Ku/Ka band mixers development for satellite communication systems

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Ku/Ka Band Mixers Development for Satellite Communication Systems

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in fulfilment of the requirement for the degree of
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Summary

This thesis focuses on Ku and Ka band mixer design, fabrication, assembly, and measurement. In order to achieve the required performance, different transmission lines or combinations of them are used for various band mixer designs.

For the Ku band mixer, microstrip is used. A Lange coupler is used for RF and LO input ports because of its capability of wideband operation and good port-to-port isolation. Beam-lead Schottky barrier diodes are used to realize the mixing operation after the Lange coupler. In general, a low pass filter (LPF) follows to take out the desired IF components while suppressing the harmonics. For this mixer, however, an additional RF short is used before the LPF for further RF and LO leakage rejection. Normally, the RF leakage rejection of a LPF is adequate for a single balanced mixer. However, for a satellite ground station, a transceiver set is designed so that antenna leakage is apparently large and LO power leakage is also significant due to the same reference crystal used. RF leakage may be originated from antenna coupling and common shared synthesizer crystal, therefore extra RF short is needed for further RF rejection, especially for wideband operation. Method of moment (MoM) based software is used for the structure analysis.

For the Ka band mixer, finline-microstrip combination structure is used. RF is fed into the mixer through a unilateral finline taper and LO is fed into the mixer through an antipodal finline taper. The two Schottky diodes are connected in series across the broadwall of the waveguide as seen from the finline RF input port and they are in parallel to the LO signal. The RF input unilateral finline taper is designed to a suitable slot width, with its characteristic impedance matched to the serial impedance of two

diodes. The LO input antipodal finline taper is also designed to a slot width with its characteristic impedance matched to the parallel impedance of two diodes. Therefore, the impedance level of the unilateral finline taper, the diode and the antipodal finline taper must be in ratio 4:2:1. A LPF is used to take out the IF components. And a special structure, called “swallow tail”, is used to adjust flatness of RF/LO isolation over the whole Ka band. Finite element method based software is used for the structure analysis.

The simulation result shows the high RF/LO isolation for the Ku band mixer due to the wonderful port-to-port isolation of the designed Lange coupler. The RF and LO leakage rejections are both shown to be good for the Ku band mixer since the additional RF short structure is used. Meanwhile, the flatness adjustment ability of “swallow tail” is proved for the Ka band mixer. A scaled S band mixer is designed and measured. The results show that the designs of the Ku and Ka band mixers are feasible.

Chapter 1 Introduction

1.1 Introduction

In 1959, J. R. Pierce and R. Kompfner described the transoceanic communication by satellites. Today, there are many communication satellites in far geosynchronous orbit (GEO) (from 35.788 to 41.679 km), near low earth orbit (LEO) (from 500 to 2000 km), and at the medium-altitude orbit (MEO), which is between the GEO and LEO orbits. Technological advances have resulted in more alternatives in satellite orbits, more output power in transmitters, lower noise in receivers, higher speed in modulation and digital circuits, and more efficient solar cells. Satellite communications have provided reliable, instant, and cost-effective communications on regional, domestic, national, or global levels.

Figure 1.1 Satellite communication link

A simple satellite communication link is shown in Figure 1.1. The earth station A transmits an uplink signal to the satellite at frequency f_U . The satellite receives,

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amplifies, and converts this signal to a frequency f_D . The signal at f_D is then transmitted to earth station B. The system on the satellite that provides signal receiving, amplification, frequency conversion, and transmitting is called a repeater or transponder. Normally, the uplink is operating at higher frequencies because higher frequency corresponds to lower power amplifier efficiency. The efficiency is less important on the ground than on the satellite. The reason for using two different uplink and downlink frequencies is to avoid the interference, and it allows simultaneous reception and transmission by the satellite repeaters. Some commonly used uplink and downlink frequencies are listed in Table 1.1. For example, at the C-band, the 4-GHz (3.7-4.2 GHz) is used for downlink and the 6-GHz band (5.925-6.425 GHz) for uplink.

Table 1.1 Commercial satellite communication frequencies

Band	Uplink Frequency (GHz)	Downlink Frequency (GHz)
L	1.5	1.6
C	6	4
X	8.2	7.5
Ku	14	12
Ka	30	20
Q	44	21

The earth stations and satellite transponders consist of many RF and microwave components. As an example, Figure 1.2 shows a simplified block diagram operating at the Ku band with the uplink at 14-14.5 GHz and downlink at 11.7-12.2 GHz. The earth terminal block consists of two upconverters converting the baseband frequency of 70 MHz to the uplink frequency. A power amplifier (PA) is used to boost the output power before transmitting. The received signal is amplified by a low-noise RF amplifier (LNA) before it is downconverted to the baseband signal. The transponder

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block on the satellite receives the uplink signal (14-14.5 GHz). It amplifies the signal and converts the amplified signal to the downlink frequencies (11.7-12.2 GHz). The downlink signal is amplified by a power amplifier before transmitting.

