.29)$$

The parameters used in the above equations representing the elements of the diode model are as follows

I_S is the saturation current,

η is the ideality coefficient,

$TEMP$ is the temperature,

V_B is the breakdown voltage,

V_D is the applied voltage across the terminals of the basic model,

I_{B0} is the breakdown current at $V_D = V_B$.

EB is the breakdown power law exponent,

C_{J0} is the junction capacitance at zero bias,

V_J is the junction potential,

FC is the coefficient for forward bias in the capacitance equation,

EJ is the junction capacitance exponent,

C_{D0} is the diffusion capacitance at zero bias.

The reverse current (I_B) can be ignored since it is only significant when the applied voltage approaches the breakdown voltage, an area where the diode will not operate. Also, the Schottky-barrier diode is a majority carrier device which operates under low level injection and does not support any significant stored charge under forward bias. Therefore the diffusion capacitance can be ignored as well.

The elements representing the added portion of the model are described as follows

R_{series} is the series resistance of the diode,

L_{bond} is the series inductance associated with the diode leads,

C_{par} is the overlay parasitic capacitance associated with the diode,

C_{shunt} is the parallel capacitance that takes into account fringing fields.

2.12.2 Initial Values For Diode Parameters

It is desirable to choose initial estimates for the diode parameters that are close to their final values, so that the time and effort that is required for convergence between simulation and measurement can be minimized. Good estimates for initial values can be obtained from manufacturer's specifications, DC I-V measurements, and DC C-V measurements.

Consulting the manufacturer's data, maximum values for R_{series} , C_{J0} , and C_{par} are 8 ohms, 0.025 pf, and 0.02 pf respectively.

Estimates for the saturation current (I_s) and the ideality constant (η) can be

made from DC I-V measurements of the diode [11]. Figure 2.12 shows a DC I-V curve where $\log(I)$ is plotted as a function of applied voltage (V). Looking at equation (2.19) which relates forward current (I) to applied voltage (V) and taking the log of both sides leads to

$$\log(I) = \log(I_S) + \log(e^{(qV)/(\eta kT)} - 1) \quad (2.30)$$

Noting that $\log = 2.3 \cdot \ln$ and that the -1 in the second term on the right hand side of equation (2.30) is insignificant compared to the exponential term, leads to

$$\log(I) = \log(I_S) + \frac{qV}{2.3\eta kT} \quad (2.31)$$

This expression is the general equation for a line where the y-intercept ($y\text{-int}$) and the slope (m) are defined as

$$y\text{-int} = \log I_S \quad (2.32)$$

$$m = \frac{q}{2.3\eta kT} \quad (2.33)$$

Therefore the linear portion of the DC I-V curve can be fitted to equation (2.31) and the above equations used to obtain estimates for the ideality constant and the saturation current.

An estimate for the series resistance can also be obtained from DC I-V measurements. Figure 2.12 illustrates that as the applied voltage is increased the $\log(I) - V$ response becomes nonlinear. This is as a result of the voltage drop across the series resistance becoming more significant. Therefore to estimate the series resistance (R_{series}), the deviation in voltage between the straight line and the measured I-V curve at a specified current can be used as follows

$$R_{series} = \frac{\Delta V}{I} \quad (2.34)$$

The junction capacitance at zero applied voltage, (C_{J0}), and the built in junction potential, Φ_{BI} , can be estimated from DC C-V measurements. By plotting $1/C$ versus voltage and fitting a straight line to these points the resulting y-intercept and x-intercept will be equal to $1/C_{J0}^2$ and Φ_{BI} respectively as shown in the following figure.

Figure 2.10 DC C-V measurements illustrating extraction of C_{J0} and Φ_{BI} [11]

2.12.3 Diode Measurement Setup

To perform DC I-V measurements requires mounting the diode in a test jig, a means to sweep the diode over a range of voltages, and a method to read the generated output current. In our case, diodes were mounted on an alumina substrate in two configurations. In the series configuration, a diode was mounted between two microstrip transmission lines. In the shunt configuration, another diode was placed between a microstrip transmission line and a metal shim that connected one end of the diode to the ground plane of the substrate. Both assemblies were placed in a Wiltron Universal transmission line test fixture. An automated setup as illustrated in Figure 2.11 was then used to run a software package, IMATM [38], on an HP computer that implemented the necessary routines to

sweep voltage, record, and store the resulting current.

Figure 2.11 DC I-V and S-parameter measurement setup

DC C-V measurements are more difficult to implement. They require the use of capacitance bridges. As diodes are made smaller, the corresponding zero bias junction conductance also gets smaller and it becomes increasingly more difficult to measure. There are techniques around this limitation but they are quite involved and impractical for the majority of diode users [11]. For these reasons DC C-V measurements were not performed.

Another set of diode measurements that can be performed are the small signal scattering parameters. At high frequencies these are the most common set of measurements employed to characterize operation of microwave passive and active circuitry. Scattering parameters relate reflected voltage waves from electrical ports to the incident voltage waves on the same or other ports [21].

S-parameter measurements require the use of a sophisticated and expensive

measuring instrument, namely a network analyzer. Typically network analyzers are two port instruments to which the device under test is connected. The network analyzer is able to send an incident voltage to one port of the circuit, measure the reflected voltage and transmitted waves off both electrical ports, and present a ratio of the reflected to incident voltage waves in a graphical format to the user. Figure 2.11 illustrates the experimental setup required for S-parameter measurements. It is basically the same setup employed to obtain the DC I-V response of the diode except for the inclusion of a Wiltron network analyzer and its cables. Once again the diodes were mounted on microstrip transmission lines in a series and shunt orientation and placed in a test jig; the Wiltron universal transmission line test fixture. For S-parameter measurements, the automated setup was driven by a PC computer implementing WafermapTM[39], another data-retrieval software package.

At high frequencies, the cables and connectors used to connect the network analyzer to the device under test affect measurements. For accurate measurements that truly reflect the operation of the diode, these effects must be removed. The means to remove these effects is to perform what is known as a premeasurement calibration. By performing a calibration, the plane of reference for measurements can be positioned so as not to include the cables and connectors. For this experimental setup, a Thru-Reflect-Line (TRL) two-port calibration was used. The principle behind the TRL calibration method can be found in [36].

2.12.4 Parameter Estimation From DC I-V Measurements

Figure 2.12 and Figure 2.13 show the DC I-V results obtained for both a series and shunt mounted diode respectively using the HP automated setup as illustrated in Figure 2.11. Here the log of the measured output current versus applied voltage was plotted for estimating the series resistance, the ideality constant, and the saturation current. Note that the data was collected for an applied voltage that was swept from slightly above 0 volts to approximately 1.3 volts. Measurement beyond 1.3 volts was not conducted for

fear of damaging the device with excessive current. Also the voltage was not swept below 0 volts since the diode's response below a junction voltage of 0.6 - 0.8 volts is approximately constant until the breakdown voltage is reached, an area where the diode will not be operated. Also displayed is the line segment that best fits the linear portion of the $\log(I)$ -V measurements. In both configurations the trace of current versus applied voltage is very similar. As a result the line segment that best fits the data for each case are also comparable and the extracted parameters should be similar.

Figure 2.12 Log(I) -V measurement for series mounted diode

Figure 2.13 Log(I) -V measurement for shunt mounted diode

As outlined in section 2.12.2 the line segment that best fits the linear section of the $\log(I) - V$ data can be used to estimate initial values for the saturation current and the ideality factor. Also, the nonlinear portion of the $\log(I) - V$ curve can be used to evaluate the series resistance. The saturation current, the ideality factor, and the series resistance for both the series and shunt cases were thus estimated and the results tabulated.

Table 2.1 Shunt And Series Diode Parameters

Parameter	Series	Shunt
I_S	1.185×10^{-14}	0.986×10^{-14}
η	1.1197	1.122
R_{series}	3.20Ω @ 18mA	4.56Ω @ 79mA
R_{series}	3.41Ω @ 31.2 mA	4.19Ω @ 39.7 mA

For both orientations the parameter results are very similar to one another except for the series resistance which differs by approximately 1 ohm but is within the maximum value of 8Ω as specified by the manufacturer. Parameter differences can be attributed to the fact that different diodes and mounting assemblies were used for each orientation. However slight parameter differences are not a major concern since these parameter values are starting points for simulation and measurement fits.

2.12.5 Diode Parameter Extraction

Once initial estimates have been obtained for the diode parameters, the next step is to extract parameters for the diode model. This is performed by fitting simulation to measurement. The two simulations that can be readily performed and compared

with measurements are the DC I-V and the small-signal scattering parameters.

Under DC analysis the capacitive and inductive components of the diode model become open circuits and short circuits respectively. Therefore the DC response is determined solely by the saturation current, the ideality constant, and the series resistance. Estimates for these parameters are provided in Table 2.1 and they can be used as initial values for the diode model. By optimizing the diode model such that the simulated DC I-V response approaches measurement more accurate values for these parameters can be obtained. At the same time these parameters must be such that simulated small signal S-parameters agree with measured small signal S-parameters which will in turn be used to extract the remaining parameters.

The following is a list of initial values used for the diode parameters for both the series and shunt mounted cases.

$C_{par} = 0.02$ pf as determined by the manufacturer

$C_{DO} = 0$ pf since diffusion capacitance is insignificant

$C_{JO} = 0.01$ pf lies within specified maximum of 0.025 pf

$L_{series} = 0.1$ nH is a reasonable estimate

$TEMP = 298$ K is equivalent to room temperature

$V_B = -\infty$ is the default value and is not of concern

$I_{B0} = 0$ is the default value and is not of concern

$EB = 10$ is the default value and is not of concern

$V_J = 0.7$ V is a typical value for a Schottky-barrier diode

$FC = 0.8$ is realistic for Schottky-barrier diodes as given by [35]

$EJ = 0.5$ is a typical value

$C_{shunt} = 0.01$ pf is reasonable estimate

The measured and simulated DC I-V responses for both the series and shunt mounted diode are shown below. The diode model was able to simulate a DC I-V response that was a better fit to measurement for the series case than for the shunt case. A

model's fit to the small signal S-parameter measurements.

Figure 2.14 Measured and simulated DC I-V for series mounted diode

Figure 2.15 Measured and simulated DC I-V for shunt mounted diode

Two-port and one-port S-parameter measurements for both a series and a shunt mounted diode respectively were obtained. These small-signal S-parameter mea-

measurements were performed for a sweep of applied voltages from -1.3 volts to 1.3 volts over a frequency range that spanned from 15 GHz to 25 GHz. The attraction of the shunt mounted diode is that it requires only one S-parameter measurement for each frequency point whereas the series case requires four. However, one measurement might not be enough to accurately characterize the device especially when it will be used in a series configuration in the final mixer circuit. Therefore both sets of data were used to see if the model could be optimized to converge to the same model parameters when simulation was optimized to fit measurement.

For the series configured diode, the plane of reference was chosen to be at the diode terminals. However, the diode in shunt was mounted on transmission lines that were longer than the ones used in the series configuration. Using the same calibration standards would not place the reference plane at the diode terminals for the shunt case. Therefore when the diode model will be used to simulate the shunt mounted diode, extra pieces of transmission line must be included in the model to account for their presence during measurement. Also a resistance of several ohms at each port was measured in the path connecting the applied voltage to the network analyzer. These resistances must also be taken into account when we attempt to fit the simulated diode model to measurements.

Figures 3.8 - 3.11 illustrate the measured S-parameters obtained in both an off state ($V_D = -0.6$ V) and an on condition ($V_D = 1.1$ V) over 15 to 25 GHz for the series mounted diode. Also displayed on the same graphs are the simulated results that were found to best fit measurements for the optimized diode model. Examining the graphs more closely, we see that S_{11} and S_{22} are similar for both measurement and simulation and the same holds true for S_{21} and S_{12} respectively. More simulation and data for various biases can be found in Appendix B along with HarpeTM program listings. The final parameter values obtained from the optimized model can be found in Table 2.2.

Figure 2.16 Measured and simulated S_{11} parameter for series mounted diode

Figure 2.17 Measured and simulated S_{22} parameter for series mounted diode

Figure 2.18 Measured and simulated S_{21} parameter for series mounted diode

Figure 2.19 Measured and simulated S_{12} parameter for series mounted diode

Figure 2.20 and Figure 2.21 illustrate S_{11} parameter measurements and simulations for the shunt mounted diode when it is off ($V_D = -0.6$ V) and when it is on ($V_D = 1.1$ V) over 15 to 25 GHz. In both graphs the start and stop frequency points are denoted S and E for both the measurement and simulation curves. The subscripts S and M further denote whether the start and end points of the curve in question pertain to simulation or measurement. Looking at the off condition, the magnitude of the simulated S_{11} parameter does not agree with measurement. However in the on condition the agreement between simulation and measurement in terms of magnitude is a much better fit. In both cases it is noticeable that the phases between simulation and measurements are not in agreement.

Figure 2.20 Measured and simulated S_{11} parameter of shunt mounted diode in the off condition

Figure 2.21 Measured and simulated S_{11} parameter of shunt mounted diode in the on condition

Table 2.2 summarizes the final parameter values for the diode model as obtained by fitting simulation to measurement for both the series and shunt mounted diode. The DC I-V and S-parameter simulations in the series case fitted measurements quite well. However the same cannot be said for the shunt case. Despite repeated attempts, the DC I-V and S-parameter simulations illustrated in the above figures provided the best attainable fits.

A possible explanation as to why the S-parameter fit for the series case was much better than for the shunt case might lie in the measured series resistances of the network analyzer's ports that were added to the diode model. Using the ports of the network analyzer for DC measurements would require that these resistances be taken into account. However for S-parameter measurements, RF calibration takes into account these resistances and therefore they should not be included in the diode model as was done. For the series case an added small series resistance is insignificant since the diode is terminated in 50 ohms. However, for the shunt case a small series resistance is significant since the

diode is terminated in a short circuit to ground. Although this realization was made to late for verification, the model as determined for the series mounted diode was the one chosen to be used in the mixer circuit since it provided good agreement between measurement and simulation at the time of parameter extraction.

Table 2.2 Final Parameter Values For Series And Shunt Mounted Diodes

Parameters	Shunt	Series
I_{sat}	$1.074 \times 10^{-14} \text{A}$	$1.07 \times 10^{-14} \text{A}$
η	1.24	1.11
R_{series}	3.81Ω	4.04Ω
C_{par}	0.02pf	0.02pf
L_{series}	0.04nH	0.025nH
C_{j0}	0.032pf	0.025pf
V_J	0.6V	0.76V
FC	0.8	0.8
EJ	0.5	0.5

CHAPTER 3: THE 180-DEGREE HYBRID COUPLER

3.1 Introduction

Figure 3.1 illustrates a single-balanced mixer consisting of two nonlinear devices (Schottky-barrier diodes) connected in an anti-parallel configuration. Each diode is a mixing element and thus requires that both the LO and RF waveforms be present across their terminals for the mixing operation to occur. Therefore a method is needed that will allow the RF and LO signals to be applied to the mixer circuit such that they appear across both nonlinear devices. At the same time it is also necessary that both these signals travel from their respective input ports to the mixer and do not appear at the other signal's port to ensure efficient mixer operation. The most common method employed to meet all these concerns is to apply the LO and RF signal into the mixer circuit via a hybrid coupler.

Figure 3.1 The single-balanced diode mixer [11]

3.2 Structure And Characteristics

The hybrid coupler is a passive structure consisting of four ports, two of which are used as inputs and the others as outputs. Ideally it should have the following characteristics:

- all ports are matched
- equal power split in output ports
- input ports are isolated from each other

- output ports are isolated from each other

According to [21], the properties of any lossless four port structure with the above characteristics are such that the hybrid coupler can only be realized in two forms. These two varieties are commonly referred to as the 180-degree and 90-degree coupler.