The purpose of this project is to develop broadband mixers for Ku band downconversion and Ka band upconversion. Ka band design will use similar balanced structure for downconversion and possibly, to inject the power reversely to achieve upconversion capability. For some transceiver sets using in satellite communication system, the same reference crystal is used for LO power as shown in

Figure 1.2 A simple block diagram operating at Ku band

Figure 1.3 to save cost. However, the RF power of the upconverter may leak to the downconverter due to antenna coupling and therefore interfere the weak received IF signal. Also large LO power may leak to antennas and then be transmitted out to interfere other systems. In this project, an additional RF short is used for the Ku band mixer to reject the RF and LO leakage and Lange coupler is used for RF and LO input ports due to its high port-to-port isolation and wideband operation property. For the Ka band mixer, RF/LO isolation is due to the field orthogonality of the RF and LO signals propagated in finline and microstrip. Gap transition between finline

and microstrip can further improve it. In addition, a special structure, “swallow tail”, is used to adjust the flatness ability of RF/LO isolation.

Figure 1.3 A transceiver set with the same crystal

1.2 A brief introduction of mixer

Either a downconverter or an upconverter is basically a mixer. Before starting the design, basic knowledge of mixers is necessary, such as their operation principle and categories. Microwave mixers make use of nonlinear semiconductor devices, usually Schottky barrier diodes due to their low impedance at high frequency and also low transit time for mixing operation [1].

1.2.1 Basic principle of mixer action

Figure 1.4 The basic mixer operation

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Consider a single-ended mixer as shown in Figure 1.4 with RF and LO signals as

$$V_{RF}(t) = V_{RF} \sin \omega_{RF} t$$

(1.1)

$$V_{LO}(t) = V_{LO} \sin \omega_{LO} t$$

(1.2)

are fed into a diode with V_{RF} and V_{LO} the amplitudes, ω_{RF} and ω_{LO} the frequencies of the RF and LO signals respectively. In this case, the output voltage can be expressed as

$$V = V_{RF} \sin \omega_{RF} t + V_{LO} \sin \omega_{LO} t$$

(1.3)

Since a diode has a nonlinear IV characteristic with output current generally written as

$$I = a_1 V + a_2 V^2 + a_3 V^3 + \dots$$

(1.4)

Then

$$I = a_1(V_{RF} \sin \omega_{RF} t + V_{LO} \sin \omega_{LO} t) + a_2(V_{RF} \sin \omega_{RF} t + V_{LO} \sin \omega_{LO} t)^2 + \dots \quad (1.5)$$

$$= a_1(V_{RF} \sin \omega_{RF} t + V_{LO} \sin \omega_{LO} t) + a_2\{0.5V_{RF}^2(1 - \cos 2\omega_{RF} t) + V_{RF}V_{LO}[\cos(\omega_{RF} - \omega_{LO})t - \cos(\omega_{RF} + \omega_{LO})t] + 0.5V_{LO}^2(1 - \cos 2\omega_{LO} t)\} + \dots$$

(1.6)

The frequency spectrum of this output is illustrated in Figure 1.5. The difference or intermediate frequency is

$$\omega_{IF} = (\omega_{RF} - \omega_{LO}) \text{ or } (\omega_{RF} + \omega_{LO})$$

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Since ω_{RF} is usually near ω_{LO} , the different frequency ($\omega_{RF}-\omega_{LO}$) can be extracted from the mixer by using a low pass filter if the mixer is used as a downconverter. However, if using as an upconverter, the IF component, ($\omega_{RF}+\omega_{LO}$), can be selected by a band pass filter. The spectrum of output signals is shown in Figure 1.5.

Figure 1.5 Frequency spectrum of a mixer

1.2.2 Mixer category

Practical mixer configurations can be broadly divided into three categories. There are, single ended mixers, single balanced mixers, and double balanced mixers. Single ended mixer, shown in Figure 1.4, is the simplest and least efficient mixer. It is also rather easy to construct. However, this design is difficult to provide LO energy while maintaining the separation of LO, RF, and IF signals for broadband application.

The single balanced dual-diode or dual-transistor mixer in conjunction with a hybrid coupler offers the ability to conduct such broadband operations. Moreover, it provides further advantages related to noise suppression and spurious mode rejection. Spurious arises in oscillators and amplifiers due to parasitic resonance and nonlinearity and they are only partially suppressed by the front end. Thermal noise

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can critically raise the noise floor in the receiver. Figure 1.6 shows the basic single balance mixer design featuring a quadrature coupler and a dual-diode followed by a low pass filter (LPF). Besides an excellent voltage standing wave ratio (VSWR), this design is capable of suppressing a considerable amount of noise because the opposite diode arrangement in conjunction with the 90° phase shift provides a good degree of noise cancellation.

Figure 1.6 A single balanced mixer

The double balanced mixer can be constructed by using four diodes arranged in a rectifier configuration. The additional diodes provide better isolation and an improved suppression of spurious modes. Unlike the single balanced approach, the double balanced design eliminates all even harmonics of both the LO and RF signals. However, the disadvantages are considerably high LO driving power and increased conversion loss [2].

Single ended design is difficult to provide LO power while maintaining the separation of LO, RF, and IF signals for broadband application. Double balanced design requires higher LO driving power and increases conversion loss, which results in high cost for both high LO power and amplifiers at high frequency (Amplifier is used to compensate the high conversion loss). Therefore, a single balanced mixer structure is selected for both the Ku and Ka band mixer design. In order to achieve the specified

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performance requirements, a mixer circuit may consist of additional RF short, coupler and other structures. The details will be discussed in the rest of this thesis.

1.3 Motivation

Microwaves are desirable for communications and radar applications because of their high frequency and short wavelength. The high frequency provides wide bandwidth capability. Because of the short wavelength of microwaves, high-gain antennas with narrow beamwidths, used in radar applications, can be constructed. At the very high frequency of microwaves, conventional transistors, ICs, and wirings will not work well due to the lead reactance and the transit time so that special microwave devices are required. This project is aimed to develop broadband mixers at Ku and Ka band according to the performance requirements of satellite communication systems.

For the Ku band mixer, some structures consisting of two quarter-wavelength open circuited lines, which are connected in shunt behind the mixer diodes and providing the diodes with low impedance RF ground, had already been proposed [1,3]. But most of them are only applicable to narrow band. Although some designs used broadband RF short for the designed mixer, two open quarter wavelength radial stub designs, as a band stop filter to reject the LO signal [4], the branch line coupler limited the broadband operation of the mixer. The mixer with a Lange coupler and two open quarter wavelength radial stubs was reported in [5]. The radial stub IF filter was able to achieve RF rejection more than 30dB at Ku band. But the structure was relatively complex to fabricate. In order to achieve high RF rejection at Ku band, a Lange coupler, two open radial stubs followed by a LPF structure has been used in

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this design. It shows good RF leakage rejection and harmonics suppression over the whole Ku band with the relatively simple structure.

For the Ka band mixer to achieve wide bandwidth, high RF/LO isolation and low conversion loss, most of mixers use combination of different transmission lines, i.e. finline-microstrip structures. Finline structure mixers for Ka band and above were reported to have good performance [6-9]. Among which, finline balanced mixer in [7] operated over a 32GHz IF instantaneous bandwidth with a conversion loss of 8 to 12 dB, but the LO was restricted at 74GHz as the RF was swept from 76 to 108GHz. The other one in [9] had high Q injected LO signal, also restricted in LO frequencies. Moreover, both band pass filters limited for narrow band LO frequency and the desired frequency would be shifted much if the band pass filter were not manufactured exactly. On the other hand, [6] and [8] both showed good broadband RF and LO frequency coverage. For these two structures [8], however, had additional bulky tuning short that was sizable in dimension. The structure suggested in [6] used only microstrip coupler to prevent IF short and achieve wideband LO frequency. It sacrificed the pass loss of LO power about 2 to 3 dB but had benefit in size reduction. Also, the structure provided novel “swallow tail” adjusting element for frequency flatness tuning, which was beneficial to broadband high RF/LO design. Previous experiment result for RF/LO isolation in [6] was as high as 30dB over the frequency range 27-40GHz. But it lacked of structure analysis. A broadband and good RF/LO isolation mixer is to design at Ka band. Its structure and performance analysis will be given in this thesis.