Figure 3.2 and Figure 3.3 illustrate the operation of the 180-degree coupler and the 90-degree coupler respectively. The signal flow between ports is illustrated graphically and mathematically through an S-parameter matrix. Referring to Figure 3.2, the 180-degree hybrid operates in the following manner. An incident signal applied at port 1

Figure 3.2 The 180-degree hybrid coupler [11]

divides equally in power and appears at ports 3 and 4 in phase. Ports 1 and 2 are inherently isolated and therefore no portion of the signal applied at port 1 appears at port 2. Similarly, a signal incident on port 2 divides equally in power, but appears at ports 3 and 4 out of phase by 180 degrees. Again isolation between ports 1 and 2 exists and does not allow the applied signal to reach port 1. Signals applied at both input ports would thus appear as a summation at output port 3 and as a difference at output port 4. For this reason ports 3 and 4 are often called the sigma (Σ) and delta (Δ) ports respectively. Similar operation would hold if ports 3 and 4 were used as the inputs and ports 1 and 2 as the out-

puts. Referring to Figure 3.3, the operation of the 90-degree hybrid is very similar except that the output signals are 90 degrees out of phase at the output ports when an incident voltage is applied at either input port.

Figure 3.3 The 90-degree hybrid coupler [11]

3.3 Non-ideal Hybrid Couplers

Hybrid coupler operation as described in the previous section assumed an ideal hybrid coupler, however in reality various non-idealities exist that will make operation deviate from the ideal. Power losses will always be present in the components used to realize the hybrid structure. As a consequence the signals appearing at the output ports will not contain half ($S_{21} = -3\text{dB}$) of the input signal's power but less than half ($S_{21} < -3\text{dB}$). Also, complete isolation between ports cannot be achieved and therefore portions of unwanted signals will find their way to ports that they were not intended for. Perfect port matching cannot be implemented and therefore reflections will exist degrading performance even further. Finally, amplitude and phase balance on each arm of the output ports will deviate from the ideal condition leading to less than ideal hybrid performance.

3.4 The Waveguide "Magic-T"

The waveguide "Magic-T" of Figure 3.4 is one type of 180-degree hybrid coupler that could be implemented for use in a single-balanced mixer. In this case the RF and LO signals are applied to the inputs of the "Magic-T" as shown. The summation and difference of the applied LO and RF voltage signals appear at the output ports as illustrated. In a later section it will be shown that for a balanced mixer, the RF and LO signals can be applied to either input port and operation would be essentially the same.

Figure 3.4 The waveguide "Magic-T"

Ideally, if perfect isolation exists between RF and LO input ports and between output ports, the "Magic-T" can be partitioned into an E-plane tee and an H-plane tee in order to better describe its operation under both RF and LO excitation. Figure 3.5 illustrates the electric field distribution associated with both these structures for an applied RF and LO signal. Here the LO signal and the RF signal are the inputs to the H-plane tee and the E-plane tee respectively. Ideally, the input signal power to each structure divides evenly into both output branches as depicted by the voltage output signals. Examining the electric fields between output ports, it is observed that their orientation is the same in the H-plane tee and opposite in the E-plane tee. This implies that no phase difference exists

between output signals for the H-plane tee and that 180 degree phase difference exists between output signals for the E-plane tee. Therefore combining both structures to form the "Magic-T" leads to a summation and difference of the applied signals at the output ports of the "Magic-T" as shown in Figure 3.4.

Figure 3.5 The waveguide H-plane and E-plane tees

3.5 The Microstrip T-Junction: H-Plane Tee Realization

The T-junction as illustrated in Figure 3.6 is the microstrip counterpart to the waveguide H-plane tee. It basically consists of two microstrip transmission lines oriented perpendicularly to each other such that the top and bottom conductors of each one lie on the same plane. Thus the field associated with an applied signal at any one port (labelled 1, 2, and 3) will be guided along between the conductors and the ground plane.

Figure 3.6 The microstrip T-junction [17]

References [22], [37] discuss this microstrip structure more extensively than will be covered here. Figure 3.7 highlights one equivalent circuit that is used to represent the microstrip T-junction. Looking at ports labelled 2 and 3, it is evident that the circuit equivalents of these two arms are identical. Therefore, it can be deduced that a signal applied at port 1 will reach the junction and the portion of this signal remaining at this point will divide evenly in power into both arms (ports 2 and 3). Also this structural symmetry ensures that there is no phase difference between the signals appearing at output ports 2 and 3.

Figure 3.7 The equivalent circuit of the microstrip T-junction [17]

3.6 Aperture Coupling: E-Plane Tee Realization

The realization of the E-plane tee in microstrip requires a structure that splits the power of an input signal equally between two output ports and 180 degrees out of phase. At the same time this structure must be such that it can be readily combined with the microstrip T-junction in order to create the "Magic-T". It is the combination of these E-plane and H-plane tee structures that is needed to fulfill the operating characteristics of the "Magic-T". The outputs of this structure must be a difference of input signals at one port and a summation of input signals at the other.

The aperture-coupled structure that is to be used to implement the E-plane tee is displayed in Figure 3.8. It is a three port circuit consisting of two parallel microstrip transmission lines mounted back-to-back along a common ground plane. The bottom transmission line is terminated in an open circuit and is used to feed the input signal into the circuit. An aperture in the ground plane oriented perpendicularly to both transmission lines allows the signal to couple from the bottom transmission line to the top one. Here the coupled signal will divide in two and will proceed towards ports 2 and 3 180 degrees out of phase. The next sections will explain how such a structure can have the properties of an E-plane tee.

Figure 3.8 The aperture-coupled microstrip line structure

3.7 Introduction To Aperture Coupling

The general concept of aperture coupling between two regions is illustrated in Figure 3.9. These regions are separated from one another by an electric wall (perfect conductor) and are joined together by an aperture in this common wall. The two regions are bounded by walls, however one or both could be unbounded. The source of field excitation can be in the form of an electric current, magnetic current, or both and it can be present in either or both regions. In the example shown, region A contains the excitation represented as impressed electric (J_i) and magnetic (M_i) current densities.

The sources J_i and M_i will radiate electric and magnetic fields in the presence of the aperture into region A denoted E_A , H_A [44]. The aperture will also allow electric and magnetic fields denoted E_B , H_B to be set up in region B. The analysis of the aperture coupled structure of Figure 3.9 involves the determination of these fields under the existing boundary conditions. Since more than one source is present an analysis can

consider the fields due to one source at a time and then use superposition to determine the resultant fields.

Figure 3.9 Aperture coupling between two regions

3.8 The Field Distribution In The Aperture-Coupled Line Structure

The previous section introduced the concept of aperture coupling. In this section the same concepts will be applied to the aperture-coupled transmission line structure in order to determine its field distribution. This will be necessary to illustrate E-plane operation later on.

The cross-sectional view of Figure 3.10 shows two microstrip transmission lines mounted back-to-back along a common ground plane on which a slot has been cut to allow coupling between the two regions. Region A consists of a microstrip feed line which carries two current densities ($\vec{J}_{inc}$ and $\vec{J}_{scat}$). These current densities are associated with the applied signal and that which is scattered back onto the conductor by the discontinuity in the ground plane presented by the aperture respectively. Region B contains the microstrip line to which the signal travelling along the transmission line in region A is

to couple to. The current density of the coupled signal is denoted by $\vec{J}_{coup}$ as shown in Figure 3.10.

Figure 3.10 The microstrip E-plane tee [44]

The sources $\vec{J}_{inc}$, $\vec{J}_{scat}$, and $\vec{J}_{coup}$ will radiate fields into their respective regions and the presence of the aperture allows them to set up fields in the other region as well. The total field in each region A and B denoted $\vec{E}_A^{tot}$, $\vec{H}_A^{tot}$, and $\vec{E}_B^{tot}$, $\vec{H}_B^{tot}$ respectively is a superposition of the field contributions in that region from each source. Thus the total field and the contribution from each source can be expressed as

$$\vec{E}_A^{tot} = \vec{E}_A(\vec{J}_{inc}) + \vec{E}_A(\vec{J}_{scat}) + \vec{E}_A(\vec{J}_{coup}) \quad (3.1)$$

$$\vec{H}_A^{tot} = \vec{H}_A(\vec{J}_{inc}) + \vec{H}_A(\vec{J}_{scat}) + \vec{H}_A(\vec{J}_{coup}) \quad (3.2)$$

$$\vec{E}_B^{tot} = \vec{E}_B(\vec{J}_{inc}) + \vec{E}_B(\vec{J}_{scat}) + \vec{E}_B(\vec{J}_{coup}) \quad (3.3)$$

$$\vec{H}_B^{tot} = \vec{H}_B(\vec{J}_{inc}) + \vec{H}_B(\vec{J}_{scat}) + \vec{H}_B(\vec{J}_{coup}) \quad (3.4)$$

3.9 Aperture-Coupled Line Equivalent

The analysis of many electromagnetic problems can be simplified by forming equivalent problems through the application of what is commonly referred to as the equivalence principle. In this principle, the discontinuity is replaced by an equivalent source which radiates the same fields as scattered by the discontinuity [23], [24]. Various implementations of this method are typically used to aid in the evaluation of scattering; diffraction, and aperture antenna problems.

With respect to the structure of Figure 3.10, two imaginary surfaces are placed along the x-z plane on both sides of the aperture extending to infinity in both planes. These imaginary surfaces are chosen to coincide with the tangential components of the electric and magnetic fields they contain. Knowing the tangential fields along these surfaces the corresponding surface current densities can be evaluated. In turn, these current densities will set up fields in each region such that the original field distribution is maintained. Figure 3.11 illustrates the aperture-coupled structure where an electric wall has been used to close the aperture leading to current densities $\vec{J}_M$ and $-\vec{J}_M$. The resulting fields can be expressed as

$$\vec{E}_A^{tot} = \vec{E}_A(\vec{J}_{inc}) + \vec{E}_A(\vec{J}_{scat}) + \vec{E}_A(-\vec{J}_M) \quad (3.5)$$

$$\vec{H}_A^{tot} = \vec{H}_A(\vec{J}_{inc}) + \vec{H}_A(\vec{J}_{scat}) + \vec{H}_A(-\vec{J}_M) \quad (3.6)$$

$$\vec{E}_B^{tot} = \vec{E}_B(\vec{J}_M) + \vec{E}_B(\vec{J}_{coup}) \quad (3.7)$$

$$\vec{H}_B^{tot} = \vec{H}_B(\vec{J}_M) + \vec{H}_B(\vec{J}_{coup}) \quad (3.8)$$

Figure 3.11 The microstrip E-plane tee after application of the equivalence principle [44]

3.10 The Planar Waveguide Model For Microstrip Lines

Unfortunately the solution to the field equations in the previous section does not lead to simple closed form analytical expressions that can be readily used to solve for the field distribution of planar transmission lines such as microstrip. In an ideal transmission line the corresponding fields can be described as TEM (transverse electric and magnetic fields) to the direction of wave propagation. Maxwell's electromagnetic equations can be readily solved to determine the field distribution in this case. However, microstrip transmission lines are inhomogeneous structures that radiate fields both in the dielectric material between the top and bottom conducting portions as well as in the air surrounding the top conductor as shown in Figure 3.12 and Figure 3.13. As a result the fields in microstrip are referred to as quasi-TEM in nature at low frequencies (a few GHz). As the frequency approaches zero the field distribution looks more like that of a TEM mode. At higher frequencies other hybrid modes enter into the picture and the field distribution becomes more complicated as shown in Figure 3.13.

Figure 3.12 The microstrip line (a) and its field distribution at low frequencies (b) [22]

Figure 3.13 The magnetic (a) and electric (b) field distribution in microstrip line at high frequencies [22]

In general, hybrid modes can be described in terms of a summation of TE (transverse electric) and TM (transverse magnetic) modes with respect to the direction of wave propagation. Typically these modes are associated with waveguides and through the

application of the planar wave guide model [22], the microstrip transmission line can be treated as a special type of waveguide. Here the side walls are treated as magnetic walls and the top and bottom conductors are treated as electric walls to meet the same boundary conditions on its surface as the microstrip line, namely $\vec{n} \times \vec{H} = 0$ on the side walls and $\vec{n} \times \vec{E} = 0$ on the top and bottom walls. The fields in this structure can be represented by a summation of an infinite series of waveguide modes. In this case these modes are the TM and TE fields discussed earlier. The field distribution can then be expressed as follows

$$\vec{E} = \sum a_n \vec{E}_n^+ \text{ for } z > 0 \quad (3.9)$$

$$\vec{H} = \sum a_n \vec{H}_n^+ \text{ for } z > 0 \quad (3.10)$$

$$\vec{E} = \sum b_n \vec{E}_n^- \text{ for } z < 0 \quad (3.11)$$

$$\vec{H} = \sum b_n \vec{H}_n^- \text{ for } z < 0 \quad (3.12)$$

where the fields $\vec{E}_n^+$, $\vec{H}_n^+$, $\vec{E}_n^-$, $\vec{H}_n^-$ can be solved using Maxwell's equations subject to boundary conditions. The superscripts + and - denote propagation in the +z direction and the -z direction respectively. The expansion coefficients a_n , b_n can be determined using what is known as the Lorentz Reciprocity principle as outlined in [25].

3.11 Bethe's Aperture Coupling Theory

According to the theory as proposed by Bethe [46], the effect of small apertures in a conducting wall between two regions on the field distribution can be equated to electric (P) and magnetic (M) dipoles radiating in the presence of an electric wall. In section 3.9 the aperture was replaced by an electric wall and current densities $\vec{J}_M$ and $-\vec{J}_M$. The relationship between dipoles and current densities can be expressed as

$$\vec{P} = \epsilon \int_{S_A} -\frac{\vec{r} \times \vec{J}_m}{2} dS \quad (3.13)$$

$$\vec{M} = \frac{1}{\mu} \int_{S_A} -\vec{r} (\nabla \cdot \vec{J}_m) dS \quad (3.14)$$

where r is the radius of the aperture, ϵ is the aperture permittivity, μ is the aperture permeability, and the integral is evaluated over the surface area of the aperture.

The fields coupled through an aperture are attributed to the normal electric and tangential magnetic fields incident to it. Figure 3.14 (a) and Figure 3.14 (d) shows a normal electric field and a tangential magnetic field incident on an electric wall respectively. Figure 3.14 (b) and Figure 3.14 (e) illustrates what occurs to these fields when an aperture is placed in the electric wall. The same effect can be realized by closing the aperture by an electric wall and placing an electric dipole normal to it, Figure 3.14 (c), and a magnetic dipole tangential to it, Figure 3.14 (f), to account for the fringing of electric and magnetic fields respectively.

Figure 3.14 The field distribution through an aperture and its equivalent dipole representation [21]

As a first approximation the electric dipole is proportional to the normal electric field and the magnetic dipole is proportional to the tangential magnetic field incident to the aperture with it replaced by an electric wall. The constants of proportionality are referred to as electric α_e and magnetic α_m polarizabilities and depend on the aperture size and shape. [26] outlines a means to evaluate them. From [17], the electric and magnetic dipoles can be expressed by the following equations.

$$\vec{P} = -\epsilon_o \epsilon(f) \alpha_e \vec{E}_n \quad (3.15)$$

$$\vec{M} = -\alpha_m \vec{H}_t \quad (3.16)$$

$$\epsilon(f) = \frac{\epsilon_{eA}(f) \bullet \epsilon_{eB}(f)}{\epsilon_{eA}(f) + \epsilon_{eB}(f)} \quad (3.17)$$

where ϵ_{eA} and ϵ_{eB} are the frequency dependent effective dielectric constants in region A and B respectively. It should be noted that equations (3.15) and (3.16) do not portray the aperture coupling completely but are only approximations. References [25] and [45] pro-

vide a more thorough representation of the dipole strengths. Reference [17] also provides factors to take into account larger apertures and finite ground plane thickness.

3.12 E-Plane Operation Of Aperture-Coupled Lines

In Section 3.8 the field distribution in the aperture-coupled microstrip line structure brought about by impressed current sources was examined. In the following section the aperture was replaced by a magnetic current (J_m) radiating in the presence of an electric wall to generate the same field distribution as in the original structure. Reviewing Bethe's theory for coupling through small apertures it was shown that the magnetic current (J_m) could be replaced by equivalent electric and magnetic dipoles to describe the aperture-coupled fields. Section 3.10 showed that the field distribution in a microstrip line can be expressed in terms of a summation of TM and TE modes. Using this dipole representation and expressing fields in terms of TE and TM modes, E-plane operation of the aperture-coupled microstrip line structure can now be demonstrated.