1.4 Objectives

The objectives of this project are

- To understand the principle operation of a single balanced mixer
- Using the method of moment (MoM) based software to evaluate the following:
 - The properties of Lang coupler, including 3 dB bandwidth and phase shift over wide frequency range
 - The design of low pass filters using microstrip with 0.1dB attenuation at the pass band and more than 20dB attenuation at the stop band
- Using the finite element method (FEM) based software to evaluate the following
 - The unilateral finline taper design to achieve an impedance transformation from waveguide impedance of 480Ω to beam lead diode impedance of 180Ω .
 - The antipodal finline taper design to achieve an impedance transformation from waveguide impedance of 480Ω to 50Ω of microstrip line.
 - The translation from finline to microstrip with less insertion loss at the required frequency range and also be able to provide good port-to-port isolation

1.5 Major contribution of the thesis

- Realize the broadband mixer with Lange coupler and finline tapers

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- Achieve high LO/RF isolation using high port-to-port isolation property of Lange coupler for the Ku band mixer and using signals' orthogonality and gap transition for the Ka band
- Perform high and wideband RF and LO leakage rejection using combination of RF short and LPF for the Ku band mixer
- Study the finline to microstrip line transition using FEM method

1.6 Organization

The thesis is arranged in the following way. The basic microstrip components of the mixers are introduced and designed in Chapter 2. Chapter 3 gives a detail introduction of finline components and their design for the Ku and Ka band mixers. The structure and performance analysis of these two mixers are presented in Chapter 4. Chapter 5 provides the scaled design and measurement of a S band mixer for evaluation purpose. Finally, the thesis ends with conclusion and future work in Chapter 6.

Chapter 2 Microstrip Components of the Mixers

2.1 Introduction

To achieve the required performance, different transmission lines are used for various band mixer designs [10-12]. Normally, microstrip transmission lines are used in order to simplify the problem of making connections in a complex microwave circuit. However, as the frequency becomes higher, the MIC equations are no longer applicable due to dispersion effect. In this design, microstrip is used for the Ku band mixer, while finline-microstrip combination is used for the Ka band mixer due to the frequency requirement shown in Figure 2.1.

Figure 2.1 Frequency specifications of the mixers

As discussed in Chapter 1, the single balanced mixer structure is selected for the two mixers since it is able to provide good broadband RF/LO isolation and noise suppression and spurious mode rejection comparing to single-ended structure, while require considerably lower LO driving power and has less conversion loss comparing to double-balanced structure.

Chapter 2 Microstrip Components of the Mixers

Figure 2.2 shows the design pattern of the Ku band mixer. RF and LO are both fed to the mixers through two ports of a Lange coupler, which has broadband and good port-to-port isolation features. Then two Schottky diodes, in opposite polarities, follow to make mixing operation. A microstrip radial stub or RF short, and a LPF are connected to get the desired IF components while suppress some byproducts, for example, harmonics. Therefore, these components form a single balanced mixer.

Figure 2.2 The circuit pattern of the Ku band mixer

The finline-microstrip combination structure for the Ka band mixer is given in Figure 2.3. The RF is fed to two diodes through a unilateral finline taper. The LO pumped power is fed to Schottky diodes through an antipodal finline taper and a coupled line DC-block (it also serves as IF-block). A microstrip low pass filter is used to take out the IF output. On the backside of the unilateral finline taper, there are two triangular metal pieces, which are called “swallow tail” as indicated in Figure 2.3. Its flatness adjustment ability on RF/LO isolation will be discussed in Chapter 4.

In the two circuit patterns shown above, the Ku band mixer makes use of microstrip components, including a Lange coupler, a RF short and a low pass filter. However, the Ka band mixer, finline-microstrip combination is used. RF and LO are both fed in through finline tapers and the rest is microstrip.

Chapter 2 Microstrip Components of the Mixers

Figure 2.3 The circuit pattern of the Ka band mixer

In the rest of this chapter, the design and simulation of the microstrip components of the two mixers will be discussed. The finline components will be discussed in Chapter 3. The complete structure and performance analysis of those mixers will be discussed in Chapter 4.

2.2 Schottky diodes

For the Ku band mixer, two diodes follow a Lange coupler to make mixing operation in opposite polarities and then RF short and LPF are used to take out the desired components. Also, two diodes are connected to the hybrid of the unilateral finline taper and microstrip coupler for the Ka band mixer to realize the single balanced mixer structure. All diodes used here are Schottky diodes since they have low impedance at high frequency and low transit time, which are suitable for high frequency operation. They help to mix the two signals (RF and LO) to give rise to the IF signal along with a number of spurious frequencies.

Chapter 2 Microstrip Components of the Mixers

Schottky mixer diodes have attracted considerable attention in microwave mixers [10, 13-14]. Properly designed Schottky diodes are rectified by majority carrier action and therefore do not have the stored charge that prevents the use of PN semiconductor junction diodes as high frequency mixers. The speed of majority carrier diodes is limited by the RC time constant of the diode. By proper design, which includes the use of a thin high resistivity epitaxial layer, diodes can work at millimeter frequencies [11]. The design of Schottky diodes is straight forward. Fabrication techniques for accomplishing the design are the key to making useful diodes.

The basic operation of the Schottky diode is based on the energy band theory in semiconductors. Schottky diode, unlike the p-n junction diode, is formed, by bringing the Fermin energy levels of a metal and a semiconductor to the same level. The band energy diagrams of the metal and semiconductor, before and after they are brought into contact, are shown in Figure 2.4.

Figure 2.4 Energy levels of metal and semiconductor (a) before contact and (b) after contact

Chapter 2 Microstrip Components of the Mixers

In the diagram E_{FM} represents Fermi energy level of the metal and E_{FS} represents that of the semiconductor. It can be seen from the above figure that the electrons presented in the conduction band of the semiconductor always face an energy barrier after the contact is made. As a result the electrons cannot flow from the semiconductor side to the metal side. But on the other hand, the energy difference between the conduction band and the Fermi energy level of the metal is always constant (shown by the double headed arrow) before and after the junction is made. So electrons can flow from the metal side to the semiconductor side as the metal has got an excess pool of electrons.

Schottky diodes are preferred to the normal p-n junctions because the Schottky diodes do not have the problem of storage and junction capacitances, which dampen the switching speed to diodes when the signal changes from the positive to the negative cycle.

Two diodes models, M/A COM's MA4E2037 and Alpha's DMK2606, are readily available and can be used in this design. MA4E2037 is selected for this project.

2.3 Lange coupler

A popular implementation of the quadrature hybrid in microstrip line form is the so-called Lange coupler. It consists of multiple edge coupled interlaced lines with the alternate lines connected electrically in the middle and at the ends [15]. A single ground plane, a single dielectric, and a single layer of metallization are used. Normally it is a four-strip configuration. Additional variations involve six- and eight-strip realizations.

Chapter 2 Microstrip Components of the Mixers

Comparing to branch line couplers, which consists of direct quarter-wave connections between the two main lines of the coupler, Lange coupler has wide bandwidth [16]. Also unlike branch line couplers, Lange coupler has more parameters to optimize, such as width (W), spacing (S), length (L) and bonding wires.

Figure 2.5 A Lange coupler

Typical coupling values range between -5 and -1 dB. By choosing the length of the microstrip elements appropriately, a very broadband bandwidth of up to 40% bandwidth has been achieved [17]. There are two basic types: 180° hybrid and 90° hybrid. The latter is also called a 3dB directional coupler. For a 3dB coupler, a single wave incident in port 1 couples equal power into port 2 and 4, but no into port 3. In another words, it has a voltage-coupling coefficient of 0.707.

The outstanding features of Lange coupler are its compact size and the relatively large line separation when compared with the gaps of a conventional two-coupled line device and its relatively large bandwidth when compared with branch-line couplers. It is an ideal component for balanced MIC amplifiers and mixers, and for binary power divider trees. Lange coupler is often used as an input coupler in balanced microwave amplifier circuits. For this application it is designed as a 3dB coupler and the output signals are in phase quadrature, so that it is a 90° hybrid junction. The main

Chapter 2 Microstrip Components of the Mixers

disadvantage is that lines are very narrow, close together and it is difficult to fabricate the necessary bonding wire between the lines.

2.3.1 Design of Lange coupler for the Ku band mixer

The design and analysis of Lange coupler have been studied in [18-20]. In this project, Presser's work [21], Figure 2.6 and Figure 2.7, is selected for the initial design due to its simplicity. Figure 2.6 shows the ratios of W/h and S/h corresponding to the dielectric constant. It means that the width (W) and spacing (S) of a Lange coupler can be obtained as long as the dielectric constant and height (h) of a substrate are given using this figure.

Figure 2.6 Aspect ratios of 3 dB Lange coupler against dielectric — work of Presser

For the Ku band mixer, the substrate is alumina with height 25 mil and dielectric constant 9.6. Based on Presser's works, as shown in Figure 2.6, the ratios of W/h and S/h are as 0.114 and 0.0845 respectively. Since the substrate thickness, h , is 25 mil, therefore W is 2.85 mil and S is 2.1125 mil.

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In addition to the width (W) and spacing (S), there is another parameter, length (L), needed to be determined for a Lange coupler. It is basically the quarter wavelength at center frequency of the specified frequency range. Figure 2.7 is used to get this value. However, it can only get the even- and odd-mode velocities for the corresponding dielectric constant, not as directly as the previous one. Before using the figure, the even- and odd-mode impedance should be calculated first. In order to get the even- and odd-mode impedance, the voltage-coupling coefficient k is calculated as below.

$$k = \log_{10}^{-1}(A/20)$$

Where A is the desired attenuation in dB and it is 3 dB in this design.