The fields radiated by the electric dipole into region B can be stated as follows

$$\vec{E}_{Be} = \begin{cases} A_1 \vec{E}_A^+ & z \geq 0 \\ A_2 \vec{E}_A^- & z \leq 0 \end{cases} \quad (3.18)$$

$$\vec{H}_{Be} = \begin{cases} A_1 \vec{H}_A^+ & z \geq 0 \\ A_2 \vec{H}_A^- & z \leq 0 \end{cases} \quad (3.19)$$

and those radiated by the magnetic dipole into region B can be expressed as

$$\vec{E}_{Bm} = \begin{cases} B_1 \vec{E}_A^+ & z \geq 0 \\ B_2 \vec{E}_A^- & z \leq 0 \end{cases} \quad (3.20)$$

$$\vec{H}_{Bm} = \begin{cases} B_1 \vec{H}_A^+ & z \geq 0 \\ B_2 \vec{H}_A^- & z \leq 0 \end{cases} . \quad (3.21)$$

$\vec{E}_A^+$, $\vec{H}_A^+$, $\vec{E}_A^-$, $\vec{H}_A^-$ are the incident electric and magnetic fields in region A and represent a summation of TE and TM modes. The total electric and magnetic field radiated into region B can be expressed as

$$\vec{E}_{B(e+m)} = \begin{cases} (A_1 + B_1) \vec{E}_A^+ & z \geq 0 \\ (A_2 + B_2) \vec{E}_A^- & z \leq 0 \end{cases} \quad (3.22)$$

$$\vec{H}_{B(e+m)} = \begin{cases} (A_1 + B_1) \vec{H}_A^+ & z \geq 0 \\ (A_2 + B_2) \vec{H}_A^- & z \leq 0 \end{cases} \quad (3.23)$$

The amplitude coefficients A_1 , A_2 , B_1 , B_2 can be evaluated as in [17]

$$A_1 = A_2 = \frac{j\omega}{P_N} \left(\vec{E}_A \cdot \vec{P} \right) \quad (3.24)$$

$$B_1 = -B_2 = \frac{j\omega\mu_o}{P_N} \left(\vec{H}_A \cdot \vec{M} \right) \quad (3.25)$$

where P_N is a normalization factor defined at $z=0$ as

$$P_N = 2 \iint_S \left(\vec{E}^+ \times \vec{H}^+ \cdot \vec{z} \right) dS = abY_\omega \quad (3.26)$$

and a , b are the width, height of the cross sectional area of the dielectric region in the direction of propagation. Y_ω is the wave impedance in the path of propagation.

As an approximation, the incident signal to the aperture from the microstrip feedline can be considered to be a TE mode to the direction of propagation which is along the z -axis. Considering only this mode the incident fields propagating in the z direction can be described as

$$\vec{E}_A^+ = \vec{E}_A^- = C \sin\left(\frac{\pi x}{a}\right) \vec{y} \quad (3.27)$$

$$\vec{H}_A^+ = -CY_\omega \sin\left(\frac{\pi x}{a}\right) \vec{x} + j \frac{\pi Y_\omega}{\beta a} C \cos\left(\frac{\pi x}{a}\right) \vec{z} \quad (3.28)$$

$$\vec{H}_A^- = CY_\omega \sin\left(\frac{\pi x}{a}\right) \vec{x} + j \frac{\pi Y_\omega}{\beta a} C \cos\left(\frac{\pi x}{a}\right) \vec{z} \quad (3.29)$$

where C is the amplitude of the incident field and β is the propagation phase constant.

Consequently A_1 , A_2 , B_1 , and B_2 are given by

$$A_1 = A_2 = -\frac{j\omega}{abY_\omega} \vec{P} \cdot \vec{y} C \sin\left(\frac{\pi}{2}\right) = -j \frac{\omega P_y}{abY_\omega} C \quad (3.30)$$

$$B_1 = \frac{j\omega\mu}{ab} \left(C \sin\left(\frac{\pi}{2}\right) \vec{x} + j \frac{\pi}{\beta a} C \cos\left(\frac{\pi}{2}\right) \vec{z} \right) \cdot \left(M_x \vec{x} + M_z \vec{z} \right) = \frac{j\omega\mu}{ab} C M_x \quad (3.31)$$

$$B_2 = \frac{j\omega\mu}{ab} \left(-C \sin\left(\frac{\pi}{2}\right) \vec{x} + j \frac{\pi}{\beta a} C \cos\left(\frac{\pi}{2}\right) \vec{z} \right) \cdot \left(M_x \vec{x} + M_z \vec{z} \right) = -\frac{j\omega\mu}{ab} C M_x \quad (3.32)$$

where $P_N = abY_\omega$ and $x = a/2$ (at the centre of the aperture). Therefore the electric

and magnetic fields radiated into region B by the electric and magnetic dipoles are given by

$$\vec{E}_{B(e+m)} = \begin{cases} \left(j\frac{\omega P_y}{abY_\omega}C + \frac{j\omega\mu}{ab}CM_X \right) \vec{E}_A^+ & z \geq 0 \\ \left(j\frac{\omega P_y}{abY_\omega}C - \frac{j\omega\mu}{ab}CM_X \right) \vec{E}_A^- & z \leq 0 \end{cases} \quad (3.33)$$

$$\vec{H}_{B(e+m)} = \begin{cases} \left(j\frac{\omega P_y}{abY_\omega}C + \frac{j\omega\mu}{ab}CM_X \right) \vec{H}_A^+ & z \geq 0 \\ \left(j\frac{\omega P_y}{abY_\omega}C - \frac{j\omega\mu}{ab}CM_X \right) \vec{H}_A^- & z \leq 0 \end{cases} \quad (3.34)$$

According to [44] the main source of radiation from region A to region B is due to the magnetic dipole and therefore the above expressions reduce to

$$\vec{E}_{B(e+m)} = \begin{cases} \left(\frac{j\omega\mu}{ab}CM_X \right) \vec{E}_A^+ & z \geq 0 \\ \left(-\frac{j\omega\mu}{ab}CM_X \right) \vec{E}_A^- & z \leq 0 \end{cases} \quad (3.35)$$

$$\vec{H}_{B(e+m)} = \begin{cases} \left(\frac{j\omega\mu}{ab}CM_X \right) \vec{H}_A^+ & z \geq 0 \\ \left(-\frac{j\omega\mu}{ab}CM_X \right) \vec{H}_A^- & z \leq 0 \end{cases} \quad (3.36)$$

Substituting equations (3.27), (3.28), and (3.29) for $\vec{E}_A^+$, $\vec{E}_A^-$, $\vec{H}_A^+$, and $\vec{H}_A^-$ into the above expressions we can rewrite equations (3.35) and (3.36) as

$$\vec{E}_{B(e+m)} = \begin{cases} \left(\frac{j\omega\mu}{ab} C^2 M_x \right) \sin\left(\frac{\pi x}{a}\right) \hat{y} & z \geq 0 \\ \left(-\frac{j\omega\mu}{ab} C^2 M_x \right) \sin\left(\frac{\pi x}{a}\right) \hat{y} & z \leq 0 \end{cases} \quad (3.37)$$

$$\vec{H}_{B(e+m)} = \begin{cases} \left(\frac{j\omega\mu}{ab} C^2 M_x \right) \left(-Y_\omega \sin\left(\frac{\pi x}{a}\right) \hat{x} + j \frac{\pi Y_\omega}{\beta a} \cos\left(\frac{\pi x}{a}\right) \hat{z} \right) & z \geq 0 \\ \left(\frac{j\omega\mu}{ab} C^2 M_x \right) \left(-Y_\omega \sin\left(\frac{\pi x}{a}\right) \hat{x} - j \frac{\pi Y_\omega}{\beta a} \cos\left(\frac{\pi x}{a}\right) \hat{z} \right) & z \leq 0 \end{cases} \quad (3.38)$$

Looking at the electric fields coupled into region B, propagating in the +z and -z directions, it is noted that their magnitudes are equal but opposite in phase. The voltage of the coupled signal is related to the coupled electric field through the following relation

$$V = - \int_y E dy. \quad (3.39)$$

As a result the voltage in the +z and -z direction also has the same magnitude and opposite phase. Therefore, E-plane operation can be realized with aperture coupled microstrip lines as long as the aperture dimensions are chosen such that as close to half of the incident signal power as possible is coupled onto each output arm of the output line.

3.13 The Microstrip Based "Magic-T"

Integrating the microstrip based H-plane tee and E-plane tee, namely the microstrip T-junction and aperture-coupled microstrip line structure respectively, the result is the microstrip "Magic-T" as shown in Figure 3.15. Here the aperture-coupled structure consists of a feedline and that portion of the T-junction parallel to this feedline. The integration of these structures for Magic-T realization is only possible if both can

operate independently from one another. In other words isolation between the input ports of the "Magic-T" must be present.

Figure 3.15 The microstrip "Magic-T"

The isolation between structures is achieved by orienting the aperture in the ground plane such that its major dimension is perpendicular to the feedline. In this manner the input signal applied to port 1 will couple through the aperture onto the microstrip line parallel to the feedline and appear at ports 3 and 4 180 degrees out of phase and not port 2. Likewise an input signal applied at port 2 will reach ports 3 and 4 in phase but not appear at port 1.

The coupling from the feedline to the upper microstrip line can be attrib-

uted to a magnetic dipole (M_x) radiating in the presence of an electric wall that replaces the aperture as was illustrated in the previous section. This dipole due to the magnetic field component in the feedline parallel to the longer dimension of the slot radiates a field distribution in the upper microstrip line that propagates along the z-direction. For the coupled field distribution to propagate along the x-direction to port 2 requires a magnetic dipole (M_z) radiating in the vicinity of the slot. Since the magnetic field and the slot length in the z-direction is small, this magnetic dipole will not be a significant radiator and field coupling will not occur. Similarly a signal applied at port 2 will have a weak magnetic field tangential to the long dimension of the slot and a magnetic field tangential to the shorter dimension. Therefore the resulting magnetic dipoles would not be strong enough to radiate fields into the feedline.

3.14 Introduction To Numerical Methods

The design of the proposed aperture-coupled parallel line structure will require a means to analyze it and adjust tunable parameters to achieve the desired results. Due to the nature of this structure, it is not feasible or practical to adjust these parameters once it is built. Features like slot size cannot be modified accurately or easily and to examine its effect on performance would require the fabrication of several structures with different aperture sizes. Also this structure, like many planar integrated circuits, is not amenable to closed-form expressions that can be used to characterize performance and therefore simulation using numerical methods is needed.

Many different numerical methods exist that can be implemented to solve electromagnetic problems such as that of the aperture coupled structure. Numerical methods can be categorized into three different types commonly referred to as differential, integral, and ray methods. The choice of which one to use is based on trade-offs between accuracy, speed, storage requirements, versatility, and data preprocessing.

All numerical methods proceed to solve the electromagnetic behaviour of a structure in the same manner. Using Maxwell's equations, expressions for a harmonic solution based on a set of differential or integral equations is obtained that reflect the boundary conditions of the particular problem. These equations are processed analytically as far as possible and then the numerical method is used to determine the final solution. In many instances employing the numerical method might require considerable preprocessing before obtaining a computationally efficient program. An overview of the various numerical methods employed in the solution of electromagnetic field distribution is available in [10]. The transmission-line-matrix method (TLM) was selected as the means to design the aperture coupled structure.

3.14.1 The Transmission-Line-Matrix Method

The basic operational principle behind the TLM method has its origin in Huygen's model of wave propagation [1]. According to this model a wavefront consists of secondary radiators which give rise to spherical wavelets. The next wavefront is then formed by the envelope of these generated spherical wavelets and this process repeats itself continuously. This description of wave behaviour provides an accurate representation of the propagation and scattering of electromagnetic radiation. For this reason the transmission line method has been used successively in the analysis of two dimensional electromagnetic field problems and has been extended to solve electromagnetic field problems in three dimensions.

Johns and Beurle [2] were the first to introduce a numerical technique that applied Huygen's model in the evaluation of two-dimensional electromagnetic field problems in the time domain. Figure 3.16 portrays the discrete implementation of Huygen's wave model in two-dimensional homogeneous space as it applies to electromagnetic propagation. This region is defined by a mesh of interconnecting nodes spaced equidistantly

from one another by an amount ΔL .

Figure 3.16 Pulse propagation in a TLM mesh [4]

The nodes are joined together by identical orthogonal transmission lines which are considered dispersiveless and lossless. The basic building block is the shunt node which consists of four sections of transmission line, each of length $\frac{\Delta L}{2}$. A series node representation is also possible [3]. Figure 3.17 illustrates the shunt node and its equivalent lumped element

model where L and C are the inductance and capacitance per unit length.

Figure 3.17 The TLM shunt node and its equivalent model

Let us examine what occurs when an excitation is introduced at a particular node of the TLM mesh in the form of a Dirac voltage impulse. This impulse will generate pulses which expand outwards in all four directions with equal energy as illustrated in Figure 3.16a. These pulses will eventually reach adjacent nodes after an elapsed time Δt , Figure 3.16b. These nodes can be considered to be secondary radiators that lead to the generation of new pulses that scatter isotropically in all four directions as shown. Each scattered pulse will carry $1/4$ of the incident energy or $1/2$ the magnitude of the corresponding field quantity as shown. The scattered pulses will eventually reach other nodes after $2\Delta t$, Figure 3.16c, where this chain of events is repeated. The pulses emanating from several such radiators propagate along combining destructively and constructively with others to form the overall waveform.

At any one node the reflected voltage pulse from one node at time $(k + 1) \Delta t$ can be evaluated if the incident voltage pulse at the same node at time $k\Delta t$ is known. For a shunt node this relationship between incident and reflected voltages is as follows [4]

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_{(k+1)\Delta t}^{REF} = \frac{1}{2} \begin{bmatrix} -1 & 1 & 1 & 1 \\ 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \end{bmatrix} \cdot \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_{k\Delta t}^{INC} \quad (3.40)$$

Using the above relation and the fact that the reflected voltage pulse from one node is the incident voltage pulse to another, the voltage at any one node at any time can be evaluated. Applying Kirchhoff's laws, the variations of circuit quantities (currents I_X , I_Z and voltage V_Y in Figure 3.17) with respect to time and space can be described by two-dimensional wave equations. Maxwell's equations relating electromagnetic fields can also be described in terms of two-dimensional wave equations. Therefore an equivalence between circuit and field quantities can be established [3], [5] that can lead to the evaluation of electromagnetic fields in the region under consideration. For handling cases where different materials, losses, and boundary conditions are present various techniques are employed that are described in more detail in [4], [9].

In 1975, Akhartzad and Johns extended the TLM method to the evaluation of electromagnetic fields in three dimensions [7]. In this method the shunt node (or series node) used in the two-dimensional TLM method was replaced by a hybrid TLM cell consisting of three shunt and three series nodes. This expanded node implementation as it was called was used to describe all six field components (E_X , E_Y , E_Z , H_X , H_Y , H_Z) simultaneously. The nature of this structure was such that field components were physically separated which led to difficulties in applying boundary conditions. To alleviate this problem condensed node models, the asymmetrical condensed node [9] and the symmetrical condensed node [8], were introduced where all field components are evaluated at the same point.

The symmetrical condensed node TLM model is the most widely used TLM method in the evaluation of three dimensional field problems. In this method a TLM

cell is represented by a node consisting of twelve ports to represent two polarizations in each coordinate direction. The voltage pulses associated with these polarizations travel along twelve transmission lines that make up the node. The node properties are not described by equivalent lumped element circuits but rather are derived from general energy and charge conservation principles. The excitation and evaluation of the impulse response follows the same procedure as in the case for two dimensional TLM methods.

The three-dimensional symmetrical condensed node TLM method as implemented at CRC [14] was used to simulate the aperture-coupled structure. In this implementation the substrates, the aperture, the ground plane, and the free space region above each substrate are discretized. The conducting portions of the transmission lines are considered to be perfectly conducting and infinitely thin. Substrate permittivities and permeabilities are modelled by specifying appropriate stub values [6]. Excitation is provided by a Gaussian pulsed TEM voltage source. The dominant quasi-TEM microstrip mode evolves from this ideal TEM excitation. Other excitation schemes are applicable [7].

3.15 The Design Of The Aperture-Coupled Line Structure

An aperture-coupled line structure (Figure 3.10) will be used to feed an RF signal centred at 20 GHz with a bandwidth of 1 GHz into the mixer. Therefore over this frequency span this aperture-coupled line structure must be designed and developed to achieve optimum E-plane performance. This will require that the input signal is effectively coupled from the lower transmission line to the upper transmission line where its power can split evenly and 180 degrees out of phase into both output lines.