Then

$$k = \log_{10}^{-1}(3/20) = 0.707$$

The corresponding even- and odd-mode impedance Z_{oe} and Z_{oo} are calculated in order to use Figure 2.7.

$$\begin{aligned} Z_{oe} &= Z_o[1 + k/(1-k)]^{0.5} \\ &= 50[1 + 0.707/(1-0.707)]^{0.5} \\ &= 120.7 \, \Omega \end{aligned}$$

where Z_o is the characteristic impedance of the system, 50 Ω .

In this example,

$$Z_{oo} = Z_o^2/Z_{oe} = 20.7 \, \Omega$$

With the even- and odd-mode impedances known, the next step is to get the corresponding velocities for dielectric constant 9.6 using Figure 2.7. And then a

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formula is used to get the effective relative dielectric constant. Finally the quarter wavelength inside this substrate can be determined.

From Figure 2.7, even mode velocity, V_e , is 1.204×10^8 m/s and odd mode velocity, V_o , is 1.318×10^8 m/s. Therefore, the even- and odd-mode effective relative dielectric constants are

Figure 2.7 Mode velocities of 3 dB Lange coupler against dielectric constant

$$\begin{aligned}\epsilon_{re} &= (c/V_e)^2 \\ &= [(3 \times 10^8)/(1.204 \times 10^8)]^2 \\ &= 6.2085\end{aligned}$$

$$\begin{aligned}\epsilon_{ro} &= (c/V_o)^2 \\ &= [(3 \times 10^8)/(1.318 \times 10^8)]^2 \\ &= 5.181\end{aligned}$$

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The relative effective dielectric constant for the substrate with dielectric constant 9.6 is

$$\epsilon_{\text{eff}} = [\epsilon_{\text{re}}\epsilon_{\text{ro}}]^{0.5} = 5.67.$$

The quarter wavelength of the Lange coupler at Ku band is calculated below.

For Ku band, given $c = 3 \times 10^8$ m/s and $f = 20$ GHz

$$\lambda = c/f/(\epsilon_{\text{eff}})^{0.5} = 0.63 \text{ cm}$$

$$1 \text{ inch} = 1000 \text{ mil} = 2.54 \text{ cm}$$

Therefore $\lambda = 248.03$ mil and $L = \lambda/4 = 62.01$ mil.

2.3.2 Simulation and optimization of the Lange coupler in ADS

Advanced Design System (ADS) is used for microstrip components' simulation in this project. The part of the electromagnetic field analysis in ADS is Momentum. Momentum is based on the numerical discretization technique, method of moment (MoM). It is used to solve Maxwell's electromagnetic equations for planar structures embedded in a multilayered dielectric substrate. The simulation modes available in Momentum, which are available for both RF and microwave simulations of ADS, are both based on this technique, but use different technologies to achieve their results.

Momentum uses specific steps to achieve solutions. The technology used by each simulation mode varies slightly in each step. In Momentum's microwave mode, the planar metallization and aperture patterns in the signal layers are meshed using rectangular and triangular cells. Momentum RF mode generates a mesh of polygonal cells. The unknown electric and magnetic surface currents are discretized using rooftop expansion function defined over the cells in the mesh. In Momentum for RF simulation, rooftop functions are regrouped in star and loop functions to eliminate low-frequency breakdown. This discretization process is carried out to transform the

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maxwell's equation to a matrix equation for numerical calculation. The microwave mode uses fullwave Green's functions resulting in L and C elements that are real and frequency independent.

Momentum optimization employs a comprehensive set of optimizers including L1, L2 (least squares), minimax, quasi-Newton, random and simulated annealing algorithms. All of these have been adapted specifically to handle engineering design problems.

The first four optimizers are so-called gradient-based. They need first-order partial derivatives in addition to the value of the objective function to make a decision on the subsequent steps. They usually move fairly quickly towards a minimum particularly if one exists close to the starting point. If possible, they tend to search fairly limited areas of available exploration regions and are susceptible to traps of local minima. The last two optimizers do not require gradient information. Since they normally use a large number of steps, caution is needed when trying to use them in conjunction with more time consuming electromagnetic simulations.

The first four optimizers have been specifically developed for minimization of their corresponding objective functions. Therefore, their performance in handling their objective functions is likely to be far superior over any other optimizer. The last three optimizers are general-purpose techniques [22].

Simulation setup and result of the Lange coupler are given in Figure 2.8. The four ports of the Lange coupler were all terminated with 50 ohm loads. And port 1 and 3 were used for RF and LO inputs respectively. To achieve a single balanced mixer, both RF and LO power should be separated equally and signals should have 90°

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phase shift, which means S21 and S41, S23 and S43, are all supposed to be 3 dB over the required frequency band. However, only S21 and S41 are presented and discussed here since the Lange coupler will ensure S23 and S43 to be the required values as long as S21 and S41 meet the requirements due to its own structure features.