In order to maximize this coupling, microstrip transmission lines will be needed that have a high dielectric constant and a small substrate height [44]. For this reason, transmission lines were realized on a 254 μm thick alumina substrate with a dielectric constant of 9.8. This material was readily available and commonly used in the circuit

fabrication facilities at CRC, thus providing further incentive for its use. Other transmission line parameters such as line width and conductor thickness are not as significant in improving coupling.

The choice of aperture dimensions and geometry also affect E-plane performance. There has been much research into various aperture shapes and sizes [40]. The simplicity of its design and good performance make the rectangular slot one of the more attractive and commonly used configurations for aperture coupling. By making the slot width very narrow ($\lambda/10$), the dominant radiation mechanism is a y-directed magnetic current. This leads to the required 180 degree phase shift in signals travelling on either side of the slot away from the aperture as discussed previously. The realization of this property with other shapes is not readily apparent and would entail further theoretical and experimental investigation which is not within the scope of this thesis. Having fixed the slot width, the slot length becomes the only parameter which can be adjusted to improve coupling. For these reasons the rectangular slot was chosen as the aperture to be used in the realization of the E-plane tee.

The use of matching circuitry can also be employed to improve coupling. Minimization of reflections of the input signal along the feedline ensures that more of the signal can find its way to the aperture. In view of this, transmission lines will be designed with a characteristic impedance of 50 ohms to match that of the RF source. Matching the discontinuity presented by the aperture to the input transmission lines is a somewhat complicated undertaking and will not be attempted if the reflections are minimal.

Eventually the feedline must come to an end and the question that arises is what termination should it see. A termination of 50 ohms (same characteristic impedance of feedline) would absorb any incident energy that does not get coupled through the aperture but continues to propagate down the line. Other impedances would reflect a portion of this signal back onto the aperture and absorb the rest. Terminating this line such that all of the signal is reflected back onto the aperture would present another opportunity for the

signal to couple through the aperture. An open circuit or a short circuit termination could perform this task. An open circuit termination is more realizable and is the one that is implemented here. Therefore the length of this open circuit stub becomes another parameter with which to improve coupling.

3.15.1 TLM Simulation Of The Aperture-Coupled Lines

TLM simulations were performed from 2 GHz to 40 GHz in steps of 200 MHz for the combinations of aperture lengths and open circuit stub lengths listed in Table 3.1. These results can be found in Appendix C. The lengths in Table 3.1 are expressed in terms of wavelength (λ) evaluated at the operating frequency (f) of 20 GHz and effective dielectric constant (ϵ_{eff}) of 6.69 using equation (3.41)

$$\lambda = \frac{c}{f\sqrt{\epsilon_{eff}}} \quad (3.41)$$

to give $\lambda = 5800\mu m$.

Table 3.1 Aperture Length And Stub Length Combinations

Aperture Length	Open Circuit Stub Length
0.2λ	$0.1\lambda, 0.25\lambda, 0.3\lambda, 0.35\lambda, 0.4\lambda$
0.3λ	$0.1\lambda, 0.15\lambda, 0.2\lambda, 0.25\lambda, 0.3\lambda, 0.35\lambda, 0.4\lambda$
0.4λ	$0.1\lambda, 0.15\lambda, 0.175\lambda, 0.2\lambda, 0.25\lambda, 0.3\lambda, 0.35\lambda, 0.4\lambda, 0.5\lambda$
0.45λ	$0.25\lambda, 0.3\lambda$

After examining the results obtained from the above combinations (Appendix C), the following observations were made. The general trends of the S_{11} and S_{21} traces for fixed aperture length size for the various combinations of stub lengths were for the most part consistent. The only exception was for combinations where the aperture length was 0.2λ and stub lengths were less than 0.25λ , Figure 1.1 and Figure 1.2 in Appendix C. For fixed aperture lengths greater than 0.2λ , increasing the stub length did not change the general shape of the S_{11} and S_{21} traces drastically but did shift them downwards in terms of frequency, Figures 1.3 - 1.14 of Appendix C.

Figures 1.15-1.24 in Appendix C also display the effects of aperture length size for fixed stub length. In general S_{11} curves displayed the same tendencies for a given stub length for the various aperture sizes. Increasing the aperture size improved the return loss between 200 MHz and 25 GHz. Likewise, S_{21} traces were very similar for a given stub length for the various aperture sizes used. The S_{21} response improved as the aperture length was made larger.

For the aperture-coupled line structure to operate effectively, the input reflection loss (S_{11}) has to be as low as possible and the coupling through the aperture (S_{21} , S_{31}) should approach the theoretical maximum of -3 dB. Figure 3.18 and Figure 3.19 illustrate that for an aperture length of 0.4λ and a stub length in the range of 0.1λ to 0.2λ S_{11} is less than -10 dB and S_{21} is slightly more than -4 dB at 20 GHz. Looking at the case for an aperture length of 0.45λ , Figures 1.13-1.14 in Appendix C, results suggest that S_{11} and S_{21} might be improved if shorter stub lengths could be used. Therefore aperture-coupled lines were designed using aperture lengths of 0.4λ and 0.45λ for a variety of stub lengths.

Figure 3.18 Simulated S_{11} for 0.4λ aperture length at various stub lengths

Figure 3.19 Simulated S_{21} for 0.4λ aperture length at various stub lengths

3.16 Design Of The Microstrip T-Junction

The microstrip T-junction will be used to feed the local oscillator (LO) signal centred at 18.5 GHz into the mixer circuit. It is the microstrip counterpart of the waveguide H-plane tee and will be designed to operate as such over this operating frequency. The design will seek to maximize the power split and maintain zero phase difference from the input port to the two output ports.

The final structure as optimized in Libra [13] to realize the T-junction is illustrated in Figure 3.20. Since a portion of the T-junction is to be used as the top transmission line of the aperture coupled structure, it was designed on an alumina substrate 254 μm thick with a dielectric constant of 9.8.

Figure 3.20 Planar profile of the designed microstrip T-junction

To accommodate the input and output connectors of the "Magic-T" structure, the input line (port 1) was made 7500 microns long, the distance from centre to centre between the outputs (ports 2 and 3) was fixed at about 15000 microns, and the length of the output lines from the centre of the bends was set at approximately 7500 microns. Sharp corners

lead to radiation losses and therefore chamfered bends were used to minimize this effect. Finally, the line widths for the input line and the other lines were made 487 microns and 256 microns respectively. More details concerning the structure of this microstrip T-junction can be found in Appendix D.

The return loss and insertion loss of the microstrip T-junction are displayed in Figure 3.21 and Figure 3.22 respectively. The insertion loss was less than 3.5 dB over the frequency span 17.5 GHz to 19.5 GHz and the return loss was better than 20 dB over the frequency range 18.0 GHz to 19.0 GHz. Also there was no phase difference between the output arms over these frequency bands. Therefore the microstrip T-junction could operate quite effectively over a fairly wide frequency range. If the fabricated structure's response suffers a frequency translation upwards or downwards of 500 MHz or less from simulation, we can be assured that the microstrip T-junction will still operate effectively at 18.5 GHz.

Figure 3.21 Simulated S_{11} response of the designed microstrip T-junction

Figure 3.22 Simulated S_{21} response of the designed microstrip T-junction

3.17 Fabrication And Assembly

The construction of the microstrip "Magic-T" required several steps. First each of the three layers, Figure 3.15, were laid out using a graphics design package, DW-2000 [31]. In the layout, provisions for aligning the three layers were made. Markers were placed on the top layer such that they could be seen from the bottom side for aligning the slot layer. The slot layer also contained markers that were used as edge guidelines for placing the feedline circuit. These three layers were then used to generate photoplots which were in turn used to create mask patterns on glass plates. The next step implemented photolithographic techniques to fabricate the T-branch and feedline circuit from the mask patterns on the glass plates. Once this was completed the back metallization of the feedline was removed and that of the T-branch was patterned out using the slot layer. The feedline was then aligned within the edge markers on the ground plane of the T-branch and held in place with epoxy. The complete structure was mounted in a brass test

jig with appropriate connectors. The jig assembly was very similar to the one used in Figure 5.6 to mount the single-balanced diode mixer circuit.

3.18 Hybrid Coupler Measurements

The following graphs illustrate the measured results obtained for the 180-degree hybrid coupler as realized using the microstrip "Magic-T". The measurements were performed on a network analyzer using the same setup as discussed in section 2.12.3. However coaxial cables and the test jig described earlier were used instead of the Wiltron test fixture. The coaxial cables were de-embedded from measurements using an open/short/load instrument calibration. A connector model was used in LibraTM to de-embed the effects of the connectors from measurement also. Details can be found in Appendix E.

Figure 3.23 displays the LO return loss versus frequency. The LO is centred at 18.5 GHz and therefore it is desirable for the return loss to be as low as possible at this frequency. Looking at Figure 3.23 the return loss at 18.5 GHz is about -14 dB and the best return loss is roughly -19 dB at a frequency of 19 GHz. Recalling the simulations of the T-branch in section 3.16 (Figure 3.21), the return loss at 18.5 GHz was approximately 30 dB. Not only was the measured return loss much higher than simulation but the shape of the return loss versus frequency was quite different.

The LO insertion loss between the input port and each of the output arms of the microstrip "Magic-T" is shown in Figure 3.24. Over the span 17.5 to 19.5 GHz the insertion loss was better than 3.5 dB for one output arm. The insertion loss over the same span was not quite as good through the other output arm. The insertion loss through the other arm was as much as 0.5 dB less even though the insertion traces were very similar. For the most part, measurement followed the same general trend as simulation which predicted roughly 3.5 dB loss over the same frequency band.

Figure 3.23 Measured LO return loss for the microstrip "Magic-T"

Figure 3.24 Measured LO insertion loss for the microstrip "Magic-T"

The discrepancies between measurement and simulation can be attributed to several factors. First the LibraTM microstrip elements used to model the T-junction do not take into account all losses over the high frequency of operation. In hindsight a more accurate description of performance would require a full electromagnetic analysis. Also, the connector de-embedding, a necessity, was an approximation and not a standard method. Finally, connector non-uniformity and cable re-positioning from the calibrated position can attribute to the observed discrepancies. This last point is also a viable explanation as to why measured insertion losses for each output arm were different.

For effective "Magic-T" operation the LO signal must appear in phase at both output arms. Looking at Figure 3.25, there is a slight difference between the phase at each output arm. However, this difference is no greater than 8 degrees at any one point and in many cases better than 3 degrees.

The RF signal will be injected into the mixer at 20 GHz with a bandwidth as wide as possible. Therefore the RF return loss of the hybrid coupler over this span should be as low as possible. Looking at Figure 3.26 the return loss at 19.5, 20, and 20.5 GHz is approximately -10, -19, and -27 dB respectively at the RF input port. A better return loss over a 1 GHz bandwidth is obtained for a signal centred at 20.4 GHz. Over this band the return loss is below -18 dB and at the centre frequency the return loss is -28 dB. Therefore if the RF return loss trace was shifted left by about 0.4 GHz the return loss would reach optimal values. As it stands the RF return loss is acceptable.

Ideally the input RF signal power should divide evenly into both output ports over the frequency band of interest. However, realistic losses make this power division less than ideal. Looking at Figure 3.27 the measured RF insertion loss over 15 to 25 GHz is displayed. At 20 GHz the loss is about 4 dB and 4.4 dB at the band edges of 19.5 and 20.5 GHz. The insertion loss over both output ports is very similar in both shape and magnitude. The graph indicates that the best insertion loss is at the operating point designed for.

The RF phase difference between both output arms should be 180 degrees over the operating band. Figure 3.28 displays the phase of the RF signal between the input port and output ports. Examining both these traces, the phase difference between them is about 186 degrees over the frequency band of interest.

Figure 3.25 Measured LO phase at both output arms of the microstrip "Magic-T"

Figure 3.26 Measured RF return loss for the microstrip "Magic-T"

Figure 3.27 Measured RF insertion loss for the microstrip "Magic-T"

Figure 3.28 Measured RF phase at both output arms of the microstrip "Magic-T"

In mixer operation, as much of the LO and RF signals as possible should appear across the terminals of the mixing element. This entails that very little of the RF and LO signals should find their way to each other's input port. The hybrid coupler is the only component in the single-balanced mixer that will provide this isolation which should be very high. Figure 3.29 displays the measured isolation between the RF and LO input port. Over the RF operating band of 19.5 to 20.5 GHz and LO frequency of 18.5 GHz this isolation is below -30 dB. Therefore both input ports are well isolated from one another at both the RF and LO frequencies.

Figure 3.29 Measured RF- LO isolation for the "Magic-T"

CHAPTER 4: FILTER NETWORKS

4.1 Introduction

Mixer circuits produce a multitude of signals at various mixing frequencies. However, typical applications are only concerned with a specific signal and thus a means is necessary to obtain the desired signal and reject unwanted ones. Filter circuits exist that can fulfill this task. They consist of lumped elements (inductors, capacitors, and resistors), distributed elements (transmission media such as waveguides or microstrip), or a combination of both arranged in various configurations. These configurations are implemented to realize the four type of filters commonly in use. These are referred to as low-pass, high-pass, bandpass, and bandstop filters. Each type is designed so as to provide a frequency band (passband) over which desired signals experience little attenuation and a region outside this band (stopband) where unwanted signals are attenuated. The following figure illustrates the frequency response of each of these filter types.

Figure 4.1 Filter types: (a) low-pass, (b) high-pass, (c) bandpass, (d) bandstop [17]

4.2 Low-Pass Filter Design Method

The most popular method used in the design of filters is known as low-pass prototype filter synthesis [17]. In this method, the characteristic frequencies associated with the desired filter response (low-pass, high-pass, bandpass, and bandstop) are transformed to that of a prototype low-pass filter (Figure 4.2), using readily available relations. The transformation results in a normalized frequency that can be used together with the specified insertion loss in the passband and stopband to determine the number of sections and the element values required to design the low-pass prototype filter. Finally, the network is realized in terms of lumped and/or distributed circuit elements. This technique will be applied in the design of a low-pass filter for selecting the desired frequency component $\omega_{IF} = \omega_{LO} \pm \omega_{RF}$ from the myriad of signals generated at the output of the single-balanced diode mixer..

Figure 4.2 Low-pass prototype filter of n sections [17]

The realization of the low-pass prototype is based on one of two commonly used types of low-pass transfer function representations, (Figure 4.3). These representations are known as the maximally flat (Butterworth) and the equal ripple (Chebyshev) responses [41]. Both are well characterized and published tables and graphs exist that can be used to determine the required number of sections and the corresponding element values for a given response. At microwave frequencies the Chebyshev response is the more widely used. In this case, the desired attenuation (A_m) in the passband, the passband fre-

quency edge (ω'_1), and the insertion loss at the stopband frequency (ω'_L) are specified and the required number of low-pass prototype sections (n) as well as the necessary element values (g_k) can then be determined.

Figure 4.3 Butterworth (a) and Chebyshev (b) low-pass transfer functions [17]

The next step is to realize the filter using lumped and/or distributed components. At microwave frequencies it is more practical to design filters using distributed elements. However, in order to do so the lumped equivalent series inductances (L_k) and shunt capacitances (C_k) [17] must first be evaluated using the following equations

$$L_k = g_k \left(\frac{Z_0}{\omega_1} \right) \quad (4.1)$$

$$C_k = g_k \left(\frac{1}{\omega_1 Z_0} \right) \quad (4.2)$$

where Z_0 is the characteristic impedance and $k = 0, 1, 2, \dots, n + 1$. Once these values have been obtained a distributed network can then be designed. As indicated in [37] short lengths of high impedance microstrip lines and low impedance microstrip lines behave similarly to series inductances and shunt capacitances respectively. Therefore implement-

ing cascaded sections of short alternating lengths of high and low impedance microstrip lines as illustrated in Figure 4.4 can be used to realize a low-pass filter. The necessary lengths of high impedance and low impedance line required to realize a given series inductance (l_L) and shunt capacitance (l_C) are stated as follows

$$l_L = \frac{\lambda_{gL}}{2\pi} \text{asin}\left(\frac{\omega L}{Z_{0L}}\right) \quad (4.3)$$

$$l_C = \frac{\lambda_{gC}}{2\pi} \text{asin}(\omega C Z_{0C}) . \quad (4.4)$$

The quantities (λ_{gL}) and (λ_{gC}) are the wavelengths of the inductive and capacitive lines respectively evaluated at the passband frequency. (Z_{0C}) and (Z_{0L}) are the characteristic impedances of these lines. The high impedance and low impedance lines contain end capacitive and inductive effects respectively. The transition from wider (low impedance) to thinner (high impedance) lines has an associated step capacitance as well. Therefore, the required inductance (L) and capacitance (C) are overstated and should take these effects into account. It has been shown [37] that the end inductance effects are negligible and can be ignored, however the same cannot be said for the remaining effects. The end and step capacitances (C_{end}) and (C_{step}) at each end of the lines can be determined using the following equations

$$C_{end} = \frac{1}{2\pi f Z_{0L}} \tan\left(\frac{\pi l_L}{\lambda_{gL}}\right) \quad (4.5)$$

$$\frac{C_{step}}{h} = 1370 \frac{\sqrt{\epsilon_{eff}}}{Z_0} \left(1 - \frac{W_2}{W_1}\right) \left(\frac{\epsilon_{eff} + 0.3}{\epsilon_{eff} - 0.258}\right) \left(\frac{W_1/h + 0.264}{W_1/h + 0.8}\right) \left(\frac{pf}{m}\right) \quad (4.6)$$

where (W_2) is the width of the wider line and all the other parameters (substrate height (h), effective dielectric constant (ϵ_{eff}), and the characteristic impedance (Z_0) are

those associated with this line. Once all the elements have been determined, the final circuit should be entered into a numerical simulator to tune the elements to achieve a more accurate response.

Figure 4.4 Lumped element (a) and distributed (b) low pass filter [37]

4.2.1 The LowPass Filter Design

For our purposes a low-pass filter with a passband attenuation (A_m) of 0.1 dB, a passband edge (ω'_1) of 3 GHz and an insertion loss (IL) of 10 dB at a stopband frequency (ω'_L) of 4 GHz is required. Since a low-pass filter is to be designed using the low-pass prototype, the transformation of the characteristic frequencies is not required as would be the case if any of the other types of filter responses were to be realized. Using

the Chebyshev response relating insertion loss to normalized frequency $\left(\left|\frac{\omega_1}{\omega_L}\right| - 1\right)$ for a passband ripple of 0.1dB [17], the number of sections required for the desired response was determined to be five. Once the number of sections were obtained, tables as given in [17] were used to evaluate the prototype element values. These were determined to be as follows $g_0 = g_6 = 1$, $g_1 = g_5 = 1.1468$, $g_2 = g_4 = 1.3712$, and $g_3 = 1.9750$. Using equations (4.1) and (4.2), values for the corresponding lumped equivalent series inductances and shunt capacitances were obtained and further optimized through LibraTM to yield $L_2 = L_4 = 3.04$ nH, $C_1 = C_5 = 0.79$ pf, and $C_3 = 1.51$ pf.

Once the lumped element equivalents were determined, the design of the distributed low-pass filter was undertaken. Here the high impedance lines were chosen to have a characteristic impedance of approximately 90Ω which corresponded to a realizable width of $50 \mu m$. The low impedance lines were set at a characteristic impedance of 25Ω resulting in a width of $773 \mu m$. To determine the lengths required for the distributed inductances and capacitances (l_L and l_C), the effective dielectric constant (ϵ_{eff}) and the guided wavelengths (λ_g) for each line were needed at the passband edge frequency $\left(\frac{\omega_1}{2\pi}\right)$ of 2 GHz. Knowing the characteristic impedance (Z_0), operating frequency (2 GHz), and the dielectric constant (ϵ_r), a software program LinecalcTM was used to find these unknown quantities.

The lengths required for the distributed inductances were obtained using equation (4.3). Then the end and step capacitances associated with these lengths were calculated as per equations (4.5) and (4.6), subtracted from the required capacitance, and this value was then used in equation (4.4) to solve for the lengths of the distributed capacitive elements. Using LibraTM, these microstrip lengths were optimized to improve the low-pass response. The final values for the widths and lengths in microns as defined in Figure 4.4 used to realize the distributed elements were as follows $W_{C1} = 773$, $W_{C2} = 773$, $W_{C3} = 773$, $l_1 = 1500$, $l_3 = 3870$, $l_5 = 1500$, $W_{L1} = 50$, $W_{L2} = 50$, $l_2 = 4385$, $l_4 = 4385$. More circuit details can be found in the Appendix F.

The simulated low-pass response of the designed distributed low-pass filter using LibraTM is shown in the following figures. LibraTM has a built in electromagnetic simulator that is valid for simulating microstrip lines and discontinuities. It was the tool used to simulate the low-pass filter over the IF band and RF band, Figure 4.5 and Figure 4.6 respectively. Since this structure is very well characterized in LibraTM, efforts were not made to fabricate and test it independently from the single-balanced mixer.

Referring to Figure 4.5, the low-pass filter allows the IF signal at 1.5 GHz to pass through with very little reflection and attenuation. At the same time, as shown in Figure 4.6, the RF and LO signals at 20 GHz and 18.5 GHz respectively are reflected to the extent that very little of these signals pass through.

Figure 4.5 Simulated S_{11} and S_{21} response of the low-pass filter over the IF band

Figure 4.6 Simulated S₁₁ and S₂₁ response for the low-pass filter over the RF band

CHAPTER 5: THE SINGLE-BALANCED DIODE MIXER

5.1 Introduction

Mixer circuits based on single-diode designs find use in millimeter frequency band applications. However at microwave frequencies, where conversion loss is not a critical parameter, mixer circuits that use a combination of a balanced number of diodes are more common. Besides not requiring extra circuitry to inject the LO signal into the mixer circuit, they offer better power handling and reject spurious responses and LO noise. However they require more LO power to drive the increased number of diodes and generally have poorer conversion loss as compared to single-diode mixers. Figure 5.1 illustrates the various balanced circuits that are employed in microwave mixers and Table 5.1 and Table 5.2 [11] compares their performance with that of the single diode mixer.

Figure 5.1 Typical balanced mixer circuits [11]

Table 5.1 Mixer Performance Comparison

Mixer Type	VSWR RF	VSWR LO	VSWR IF	Isolation RF-IF	Isolation LO-RF	Isolation LO-IF
Single diode	Depends on matching	Depends on matching	Depends on matching	Depends on filters	Depends on filters	Depends on filters
Singly balanced (180 deg)	Depends on matching	Depends on matching	Depends on matching	Depends on filters	Good	Depends on filters
Singly balanced (90 deg)	Good	Good	Depends on matching	Depends on filters	Poor	Depends on filters
Doubly balanced (ring/star)	Good	Good	Good	Good	Good	Good
Subharmonically pumped	Depends on matching	Depends on matching	Depends on matching	Depends on filters	Good	Depends on filters
Image Rejection	Good	Good	Good	Good	Good	Good

Table 5.2 Mixer Performance Comparison (Continued)

Mixer Type	LO AM Noise Rejection	LO Spurious Signal Rejection	Low Order Spurious Response Rejection	LO Power Required	Third Order IM Intercept	1-dB Compression
Single diode	None	None	None	Low	Low	Low
Singly balanced (180 deg)	Good	Good	(2,2):good (2,1):good if LO at Δ port (1,2):good if LO at Σ port	Moderate	Moderate	Moderate
Singly balanced (90 deg)	Good	Good	(2,2):good (2,1):none (1,2):none	Moderate	Moderate	Moderate
Doubly balanced (ring/star)	Good	Good	(2,2):good (2,1):good (1,2):good	High	High	High
Subharmonically pumped	Good	Good	Mixing with odd LO harmonics	Moderate	Low	Low
Image Rejection	Good	Good	Same as above	High to very high	High	High

5.2 Theory Of Operation

A single-balanced diode mixer will be used to implement the "Magic-T" hybrid coupler into a working mixer circuit and will thus be the focus of attention. Figure 5.2 will be used to illustrate the operation of the 180-degree single-balanced diode mixer. Here the LO signal is applied to the delta (Δ) port of the hybrid which leads to an LO voltage appearing in phase across one set of diode terminals and 180 degrees out of phase across the other. The RF signal is applied to the hybrid's sigma (Σ) port and will appear in phase across the terminals of each diode.

Figure 5.2 Single-balanced diode mixer operation [11]

The resulting currents and voltages across the diodes in the 180-degree balanced diode mixer are shown in the above figure. The voltages V_2 and V_1 are the summation and difference of the LO and RF signals respectively at the output terminals of the hybrid and are given by equations (5.1) and (5.2). The diode currents I_2 and I_1 are related to these voltages by polynomial expressions as outlined in Section 2.2 and restated in equations (5.3) and (5.4). The resulting current from the output port of the mixer will then be $I_{IF} = I_1 - I_2$.

$$V_2 = V_L \cos(\omega_{LO}t) + V_{RF} \cos(\omega_{RF}t) \quad (5.1)$$

$$V_1 = -V_L \cos(\omega_{LO}t) + V_{RF} \cos(\omega_{RF}t) \quad (5.2)$$

$$I_2 = -aV_2 + bV_2^2 - cV_2^3 + dV_2^4 + \dots \quad (5.3)$$

$$I_1 = aV_1 + bV_1^2 + cV_1^3 + dV_1^4 + \dots \quad (5.4)$$

Substituting V_1 and V_2 into the current equations for I_1 and I_2 and using the trigonometric identities outlined in section 2.1 shows that the resultant current I_{IF} consists of many frequency components. These mixing frequencies are defined as $f_{IF} = mf_{RF} + nf_{LO}$, where $m + n = k$ and k is the order of the polynomial in equations (5.3) and (5.4). Spurious responses where m is even and n is even or odd are eliminated. Also AM noise and spurious signals on the LO are cancelled in the ideal 180-degree balanced mixer. Applying the LO and RF signals into the (Σ) port and (Δ) port of the hybrid leads to the same operation as described except that spurious responses where n is even and m is even or odd are eliminated instead. More detailed explanation of mixer operation can be found in [42], [43].

5.3 Mixer Analysis: Harmonic Balance

Mixer circuits operate under the influence of both a large (LO) signal and a small (RF) signal. Therefore to fully characterize their performance, mixer circuits must be analyzed under both operating conditions. For ease of analysis, multiple diode mixer circuits are often simplified to single-diode mixer equivalents. With this in mind a single-diode mixer will be used to illustrate the general theory behind the application of the harmonic balance technique. In a diode mixer circuit, the purpose of the LO signal is to vary the small-signal junction conductance and capacitance with respect to time. The RF signal is responsible for the frequency conversion that generates the small-signal mixing spectrum via these linear time-varying small-signal elements.

The most common means of analysis is to employ what is referred to as the harmonic balance technique. The circuit is first analyzed under large-signal conditions to generate a harmonically balanced system of equations. In the case of diode mixers these equations are used to determine the large-signal junction voltage across each diode. The large-signal junction voltage is then used to determine the time-varying conductances and capacitances. Once these are determined a small-signal analysis can then be performed resulting in a set of harmonically balanced equations that describe the conversion properties of the mixer.

5.3.1 Large-Signal Analysis

The equivalent circuit of a single-diode mixer under large-signal excitation is illustrated in Figure 5.3. It consists of a single-diode, matching circuitry, source impedance (Z_S), and impedance terminations (Z_{IF}) and (Z_{RF}) at the IF and RF ports respectively. The large signal (LO) excitation is represented as a voltage source denoted (V_S). The small signal (RF) excitation is ignored under large signal analysis.

Figure 5.3 Equivalent diode mixer circuit

The equivalent circuit in the above figure can be reduced to a two port network by absorbing the source and impedance terminations into a linear subcircuit. The diode can be represented by its constituent components as displayed in Figure 5.4. The linear elements of the diode can also be absorbed into the subcircuit. However, the junction conductance and capacitance being nonlinear cannot be included in the linear circuit. Figure 5.4 illustrates this circuit representation of the single-diode mixer where an admittance matrix has been used to characterize the linear subcircuit.

Figure 5.4 Two port representation of single-diode mixer

Application of an LO signal to the above circuit will generate a junction voltage (V_j), junction current (I_j), and currents (I_1) and (I_2) made up of harmonic components of the fundamental LO frequency. Therefore these voltages and currents can be expressed as a summation of phasor components as follows

$$V_j = \sum_{n=-\infty}^{n=\infty} V_{jn} e^{j(n\omega_p)t}, \quad (5.5)$$

$$I_d = \sum_{n=-\infty}^{n=\infty} I_{dn} e^{j(n\omega_p)t}, \quad (5.6)$$

$$I_1, I_2 = \sum_{n=-\infty}^{n=\infty} I_{1n}, I_{2n} e^{j(n\omega_p)t}. \quad (5.7)$$

The admittance matrix representing the linear portion of the diode mixer relates port currents to port voltages at each LO harmonic ($n\omega_p$) as follows

$$\begin{bmatrix} I_{1n} \\ I_{2n} \end{bmatrix} = \begin{bmatrix} Y_{11n} & Y_{12n} \\ Y_{21n} & Y_{22n} \end{bmatrix} \begin{bmatrix} V_{jn} \\ V_{sn} \end{bmatrix}. \quad (5.8)$$

Therefore the current into port 1 is defined as

$$I_{1n} = Y_{11n} \bullet V_{jn} + Y_{12n} \bullet V_{sn} \quad (5.9)$$

where the components (Y_{11n}) are referred to as the embedding admittances and represent the admittance at each LO harmonic as seen from the diode terminals. The current (I_d) is that which flows through the diode and is the summation of the currents across the junction capacitance (I_c) and conductance (I_g). The currents (I_c) and (I_g) are given by equations (2.20) and (2.19) respectively. The diode current at each LO harmonic is thus represented as

$$I_{dn} = I_{gn} + I_{cn}. \quad (5.10)$$

According to Kirchhoff's current law, the summation of currents at any one node must equal zero. Therefore the currents (I_1) and (I_d) must be such that they obey the following

$$I_{1n} + I_{dn} = 0. \quad (5.11)$$

The goal of the large-signal analysis is to find a set of harmonic voltage components (V_{jn}) that satisfy the above equation. An initial means of estimating (V_{jn}) and an algorithm that successively modifies it and tests it until equation (5.11) is very closely met over a finite

number of harmonics, $n = 0, \pm 1, \pm 2, \dots \pm N$, is required. The problem at hand lends itself to many algorithms. Some of the more common ones are the reflection algorithm, the splitting algorithm, and Newton's method [11]. The one chosen is the Newton Method as implemented in LibraTM.

5.3.2 Small-Signal Analysis

Application of a small RF signal voltage at a frequency (ω_s) to a diode under the influence of a large LO signal generates currents and voltages at many other frequencies. These new signals are termed the small-signal mixing components as shown in Figure 2.3 and whose frequencies are defined by equation (2.6). For ease of illustration the harmonic components of the applied RF signal are not considered ($m = 1$). These voltage and current components can be expressed as phasors (v_n) and (i_l) respectively at each mixing frequency. The total small-signal voltage (v) and current (i) components across the diode can be expressed as a summation of phasor components given by

$$v = \sum_{n=-\infty}^{n=\infty} v_n e^{j(\omega_s + n\omega_p)t} \quad (5.12)$$

$$i = \sum_{l=-\infty}^{l=\infty} i_l e^{j(\omega_s + l\omega_p)t} \quad (5.13)$$

respectively. The time-varying conductance (G_s) and capacitance (C_s) at each mixing frequency can be determined using equations (2.21) and (2.22) respectively and the large-signal junction voltage (V_{jn}) that was obtained under large-signal analysis. The time-varying conductance and capacitance can thus be expressed as

$$g = \sum_{s=-\infty}^{s=\infty} G_s e^{j(s\omega_p t)} \quad (5.14)$$

$$C = \sum_{s=-\infty}^{s=\infty} C_s e^{j(s\omega_p t)} \quad (5.15)$$

respectively. The relation between the harmonic amplitudes of the RF current (i_l) and voltage (v_n) across the diode junction in terms of the time-varying components (G_s) and (C_s) is given by

$$i_l = \sum_{s=-\infty}^{s=\infty} G_s \cdot v_n + \sum_{s=-\infty}^{s=\infty} \frac{dC_s}{dt} \cdot v_n + \sum_{s=-\infty}^{s=\infty} C_s \cdot \frac{dv_n}{dt}. \quad (5.16)$$

To solve for this expression, the number of harmonics must be limited to some value N such that $-N \leq l, n \leq N$ and the harmonics on either side of equation (5.16) must be balanced such that $l = s + n$. In matrix form and converting to the frequency domain equation (5.16) can be expressed as

$$\begin{bmatrix} i_{-N}^* \\ i_{-N+1}^* \\ \dots \\ i_N \end{bmatrix} = \begin{bmatrix} G_0 + \omega_{-N} C_0 & G_{-1} + \omega_{-N} C_{-1} & \dots & G_{-2N} + \omega_{-N} C_{-2N} \\ G_1 + \omega_{-N+1} C_1 & G_0 + \omega_{-N+1} C_0 & \dots & G_{-2N+1} + \omega_{-N+1} C_{-2N+1} \\ \dots & \dots & \dots & \dots \\ G_{2N} + \omega_N C_{2N} & G_{2N-1} + \omega_N C_{2N-1} & \dots & G_0 + \omega_N C_0 \end{bmatrix} \cdot \begin{bmatrix} v_{-N}^* \\ v_{-N+1}^* \\ \dots \\ v_N \end{bmatrix} \quad (5.17)$$

where * denotes terms with negative frequency components whose complex conjugates have been taken in order to facilitate equation handling.