Figure 2.8 Simulation result of the Lange coupler based on calculated values at Ku band

Obviously, the coupling ratio, S21, and directivity, S41, were both close to the desired 3 dB as shown in Figure 2.8. However, optimization can further improve the Lange coupler's performance. It helped to remove the uncertainty and inaccuracy of the results after the simulation was done using the initial calculated values. The optimized values were updated until the desired performance was obtained.

In this simulation, the width and gap and also the length of the Lange coupler were given a small range for adjustment during optimization so that the appropriate

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changes can be made in the design in order to meet the required specification. To obtain optimum performance, goals were set for S21 and S41 in the required frequency range. This resulted in minor design changes.

Figure 2.9 Optimization setup of the Lange coupler at Ku band

The optimization setup for the Lange coupler is shown in Figure 2.9. The initial values of Lange coupler were W=2.85 mil, S=2.1125 mil and L=62.01 mil. After optimization, W was 2.7 mil, S was 2.268 mil and L was 64.93 mil. The performance comparison before and after optimization is shown in Figure 2.10.

Figure 2.10 S21 and S41 of the Lange coupler before and after optimization

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The Opt_S21 and Opt_S41 marked the optimized results and the P_S21 and P_S41 marked the previous ones. Obviously, S21 and S41 were much closer to 3dB over the required frequency range after optimization. Also 90° phase shift is necessary to realize the single balanced mixer, which means the signals, coming out from port 2 and port 4, must have 90° phase difference. The simulation results are shown in Figure 2.11. The phase difference between S21 and S41 was always 90° over the required frequency range. Therefore, the Lange coupler satisfies the required performance, equal power separation and 90° phase shift.

Figure 2.11 Phase difference of two output ports of the Ku band Lange coupler

2.4 Low pass filter

A filter is often a frequency-selective device, which passes signals of certain frequencies and blocks or attenuates signals of other frequencies. A frequency-selective filter is one that passes signals whose frequencies are in certain range or bands, called the passband, and blocks or attenuates signals whose frequencies are in other ranges, called the stopbands. Passbands are the frequency bands occupied by the wanted signal. In these bands, the ideal requirement for the filter is to provide constant loss and constant envelope delay so that the wanted signal will be

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transmitted without distortion. On the other hand, frequency bands occupied by the unwanted signal are referred to as the stopbands. In these bands, the common form of specification merely requires the loss, relative to the lower limit set for the passband, to be equal or greater than some minimum amount. The most often encountered types of frequency-selective filters are, low pass filter, high pass filter and bandpass filter.

In this project, a low pass filter is used to extract the desired IF components. There are two main classes of tabulated approximations to the low pass filter characteristics. They are classified according to the manner in which the function approximates the constant attenuation specifications in the passband and stopband.

The two types of approximation are:

- Butterworth – this approximation provides a maximally flat passband (close to zero frequency) and a monotonically increasing attenuation function in the stopband.
- Chebyshev – this approximation provides an equip-ripple passband and a monotonically increasing attenuation function in the stopband.

2.4.1 Low pass filter design using lump components

Design Specifications

The following are the design specifications for the Ku and Ka band mixers' low pass filters.

Filter type: Chebyshev (Equi-ripple)

Ripple height, L_{ar} : 0.1 dB (set by integer)

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Cutoff frequency (f_c): $f_c = 7.5$ GHz (Ku band) and $f_c = 10$ GHz (Ka band)

Minimum Stopband(f_s), L_{as} : 15 dB at $f_s=10$ GHz(Ku band) and 20 dB $f_s=15$ GHz(Ka band)

Determining the filter order

The basic formula governing the filter order, the cutoff frequency, the stopband frequency and the minimum stopband is the following,

$$L_{as} \leq 10 \log_{10} \{1 + \epsilon \cosh^2[n \cosh^{-1}(\omega_s/\omega_c)]\}$$

Given that

$$\epsilon = 10^{L_{ar}/10} - 1$$

This can be simplified to make the variable 'n' as the subject to the following form,

$$n \geq [1/ \cosh^{-1}(\omega_s/\omega_c)] \cosh^{-1} \{(10^{L_{as}/10} - 1)/(10^{L_{ar}/10} - 1)\}^{1/2}$$

Now, putting the values of,

$$\omega_s = 2\pi f_s$$

$$\omega_c = 2\pi f_c$$

$$L_{ar} = 0.1 \text{ dB}$$

$$L_{as} = 15 \text{ dB (Ku band) and } 20 \text{ dB (Ka band)}$$

For Ku band, the cut off frequency is chosen to be 7.5 GHz and the stop frequency is 10 GHz.