Equation (5.17) is referred to as a conversion matrix and can be used to represent the diode as a multiport network as displayed in Figure 5.5. The ports of this network are not considered to be physical terminals but represent the voltage and current

components at different mixing frequencies as seen through the diode terminals. Each of these ports is terminated in an embedding impedance evaluated at its particular mixing frequency and a current source ($I_s(\omega_s)$) representing the RF excitation. The RF port and the IF port representing the desired mixing frequency are the only ones of concern since they are used in the evaluation of the mixer performance. With this in mind, the multiport network can be reduced to a two port (RF and IF) by absorbing the embedding admittances of the other ports into the conversion matrix as outlined in [11].

Figure 5.5 Multiport representation of single diode mixer

5.4 The Design Of The Single-Balanced Diode Mixer

The single-balanced mixer is made up of several components as illustrated in diagrams a) and b) of Figure 5.1. In this figure a hybrid coupler, mixing devices, and a low-pass filter are the main elements. In our case a 180-degree hybrid coupler was implemented using a microstrip "Magic-T" realization. The mixing element consists of an anti-parallel diode pair. The final element, the low pass filter, is based on cascaded sections of narrow and wide microstrip transmission lines. Besides these components, matching circuitry for the RF and LO signals is commonly utilized. Also, DC-IF blocks and return paths that shunt unwanted DC and IF signals to ground typically find a place in mixer circuits. An overview of the design and integration of all these components into a working mixer is the focus of this section. Figure 5.6, Figure 5.7, and Figure 5.8 illustrate the fabricated single-balanced diode mixer. Appendix G provides more information on its design.

5.4.1 The 180-Degree Hybrid Coupler: The Microstrip "Magic Tee"

The microstrip "Magic-T" was designed to fulfill the role of a 180-degree hybrid coupler. It was based on the union of a T-branch and aperture-coupled lines designed independently as outlined in sections 3.16 and 3.15. Since a well characterized element model did not exist, the "Magic-T" was constructed and measured. The scattering parameters as obtained from measurement were then used to represent the coupler in the overall single-balanced mixer design.

5.4.2 The Mixing Element: Schottky-Barrier Diode

A GaAs Schottky-barrier diode was chosen as the nonlinear device to be used as the mixing element. The diode is to be used under both small (RF) and large signal (LO) operating conditions where its response is quite different. Therefore small signal

S-parameter measurements cannot solely be used for its representation. A diode model that could describe its operation under both conditions was needed. Unlike the coupler, diode models are available that can be used to represent the diode. The diode's DC response and measured S-parameters were used in the parameter evaluation of the diode model. This diode model could then be used in the representation of each diode in the antiparallel combination required for the single balanced mixer.

5.4.3 The Low-Pass Filter: Microstrip Based Low-Pass Filter

A low-pass filter is needed to allow the transmission of the desired low frequency IF signal while at the same time suppressing the undesired higher frequency mixing components, the RF, and the LO signal. The designed low-pass filter was based on cascaded sections of microstrip transmission lines. The fundamental component of this filter is the well characterized microstrip transmission line element typically found in many electromagnetic simulators. For this reason it was deemed that the circuit representation of the designed low-pass filter provided enough accuracy to be used directly in the single-balanced mixer design. The other option was to construct the designed filter and use measured S-parameter data to represent the filter. However, this increased accuracy would require more time and effort than was necessary and the former option was chosen.

5.4.4 The DC-IF Block

Generated DC and IF harmonics are supposed to be directed towards the output of the mixer circuit. However, due to mismatches between the mixing elements and the output circuitry these harmonics undergo reflections and find their way back to the LO and RF signal side. A DC-IF block is needed to prevent these signals from reaching the LO and RF sources. At the same time the DC-IF block must not hinder the RF and LO signal propagation to the mixing elements. A series capacitor of appropriate value can

present a very low impedance path for RF and LO signals and at the same time a high impedance path for the DC and IF signals. A commercial parallel plate capacitor of 1 pF was chosen for the task. It had to be selected such that its series resonance frequency was much higher than the operating frequency to ensure that it did not behave as an inductor.

5.4.5 The DC-IF Return

The DC-IF harmonics that are blocked from reaching the RF and LO sources are reflected back onto the mixing elements (the anti-parallel diode pair). These harmonics should be shunted to ground rather than allowed to mix with the LO and RF signals to generate new harmonics. Therefore circuitry is commonly implemented to achieve this task without affecting the RF and LO signals that must appear at the diode terminals. In our case a piece of microstrip was connected between ground, by means of a via, and the input line feeding the RF and LO signal to the diodes. Theoretically, a transmission line terminated by a short circuit at one end can be translated into an open circuit at a specific frequency by making the line a quarter wavelength in length at the frequency of interest. The same approach was used in order to make the DC-IF return path transparent to the RF and LO frequencies. Although the microstrip line in the DC-IF return path cannot be made a quarter wavelength at both these frequencies, an intermediate length was used in order to minimize the effects of the DC-IF return path at the LO and RF frequencies.

5.4.6 The RF-Matching Network

In any mixer design it is important to ensure that as much of the RF and LO power reaches the mixing elements as possible. In this manner we avoid using excessive amounts of RF and LO power which in some applications tends to be at a premium. This requires that the points of reflection on the input side of the single balanced mixer be min-

imized at the RF and LO frequencies which necessitates the use of matching circuits. The RF and LO return loss at the input ports of the coupler were designed to be as low as possible and therefore no further matching was contemplated at this point. However, the same cannot be said about the point between the input side of the mixing elements and the output ports of the coupler. For minimal reflections, the input impedance at the diodes must be such that it presents the complex conjugate of the output impedance of the coupler to any signal travelling from the coupler to the diodes. Therefore, a single stub matching network was employed to transform the input impedance of the mixing elements to the complex conjugate of the output impedance of the coupler. Ideally, the matching network should be designed so as to minimize reflections at both the RF and LO frequency. However, due to the narrow-band nature of the matching network it is only possible to match to one frequency, typically the RF frequency. Therefore, the assumption is made that the frequency of the LO signal is close enough to that of the RF signal that it will benefit from the RF matching network.

5.4.7 The IF Combining Circuitry

It is also necessary to design circuitry to combine the IF signals from each mixing element to reach the output of the mixer. Microstrip lines and chamfered bends (to reduce signal discontinuity) were used to span the distance between each mixing element and direct the generated harmonics to the low-pass filter where the desired IF signal is to be obtained.

5.5 Single-Balanced Diode Mixer Fabrication And Assembly

The same process as implemented in the fabrication of the hybrid coupler as described in Section 3.15 was used to fabricate the single-balanced diode mixer. In this case the single-balanced mixer was laid out in three pieces. The pieces were the microstrip RF feedline, the slot layer, and the microstrip circuitry making up the T-branch, low-pass filter, and the remaining circuitry. Once fabricated and assembled, the complete structure was mounted in a test jig and connectorized as shown in the following figures.

Figure 5.6 The single-balanced diode mixer mounted in test jig

Figure 5.7 Top view of single-balanced mixer with inset view of parts

Figure 5.8 Bottom view of mixer highlighting the RF feedline

5.6 Mixer Performance Parameters

Mixer performance can be specified in terms of a variety of parameters. Some of the more common ones are port-to-port isolation, return loss, conversion loss or gain, 1 dB compression point and intermodulation product level.

In mixers, signals at specific frequencies are applied to different ports. Isolation is a measure of how well these signals are separated from one another at each port. Perfectly isolated ports should not allow the signal applied at one port to appear at the other. Unfortunately most mixer ports are not perfectly isolated and therefore a signal applied at one port inevitably emerges to some degree at another port. Isolation of a signal A from port 1 to port 2 is thus defined as

$$Isolation_{1 \rightarrow 2}^A = Power_2^A - Power_1^A \text{ (dB)} \quad (5.18)$$

where $Power_2^A$ is the power of signal A at port 2 and $Power_1^A$ is the power of signal A applied to port 1. The port-to-port isolations of concern in mixers are the LO-IF and LO-RF isolations.

In multiport circuits, a signal incident to one port is either transmitted to other ports or reflected away from the incident port. The summation of the transmitted and reflected signal power must be equal to the total power of the incident signal. Return

loss at a port is defined as the ratio of a signal's reflected power (P_{REF}) to the total power in the incident signal (P_{INC}). In equation form it can be expressed as

$$Return\ Loss = P_{REF} - P_{INC} \text{ (dB)}. \quad (5.19)$$

Both the LO and RF return losses of the single-balanced mixer will be measured.

The goal of a mixer circuit is to take two applied signals, the LO and RF, and generate a new signal at an IF frequency which is a harmonic combination of the other two. The conversion gain or loss is a parameter that is a relative measure of how well the mixer generates an IF signal with reference to the applied RF signal. It is defined as the difference between the IF power (P_{IF}) emerging at the IF port and the applied RF power (P_{RF}) at the input to the mixer according to the following equation.

$$Conversion\ Gain = P_{IF} - P_{RF} \text{ (dB)}. \quad (5.20)$$

To generate the IF signal, an RF signal of very low power is applied to the mixer. As the RF signal is increased the IF signal increases by the same amount such that the conversion loss remains constant. Eventually, increasing the RF power will lead to the nonlinear saturation of the mixer. At this stage, increases in RF power will begin to increase the conversion loss. The RF power level at which the conversion loss increases by 1 dB over its constant level is referred to as the 1 dB compression point. To avoid nonlinear distortion in converting from the RF signal to the IF signal the RF power must be limited below this amount. This defines a dynamic range of operation in terms of allowed RF input power.

Nonlinear devices such as the Schottky-barrier diode generate intermodulation products when two closely spaced RF input tones at frequencies (f_1, f_2) are fed into the mixer. These new mixing products can be expressed through the following relation

$$f_{IM} = \pm qf_1 \pm rf_2 \pm sf_{LO} \quad (5.21)$$

where q , r , and s are positive integers. The most important of these mixing frequencies is the 3rd order intermodulation product given by $(2f_1 - f_2 \pm f_{LO})$ and $(2f_2 - f_1 \pm f_{LO})$ where $(q + r = 3)$. The power level of the intermodulation product of a specific order appearing at the output of the mixer is governed by the following equation

$$P_{IM} = nP_1 - (n - 1)IP_n \text{ (dB)} \quad (5.22)$$

where $n = q + r$ is the order of the IM product, IP_n is the n th-order intercept point, and P_1 is the output power level of the linear response when only one RF signal is applied, $q = 1, r = 0$ or $q = 0, r = 1$. The n th-order intercept point (IP_n) is the extrapolated output power level at which the input-output curves of the linear and IM responses intersect. The n th order IM product varies q dB for every 1 dB change in signal level at frequency f_1 and r dB change for every 1 dB change in signal level at f_2 . Thus given any set of input power levels, the resulting power levels of the intermodulation products can be determined.

5.7 Optimized Simulated Mixer Performance

Once the circuit representation of the various mixer components were obtained (hybrid coupler, IF filter, etc.) either in measured S-parameter or circuit model format, they were all assembled to make up the single-balanced mixer circuit in LibraTM. Using harmonic balance analysis as employed in LibraTM, the performance of the single-balanced mixer was simulated. Using optimization routines in LibraTM, the lengths and widths of the RF matching network, the IF combining circuitry, and the DC-IF return were tuned for optimum mixer performance. The following table summarizes the simulated

performance of the single-balanced diode mixer.

Table 5.3 Simulated Performance For The Single Balanced Mixer

PARAMETER	19.5 GHz (RF)	20.0 GHz (RF)	20.5 GHz (RF)
RF Return Loss	14.98 dB	25.78 dB	13.63 dB
LO Return Loss	5.78 dB	5.78 dB	5.78 dB
RF-IF Isolation	55.72 dB	65.74 dB	65.38 dB
LO-IF Isolation	25 dB	25 dB	25 dB
RF-LO Isolation	28.90 dB	27.62 dB	26.63 dB
LO-RF Isolation	35.54 dB	35.54 dB	35.54 dB
Conversion Loss	7.85 dB	7.89 dB	8.10 dB

5.8 Mixer Measurement Setups

Mixer performance is specified in terms of conversion loss, port isolations, return losses, and several other parameters. Measuring these parameters requires several pieces of equipment arranged in various experimental setups. In each setup, RF and LO sources were required. HP 8350B sweep generators that were able to provide 0 to 14 dBm of signal power were used to generate the RF and LO signals. Attenuators were necessary to reduce the available power down to the required -20 dBm for the RF signal power. The measurement of power levels were conducted using an HP 8569 spectrum analyzer and/or an HP 437B RF power meter. A spectrum analyzer was utilized to display the power levels at each signal frequency at the RF, LO and IF ports of the mixer. The spectrum analyzer provides a relative indication of power levels and not an absolute value. For situations where an absolute measurement of power was required, the power meter was used to provide a reference level for spectrum analyzer readings. In many cases the power meter was not used since it cannot detect signals with very low power levels and since it

measures all power over a wide frequency spectrum it cannot distinguish between RF, LO, and IF frequencies. The following sections illustrate how these instruments and others were used to measure the performance of the single-balanced mixer.

5.8.1 LO-IF and RF-IF Isolation

Figure 5.9 illustrates the instrument setup needed for measuring the LO-IF isolation, RF-IF isolation, and conversion loss performance parameters. The RF source, LO source, and the spectrum analyzer are connected to the mixer's RF, LO, and IF ports respectively via semirigid cables. The RF and LO power entering the mixer (including effects of cables) and that emerging at the IF port after the output cable are measured separately on the spectrum analyzer. The loss across this cable can be taken into account by measuring it with the power meter. The difference between the RF input and output levels will then give us the RF-IF isolation. The same procedure can be used to determine the LO-IF isolation. By measuring the IF signal power and the RF input power into the mixer, the conversion loss can also be determined but an alternate method was used.

Figure 5.9 LO-IF and RF-IF isolation measurement setup

5.8.2 LO-RF Isolation, RF-LO Isolation, And Return Loss

The measurement of return losses and port isolations at the LO and RF ports requires the use of a directional coupler. To measure LO return loss and RF-LO isolation the directional coupler is connected between the LO generator and the LO mixer input port. This device provides one path for the LO signal to travel from the generator to the LO mixer input port with very little attenuation. The directional coupler provides another path for signals to travel from the LO input port to the spectrum analyzer with 10 dB attenuation. LO signals reflected from the LO port and RF signals transmitted from the RF input port to the LO input port can then be directed to the spectrum analyzer where their power levels can be measured. Figure 5.11 illustrates the experimental setup necessary for measuring LO return loss and RF-LO isolation. In a similar manner the RF return loss and LO-RF isolation can be measured and the required setup can be found in Figure 5.11.

Figure 5.10 Experimental setup for measuring RF-LO isolation and LO return loss

Figure 5.11 Measurement setup for LO-RF isolation and RF return loss

5.8.3 Conversion Loss And 1dB Compression Point

The experimental setup required for measuring conversion loss and the 1 dB compression point are displayed in Figure 5.12. The power meter is utilized in this setup since it can provide more accurate readings than the spectrum analyzer. By using a bandpass filter with an operating range from 1.0 GHz to 2.0 GHz, the 1.5 GHz IF signal can be filtered out for measurement with the power meter. The conversion loss can then be determined once the input RF power is measured. The evaluation of the 1 dB compression point is a little more complicated. For this measurement the RF power is increased and the resulting IF power is observed. The RF power at which the IF power drops 1 dB below its extrapolated linear value is the 1dB compression point.

Figure 5.12 Experimental setup for 1dB compression point and conversion loss

5.8.4 Intermodulation Product Measurement

Figure 5.13 shows the experimental setup for measuring 3 dB intermodulation products. In this setup two RF signals 150 MHz apart in frequency are applied to the mixer via a combiner. The input RF power is increased such that the power level of the two RF signals are kept the same. The resulting power levels of the intermodulation products in the IF passband are noted and recorded. The output power level of these intermodulation products and the linear response where only one RF input is applied is plotted versus RF power from which the 3rd order intercept point is extrapolated.

Figure 5.13 Experimental setup for measuring 3rd order intermodulation products

5.9 Measurement Results

As discussed in previous sections various measurements can be made to grade the overall performance of the single-balanced diode mixer circuit. The ones chosen were the conversion loss, LO-IF isolation, LO-RF isolation, RF-IF isolation, LO return loss, RF return loss, 1 dB compression point, IM levels, and 3rd order intercept point. The LO signal power was chosen to be 12 dBm, an adequate amount to turn on the diodes. The RF signal power was chosen to be -15 dBm, the smallest amount possible with the equipment available. The following sections state individual measurement results and observations, detailed discussion only in the following chapter.

5.9.1 Conversion Loss

The conversion loss of the single-balanced mixer circuit was the first parameter to be measured. For this measurement, the experimental setup described in Figure 5.12 was used. An RF signal centred at 20 GHz with a power level of -15.2 dBm and an LO signal fixed at 18.5 GHz with a power level of 12 dBm were applied to the mixer. The resulting IF signal was measured to have a power level of -27.57 dBm resulting in a conversion loss of 12.37 dB. The resulting IF signal centred at 1.5 GHz as viewed on a spectrum analyzer is shown in Figure 5.14.

Figure 5.14 The generated IF signal

The conversion loss of 12.37 dB for an RF and LO power level of -15.2 dBm and 12 dBm respectively was rather high. In light of this the LO power was adjusted to obtain better conversion loss. The LO power was increased from 12 dBm to 14.9 dBm and at the same time the RF power level was increased from -15.2 dBm to -10.2 dBm to verify that the conversion loss remained at a constant level. Figure 5.15 illustrates the conversion loss obtained for the various combinations of RF and LO powers. Examining these results we see that increasing the LO power to 14 dBm improves the conversion loss

to approximately 8.7 dB. Increasing the LO power further to 14.9 dBm leads to a conversion loss of about 8.3 dB. However, increasing the LO power level beyond 14 dBm leads to diminishing returns in the IF power.

Figure 5.15 Measured conversion loss at varying RF and LO power levels

The single-balanced mixer was designed to work with an LO signal fixed at 18.5 GHz and an applied RF signal centred at 20 GHz leading to the generation of an IF signal at approximately 1.5 GHz. It is noteworthy to determine over what RF bandwidth the conversion loss is at its best and remains at a constant level. Figure 5.16 displays the results obtained for conversion loss over the RF bandwidth of 19.4 to 20.5 GHz for an RF and an LO power level of -15.2 dBm and 14.9 dBm respectively. Clearly, the lowest conversion loss is approximately 7.4 dB over the RF frequency range of 19.4 to 19.8 GHz. At an RF frequency of 20 GHz the conversion loss has degraded by about 1 dB.

Figure 5.16 Conversion loss versus RF frequency at an LO level of 14.9 dBm

Simulation showed conversion loss in the range of 7.85 - 8.10 dBm over 19.5 - 20.5 GHz for an LO power of 12 dBm. In comparison, to measure the same level of conversion loss the LO power had to be increased from 12 dBm to 14.9 dBm and even then it was across a much narrower bandwidth, 19.4 - 19.9 GHz.

5.9.2 LO-IF Isolation

The LO-IF isolation was measured as outlined in Figure 5.9. Here an LO and RF signal of 12 dBm at 18.5 GHz and -15.2 dBm at 20 GHz respectively was applied to the mixer. Figure 5.17 illustrates the applied LO signal and the LO signal emerging at the IF port as viewed on the spectrum analyzer. Note that the spectrum analyzer does not measure absolute quantities which is not required for this measurement. Looking at this figure we see that the applied LO power and the LO power at the IF port reads as 6 dB and -34 dB respectively. The resulting isolation is approximately 40 dB.

Figure 5.17 LO-IF isolation measurement

Simulation predicted an LO-IF isolation of only 25 dB and not 40 dB as was measured. This warrants further investigation as measurement is not generally expected to state more favourable results than simulation.

5.9.3 RF-IF Isolation

The experimental setup for this measurement was as shown in Figure 5.9. An LO power level of 12 dBm and an RF power level of -1.80 dBm were used. The RF power was raised from -15.2 dBm to allow the small RF signal at the IF port to be more easily seen. Figure 5.18 illustrates the RF power applied to the mixer and Figure 5.19 shows the RF power emerging at the IF port. Referring to the spectrum analyzer, the measured RF power applied to the mixer was -13.2 dB and that emerging at the IF was -57.2 dB. Measuring the RF power at the IF port required the insertion of an extra piece of semirigid cable with an insertion loss of 2.2 dB. Therefore the resulting RF-IF isolation is 41.8 dB as compared to 65.74 dB under simulation.

Figure 5.18 The applied RF signal

Figure 5.19 RF-IF isolation measurement

5.9.4 RF-LO Isolation

The measurement setup of Figure 5.10 was used to determine the RF-LO isolation. With an applied RF signal of -1.8 dBm and an LO power of 12 dBm the RF power levels at the RF port and LO port as measured on the spectrum analyzer (Figure 5.18 and Figure 5.20) were -13.2 dB and -56.6 dB respectively. The RF power was increased for the same reason as in the previous section. However before the isolation can be evaluated we have to take into account the -12.2 dB of insertion loss presented by the directional coupler and a semirigid cable connection made to the spectrum analyzer. Having done so the RF -LO isolation was evaluated to be 31.2 dB. This is better than simulation which forecasted a value of 27.62 dB.

Figure 5.20 RF-LO isolation measurement

5.9.5 LO-RF Isolation

Figure 5.11 details the experimental setup required for measuring this isolation. Again an RF power level of -1.80 dBm and an LO power level of 12 dBm were used. The figure below displays an LO power of 6 dB and an LO power at the RF port of -34 dB as seen on the spectrum analyzer. As done for the RF-LO measurement, the insertion loss of 12.2 dB due to the directional coupler and cable must be taken into account. As a result the isolation was determined to be 27.8 dB, compared to a simulated value of 35.54 dB.

Figure 5.21 LO-RF isolation

5.9.6 LO Return Loss

For this measurement the setup outlined in Figure 5.10 was implemented. In this case an LO power of 12 dBm and an RF power of -15.2 dBm was used. The input LO power was measured to be 4 dB and the reflected LO power from the LO port was determined to be -12 dB, Figure 5.22. Accounting for the insertion loss of the directional coupler and cable the LO return loss was determined to be 3.8 dB as compared to 5.78 dB for simulation.

Figure 5.22 LO Return Loss

5.9.7 RF Return Loss

To measure the RF return loss the setup of Figure 5.11 was utilized. An RF power of -1.8 dBm and an LO power of 12 dBm was used. Examining the spectrum analyzer trace below, the incident RF power was -24 dB and the reflected RF power was -44 dB. Taking into account the directional coupler and cable the RF return loss was found to be 7.8 dB instead of the simulated value of 25.78 dB.

Figure 5.23 RF Return Loss

5.9.8 1 dB Compression Point

Figure 5.12 was used to determine the 1 dB compression point of the single-balanced mixer circuit. An LO power of 12 dBm was used and the RF power was set at -20 dBm and increased by 1 dB until the IF power no longer increased by a corresponding 1 dB. However, this point was never observed. Increasing the RF power to a maximum of 12 dBm resulted in a similar increase in the IF signal. According to simulation, Appendix G showed that the 1 dB compression point occurs at an RF power level of approximately 6 dBm. This discrepancy will be subject to discussion in the upcoming section.

5.9.9 Measurement OF Intermodulation Products

Figure 5.13 was used to measure intermodulation product levels. For this measurement an LO power of 12 dBm and two RF tones spaced close together (about 100 MHz) at a power of -15.2 dBm were swept in power and the corresponding level of the intermodulation products measured. The following figure shows a trace of the signals at the IF port after injection of the two RF tones. It shows two generated IF signals as well as the presence of 3rd order intermodulation products on either side of these signals. It was expected that a 1dB increase in the input RF powers would result in a corresponding 3dB change in intermod power levels. However a 1 dB increase in RF power only resulted in a corresponding 1 dB increase in intermodulation product power level. Since the same observation was made while measuring the 1 dB compression point, it is not surprising that this would also be the case when measuring intermodulation product levels.

Figure 5.24 Intermodulation Products

CHAPTER 6: CONCLUSION

6.1 Summary

A single-balanced diode mixer which utilized a novel 180-degree hybrid coupler was designed, constructed, and its performance measured. The 180-degree hybrid coupler was realized using a microstrip-based “Magic-T”. It consisted of a two layered substrate with microstrip circuitry on either side aligned back-to-back along a common ground plane. The top and bottom circuitry was used to implement a T-branch and a feed-line terminated in an open circuit respectively. A narrow aperture in the ground plane was used to allow signals in the lower circuitry to be coupled into the upper portion onto which the output ports were attached. At the same time, the aperture restricted signal flow from the top to bottom circuitry. The single-balanced diode mixer used an anti-parallel Schottky-barrier diode pair to perform the mixing function. It also incorporated RF matching circuitry, an IF return, DC - IF blocking circuitry, and a distributed element low-pass filter.

The single-balanced mixer was designed to operate as a downconverter with an RF signal of 20 GHz and an LO signal of 18.5 GHz resulting in an IF signal of 1.5 GHz. The LO signal was to be applied in phase across the diodes while the RF signal was to appear 180 degrees out of phase between diodes. In order to fulfill this role the following steps were followed in the design of the single-balanced diode mixer. First an equivalent model for the Schottky diode that was valid for both large and small signal excitation was chosen. Using measurements, manufacturer’s specifications, and common literature values parameter extraction for the model was carried out by curve fitting simulated to measured results. According to theory, the “Magic-T” could be separated into a T-branch and parallel aperture-coupled lines. A microwave CAD package was used to design the T-branch at the LO frequency and a numerical technique (TLM) was used to design the aperture-coupled lines for optimum signal coupling of the RF signal. The same microwave CAD package was also used to design the distributed element low-pass filter. Once these

three components were designed, the overall circuit was developed with RF matching, DC-IF return, and DC-IF blocking circuitry and optimized for best performance.

The resultant microstrip “Magic-T” design showed very good RF-LO isolation, better than 30 dB. The LO insertion loss was better than 4 dB over a 2 GHz bandwidth centred at 18.5 GHz and the RF insertion loss was around 4 dB centred narrowly over a 1 GHz bandwidth at 20 GHz. These results were obtained after de-embedding the effects of connectors. The in phase and out of phase properties of the “Magic-T” were observed to be present during measurement.

The single-balanced mixer produced an IF signal at 1.5 GHz as intended for an RF input at 20 GHz and an LO signal at 18.5 GHz. However, the measured performance of the mixer fell short of what was designed for. The LO return loss and RF return loss were measured to be 3.8 dB and 7.8 dB instead of the design values of 5.78 dB and 25.78 dB respectively.

As a result of the RF and LO mismatches it is not surprising that the conversion loss was much higher than simulated. To approach the design value of 7.85 - 8.10 dB of conversion loss over 19.5 - 20.5 GHz required raising the LO power from 12 dBm to 14.9 dBm and even then these levels were only obtainable over half the bandwidth, 19.5 - 19.9 GHz.

The LO mismatch explains why the LO-IF isolation was better than simulation, 40 dB compared to 25 dB. The higher the reflected LO power the less there is to reach the IF port. The same reasoning cannot explain why the RF-IF isolation degraded from a design value of 65.7 dB to 41.8 dB. However, since these values represent very low power levels to begin with, this discrepancy is not very significant.

Looking at the RF-LO isolation we see that it improved over its design value of 27.62 dB to 31.2 dB. Again the RF mismatch can explain this observation. The same explanation cannot be used to account for the degradation in LO-RF isolation from 35.5 dB to 27.8 dB. The prototype “Magic-T” was used in the design of the mixer circuit,

but since a “Magic-T” had to be integrated with the mixer layout, it was not used in the final mixer. Discrepancies in processing and assembly could construct a “Magic-T” whose properties are different from the prototype such that they adversely affect performance.

With regards to the absence of a 1 dB compression point, the significant RF mismatch is such that it limits the RF power reaching the diode. Therefore with the equipment available, the RF power could not be raised high enough to saturate the diode. Likewise, although the intermodulation products were observed they did not change 3 dB for every 1 dB change in RF power for the same reason.

There are several possible explanations as to why the observed LO and RF mismatches are present. One unlikely reason might be the discrepancy between components such as the “Magic-T” and the diodes used in the design and those used in the actual circuit. A more plausible explanation is that the DC-IF block, an ideal capacitor, used in the design did not model the chip capacitor accurately. Also a simplified model utilizing a series inductance and resistance combination was used to represent the vias in the mixer circuit since the via models as provided by LibraTM failed to converge during simulation. Finally, many of the elements used in the simulation and design of the mixer did not take into account the electromagnetic effects present at the high frequencies that were operated at.

In conclusion a working single-balanced mixer which incorporated a novel 180-degree hybrid coupler in the form of a microstrip based “Magic-T” was illustrated. To this researcher’s knowledge this is the first realization of this particular form of mixer circuit. The microstrip “Magic-T” showed satisfactory performance in its function as a 180-degree hybrid coupler. It presents itself as an alternative 180-degree hybrid whose structure presents distinct advantages in isolation between RF and LO ports. The single-balanced mixer despite not meeting performance expectations due to concerns already stated nevertheless was valuable in highlighting the use and integration of the microstrip

“Magic-T” in a working circuit.

6.2 Future Work

Clearly the discrepancy between simulation and measurement should be addressed and prompt further investigation. The apparent RF and LO mismatch is considered to be a major factor in this discrepancy. In hindsight, capacitor manufacturer's provide more accurate models that take into account series resistances, and stray inductances and these should have been used to represent the DC-IF block. Also a better DC-IF block could have been realized using distributed elements. The provided via models presented convergence problems in simulation and further investigation revealed that the diameter of the surface metal had no bearing in RF performance. Therefore, since the main components of this model are its series inductance and resistance, this provided some justification for using a lumped element representation. Finally, the frequency of operation is high enough to warrant the use of electromagnetic analysis in the mixer design which in this case was used sporadically. Any further work should focus on addressing these problems as well as investigating the possibility of using IF matching circuitry to improve conversion loss.

Further investigation into improving the microstrip “Magic-T” by using different upper and lower substrate materials, different aperture configurations, and T-branch improvements such as the use of cutouts in the metallization at the T-branch junction should also be conducted.

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APPENDIX A
DMK279 SCHOTTKY DIODE SPECIFICATIONS

Chip and Beam Lead Mixer Diodes

 $(T_A = 25^\circ\text{C})$

			Electrical Characteristics by Frequency Band				
Test	Test Conditions	Units		X	Ku	Ka	mm
V_f	1 mA	mV	Min	500	600	600	500
			Max	500	300	300	300
C_j	0V 1MHz	pF	Min	.03	.03	.07	.025
			Max	.12	.13		
R_s	10 mA	Ω	Max	4.0	6.0	6.0	8.0
V_s	10uA	V	Min	3.0	3.0	3.0	3.0
NF(ssb)	LO = 7 mW $N_f = 1.0\text{dB}$	dB	Max	5.0	6.3	7.0	
Style	Outline Number	Part Number	Part Number	Part Number	Part Number	Part Number	
Chip:	270-807	CMK7703-000	CMK7704-000	CMK7705-000			
Beam-Lead: Singles	481-011	CMK2504-000	CMK2505-000	CMK2506-000	DMK2791-000		
	468-011				CMK2784-000		
Pairs	378-016			CMK6591-000			
	490-025			CMK2307-000			
Quad Rings	294-002			DMK6592-000			

Notes:

Matching Criteria:

Pairs: ΔV_f @ 1mA = 15 mV max.Quads: ΔV_f @ 1mA = 20 mV max.

Junction Capacitance is determined by measuring the total capacitance and then subtracting the overlay (parasitic) capacitance which is calculated to be 0.02 pF (0.015 pF for DMK2784-000).

Cutoff Frequency (F_{co}) may be calculated as follows: $F_{co} = \frac{1}{2\pi C_{jo} R_s}$

LO = 5mW

 $N_f = 1.0\text{dB}$

Test Conditions
 $R_s = 10 \text{ ohms}$

Typical DMK6583
GaAs X-Band Mixer Diode
Admittance Characteristics

Test Conditions
 $R_s = 10 \text{ ohms}$

Typical DMK3308
GaAs X-Band Mixer Diode
Admittance Characteristics

APPENDIX B

DIODE CIRCUIT FILES AND SIMULATIONS

Diode Circuit And Simulation File (HARPE™)

```

: Small signal model for varactor
: File = ssmodel.ckt

EXPRESSION
  Isat:71.07e-14A 1.07e-14A 1.28e-14A?;
  Nideal:71.00980 1.11075 1.23420?;
  Rseries:73.8oh 4.04119oh 4.25oh?;
  Lseries:0.25nH;
  Cpar:0.02pf;
  Rqground:2600000oh;
  Cjun: 0.025pf;
  Vjun:0.764181V;
  Cshunt:0.021307pf;
  Fcoeff:0.911866;
  EXGcap:0.447466;

END

MODEL
  RES      1      4      R=3.5oh;
  RES      4      0      R=Rqground;
  CAP      4      0      C=Cshunt;
  RES      5      6      R=Rseries;
  IND      4      5      L=Lseries/2;
  DIODE    6      7      IS=Isat N=Nideal CJO=Cjun VJ=Vjun FC=Fcoeff
  CAP      5      7      C=Cpar;
  CAP      8      0      C=Cshunt;
  IND      7      8      L=Lseries/2;
  RES      8      9      R=1.7oh;
  RES      8      0      R=Rqground;
  ZPCR     1      9;

END

DATA
!S-Parameter measurement data
  #INCLUDE "d_dls.txt";
  #INCLUDE "dm3s9.dat";
  #INCLUDE "dm3s1.dat";
  #INCLUDE "dm3s2.dat";
  #INCLUDE "dm3s4.dat";
  #INCLUDE "dm3s5.dat";
  #INCLUDE "dm3s6.dat";
  #INCLUDE "dm3s7.dat";
  #INCLUDE "dm3s8.dat";
  #INCLUDE "dm3s9.dat";
  #INCLUDE "dm3s10.dat";
  #INCLUDE "dm3s11.dat";
  #INCLUDE "dm3s12.dat";
  #INCLUDE "dm3s13.dat";
  #INCLUDE "dm3s14.dat";
  #INCLUDE "dm3s15.dat";
  #INCLUDE "dm3s16.dat";
  #INCLUDE "dm3s17.dat";
  #INCLUDE "dm3s18.dat";
  #INCLUDE "dm3s19.dat";
  #INCLUDE "/home/users/almeida/thesis/Diode1_Series_DNA/dm3s01.dat";
  #INCLUDE "dm3s02.dat";
  #INCLUDE "dm3s03.dat";
  #INCLUDE "dm3s04.dat";
  #INCLUDE "dm3s05.dat";
  #INCLUDE "dm3s06.dat";
  #INCLUDE "dm3s07.dat";
  #INCLUDE "dm3s08.dat";
  #INCLUDE "dm3s09.dat";
  #INCLUDE "dm3s10.dat";
  #INCLUDE "dm3s11.dat";
  #INCLUDE "dm3s12.dat";
  #INCLUDE "dm3s13.dat";
  #INCLUDE "dm3s14.dat";
  #INCLUDE "dm3s15.dat";
  #INCLUDE "dm3s16.dat";
  #INCLUDE "dm3s17.dat";
  #INCLUDE "dm3s18.dat";

END

SWEEP
  FREQ: FROM 15GHZ TO 25GHZ STEP=0.25GHZ VG=-0.597V 0.111V
        0.492V 0.571V 0.632V 0.702V 0.764V 0.805V 0.888V 1.104V
        VD=0V MS11 PS11 MS21 PS21 MS12 PS12 MS22 PS22;

END

SPECIFICATION
  FREQ: FROM 15GHZ TO 25GHZ STEP=1GHZ VG=-0.597V 0.111V 0.492V
        0.571V 0.632V 0.702V 0.764V 0.805V 0.888V 1.104V
        VD=0V MS11=MS11 PS11=PS11 MS21=MS21 PS21=PS21;

END

```

Model And Measured S_{11} Response For $V=0.0114$ V

Model And Measured S_{21} Response For $V=0.0114$ V

Model And Measured S_{11} Response For $V=0.632$ V

Model And Measured S_{21} Response For $V=0.632$ V

Model And Measured S_{11} Response For $V=0.702$ V

Model And Measured S_{21} Response For $V=0.702$ V

Model And Measured S_{11} Response For $V=1.104$ V

Model And Measured S_{21} Response For $V=1.104$ V

APPENDIX C
TLM ANALYSIS OF APERTURE COUPLING

Figure 1.1 S_{11} For 0.2λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.2 S_{21} For 0.2λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.3 S11 For 0.3λ Aperture At Various Stub Lengths (Simulation)

Figure 1.4 S11 For 0.3λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.5 S_{21} For 0.3λ Aperture Length At Various Stub Lengths

Figure 1.6 S_{21} For 0.3λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.7 S_{11} For 0.4λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.8 S_{11} For 0.4λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.9 S_{11} For 0.4λ Aperture Length At Various StubLengths (Simulation)

Figure 1.10 S_{21} For 0.4λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.11 S_{21} For 0.4λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.12 S_{21} For 0.4λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.13 S11 For 0.45λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.14 S21 For 0.45λ Aperture Length At Various Stub Lengths (Simulation)

Figure 1.15 S_{11} For 0.1λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.16 S_{11} For 0.2λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.17 S_{11} For 0.3λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.18 S_{11} For 0.35λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.19 S_{11} For 0.4λ Stub length At Various Aperture Lengths (Simulation)

Figure 1.20 S_{21} For 0.1λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.21 S_{21} For 0.25λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.22 S_{21} For 0.3λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.23 S_{21} For 0.35λ Stub Length At Various Aperture Lengths (Simulation)

Figure 1.24 S_{21} For 0.4λ Stub Length At Various Aperture Lengths (Simulation)

APPENDIX D
T-JUNCTION CIRCUIT FILE AND SIMULATION

T-Junction Circuit File

Libra (TM) Ver. 3.500.103.3 Cfg. (800 21525 5 5540D847 9276 0 0 11FC31DE)
 tline_step.ckt Sat Apr 16 17:14:44 1994

L2=1778.5
 L3=1739.5
 WCPT#10 487.7864 15000

CKT

MSUB ER=9.96 H=254 T=1 RHO=1 RGH=0
 TAND TANC=0.001

MTEE 1 2 3 W1=256 W2=256 W3=256

MLIN 1 4 W=256 L=7400.0

MBEND3 4 5 W=256

MLIN 5 6 W=256 L=7500

MLIN 2 7 W=256 L=7400

MBEND3 7 8 W=256

MLIN 8 9 W=256 L=7500

MLIN 3 10 W^WCPT L=7500

DEF3P 6 9 10 TEE

OUT

TEE db[S13] gr1

TEE db[S23] gr1

TEE db[S33] gr2

FREQ

SWEEP 17.5 19.7 0.1

T-Junction Simulated Response

FREQ-GHZ	DB[S13] TEE	DB[S23] TEE	DB[S33] TEE
17.5000	-3.560	-3.560	-14.936
17.6000	-3.537	-3.537	-15.751
17.7000	-3.516	-3.516	-16.651
17.8000	-3.497	-3.497	-17.648
17.9000	-3.481	-3.481	-18.755
18.0000	-3.467	-3.467	-19.995
18.1000	-3.456	-3.456	-21.396
18.2000	-3.447	-3.447	-23.002
18.3000	-3.441	-3.441	-24.878
18.4000	-3.436	-3.436	-27.115
18.5000	-3.434	-3.434	-29.787
18.6000	-3.432	-3.432	-32.614
18.7000	-3.433	-3.433	-33.767
18.8000	-3.435	-3.435	-31.785
18.9000	-3.439	-3.439	-28.915
19.0000	-3.445	-3.445	-26.382
19.1000	-3.453	-3.453	-24.274
19.2000	-3.462	-3.462	-22.497
19.3000	-3.474	-3.474	-20.970
19.4000	-3.488	-3.488	-19.635
19.5000	-3.504	-3.504	-18.453
19.6000	-3.522	-3.522	-17.327
19.7000	-3.544	-3.544	-16.448

APPENDIX E

CONNECTOR MODEL AND FIT

Circuit Model Of Connectorized Transmission Line

DATA
MSUB
MSUB1
ER=9.98
H=251
T=1
RHO=1.80
RGH=0

DATA
MSUB
MSUB2
ER=1
H=251
T=1
RHO=1
RGH=0

DATA
TAND
TAND1
TAND=1.00e-03

var
Len
VAR
-VAR
Z=52.55
C=6.27e-03 unconst
L=1.3000
A=4.43e-05
F=20

Connector Model

2-Terminal Physical Transmission Line

TLINP

Symbol:

Illustration:

Parameters:

- Z = characteristic impedance, in resistance units
- L = physical length, in length units
- K = effective dielectric constant
- A = attenuation, in dB per unit length
- F = frequency for scaling attenuation, in frequency units

- TAND = ID of dielectric loss tangent (TAND) Data Item
- PERM = ID of permeability (PERM) Data Item
- SIGMA = ID of dielectric conductivity (SIGMA) Data Item
- TEMP = ID of physical temperature (TEMP) Data Item

Range of Usage:

$$Z > 0 \quad K \geq 0 \quad A \geq 0 \quad F \geq 0$$

Notes/Equations/References:

1. The A parameter specifies conductor loss only. To specify dielectric loss, specify non-zero value for TAND (to specify a frequency-dependent dielectric loss) or SIGMA (to specify a constant dielectric loss).
2. Since conductor and dielectric losses can be specified separately, the element is not assumed to be distortionless. Therefore, the actual characteristic impedance of the line may be complex and frequency-dependent. This may cause reflections in your circuit that would not occur if a distortionless approximation were made.
3. $A(f) = A$ (for $F = 0$)

$$A(f) = A(F) \cdot \sqrt{\frac{f}{F}} \quad (\text{for } F \neq 0)$$

where

f = simulation frequency

F = reference frequency

Model And Measured S_{11} (Magnitude)

Model And Measured Phase Of S_{11} (Phase)

Model And Measured Magnitude Of S_{21} (Magnitude)

Model And Measured Phase Of S_{21} (Phase)

APPENDIX F

LOW-PASS FILTER SIMULATION AND ANALYSIS

Libra And Electromagnetic Analysis Of Filter

	B1 SIJ1 mar23	B1 SIJ1 mar23	B2 SIJ1 mar23b	B2 SIJ1 mar23b	B1 SIJ2 mar23	B1 SIJ2 mar23	B2 SIJ2 mar23b	B2 SIJ2 mar23b
Frequency GHz	S[1,1] dB	S[1,1] Ang deg	S[1,1] dB	S[1,1] Ang deg	S[2,1] dB	S[2,1] Ang deg	S[2,1] dB	S[2,1] Ang deg
0.100000000	-18.6805	-3.9453	-18.9669	-7.6575	-1.0761	-8.7092	-1.0427	-8.7061
0.600000000	-18.3791	-22.1275	-19.4091	-40.4460	-1.1019	-52.2781	-1.1006	-52.3094
1.100000000	-17.3830	-39.2025	-19.8391	-56.4085	-1.1495	-96.3831	-1.1719	-96.3786
1.600000000	-18.3363	-61.6830	-20.5144	-68.9504	-1.2078	-141.8766	-1.2635	-141.7610
2.100000000	-24.4978	16.7316	-26.0258	7.5833	-1.3967	168.8585	-1.4748	169.5200
2.600000000	-7.7477	6.7574	-8.6804	5.0968	-2.7018	113.8844	-2.6349	115.0680
3.100000000	-2.5979	-36.5416	-2.9181	-38.1994	-6.1386	63.6412	-5.9428	63.6400
3.600000000	-1.0440	-68.9283	-1.1450	-71.3150	-10.2155	27.8541	-10.1244	26.6019
4.100000000	-0.5421	-91.7917	-0.5746	-94.3384	-13.5882	2.8327	-13.6642	1.0963
4.600000000	-0.3523	-109.1715	-0.3595	-111.5080	-15.9326	-16.1670	-16.1738	-17.9078
5.000000000	-0.2882	-120.7873	-0.2837	-122.8160	-17.0491	-28.9013	-17.4152	-30.4602
18.000000000	-2.1973	172.9195	-4.6902	-172.7240	-6.9194	91.7709	-4.3322	110.2050
18.500000000	-1.0631	147.4169	-2.0565	154.2300	-9.9689	62.4344	-7.3882	71.3753
19.000000000	-0.7500	127.0595	-1.3544	128.9150	-11.3516	39.7773	-9.2799	43.1657
19.500000000	-0.7154	106.9368	-1.1874	105.6850	-11.3090	17.8487	-9.8485	17.9997
20.000000000	-0.9063	84.1036	-1.2805	81.0502	-10.1134	-6.7135	-9.3050	-8.1448
20.500000000	-1.4454	56.0140	-1.6354	52.6614	-8.0814	-36.6107	-7.8995	-37.8502
21.000000000	-2.6445	20.4758	-2.4418	18.7856	-5.6886	-73.8478	-5.9275	-73.1050
21.500000000	-5.0343	-23.1574	-4.3194	-21.4126	-3.5293	-118.4680	-3.7544	-115.5300
22.000000000	-9.9367	-73.3059	-9.9517	-64.5897	-1.9475	-168.5128	-1.9300	-166.8620

B2 - Libra Simulation

B1 - Electromagnetic Analysis

APPENDIX G
MIXER CIRCUIT FILE AND SIMULATION

Single-Balanced Diode Mixer Circuit File

Test Bench For Simulating Mixer Circuit

Simulated Performance Of Single-Balanced Diode Mixer

	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb
	S11_RF	S11_LO	PF_LO_IN_RF	PF_RF_IN_LO	PF_RF_IN_IF	PF_LO_IN_IF
	May20_Mix	May20_Mix	May20_Mix	May20_Mix	May20_Mix	May20_Mix
Frequency	CUT_EQN	CUT_EQN	PF	PF	PF	PF
GHz	dB	dB	dBm	dBm	dBm	dBm
19.500000000	-5.4554	-3.1619	-25.3094	-49.3946	-30.6224	-14.5377
20.000000000	-11.9560	-3.1314	-25.1397	-49.3611	-31.5229	-14.4589
20.500000000	-5.5324	-3.1358	-25.3946	-46.6896	-105.0330	-14.4684

	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb	May20_Mix_tb
	PF_RF	PF_LO	PF_IF	Conversion_Loss	Conversion_Loss
	May20_Mix	May20_Mix	May20_Mix	May20_Mix	May20_Mix
Frequency	PF	PF	PF	CUT_EQN	CUT_EQN
GHz	dBm	dBm	dBm	Mag	Re
19.500000000	-20.2231	10.8472	-27.9569	7.7289	-7.7289
20.000000000	-20.3177	10.8294	-27.5600	7.5423	-7.5423
20.500000000	-20.3539	10.8311	-29.3340	9.4801	-9.4801

Conversion Loss Versus RF Power (1 dB Compression Point)

IMAGE EVALUATION TEST TARGET (QA-3)

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