The value of n is determined as $n \geq 5.3857$.

For Ka band, the cut off frequency is chosen to be 10 GHz and the stop frequency is 15 GHz.

The value of n is determined as $n \geq 5.0606$.

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But n has to be an integer. Therefore,

$n = 6$ for both Ku and Ka band.

Determining filter prototype element values

From the table [23] for Chebyshev filter with $n = 6$, $g_o = 1$, $w'_1 = 1$ and 0.1 dB ripple, the following values for the various parameters are determined, which can be verified using the formulas that follow the values of the parameters.

$$g_0 = 1, g_1 = 0.517, g_2 = 1.414, g_3 = 1.932, g_4 = 1.932, g_5 = 1.414, g_6 = 0.517 \text{ and } g_7 = 1$$

The following formulae, which can be used to verify the above parameters, are

$$\begin{aligned} \chi &= (1/n) \sinh^{-1}(1/\epsilon) \text{ where } \epsilon = 10^{(1\text{ar}/10)} - 1 \\ g_1 &= [1/(\sinh \chi) g_0] 2 \sin(\pi/2n) \\ g_{k,k-1} &= [1/\sinh^2 \chi + \sin(2\pi/n)] \{4 \sin[(2^{k-1} - 1)\pi/2n] \sin[(2k - 1)\pi/2n]\} \\ g_{n+1} &= 1 \quad (\text{for } n = \text{odd}) \end{aligned}$$

The corresponding capacitance and inductance can be got using the following formulae,

$$C_n = g_n/w_c Z_o$$

$$L_n = g_n Z_o/w_c$$

The prototype for the designed LPF is shown in Figure 2.12. There are totally 6 elements since n is 6 for both the Ku and Ka band mixers.

Figure 2.12 The prototype of tge low pass filter using LC components

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The L and C values of the LPFs for the Ku and Ka band mixers are calculated as below.

For the Ku band mixer,

$$L1 = 0.55 \text{ nH}, C1 = 0.6 \text{ pF}, L2 = 2.05 \text{ nH}, C2 = 0.82 \text{ pF}, L3 = 1.5 \text{ nH}, C3 = 0.22 \text{ pF}.$$

For the Ka band mixer,

$$L1 = 0.4 \text{ nH}, C1 = 0.45 \text{ pF}, L2 = 1.537 \text{ nH}, C2 = 0.615 \text{ pF}, L3 = 1.125 \text{ nH}, C3 = 0.165 \text{ pF}.$$

2.4.2 Microstrip implementation for the LPFs

For both the Ku and Ka band mixers, the microstrip line is used to realize the low pass filters. The detail calculation will be given. Normally, a short section of transmission line with length l can be represented by a T-equivalent circuit, which is shown in Figure 2.13. Z_0 is the characteristic impedance and β is the propagation constant.

Figure 2.13 Approximate equivalent circuits for short sections of transmission lines.

(a) Equivalent circuit for a transmission line section having $\beta l \ll \pi/2$.

(b) Equivalent circuit for small βl and large Z_0 .

(c) Equivalent circuit for small βl and small Z_0 .

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(i) If $\beta l < \pi / 2$,

The series elements have a positive reactance (inductors)

$$X/2 = Z_o \tan (\beta l / 2) \quad (2.1)$$

and shunt element has a negative reactance (capacitor)

$$B = \sin \beta l / Z_o \quad (2.2)$$

(ii) If $\beta l < \pi / 4$ and a large characteristic impedance Z_h ,

(1) and (2) approximately reduce to an equivalent of a series inductor

$$\begin{aligned} X_L &\cong Z_o \beta l = Z_h \beta l \\ B &\cong 0 \end{aligned} \quad (2.3)$$

(iii) If $\beta l < \pi / 4$ and a small characteristic impedance Z_l ,

(1) and (2) approximately reduce to an equivalent of a shunt capacitor

$$\begin{aligned} X &\cong 0 \\ B_C &\cong Y_o \beta l = \beta l / Z_l \end{aligned} \quad (2.4)$$

The value of $\beta = 2\pi/\lambda$ in (2.3) and (2.4) is calculated at $\omega = \omega_c$ (i.e., $\lambda = \lambda_c$) in order to get the best response near the cutoff frequency. Considering impedance scaling (denormalization), the electrical lengths in (2.3) and (2.4) are given by

$$\begin{aligned} \beta l &= L R_o / Z_h \quad (\text{for inductor section}) \\ \beta l &= C Z_l / R_o \quad (\text{for capacitor section}) \end{aligned} \quad (2.5)$$

where

R_o = impedance scaling factor

L and C = normalized element values (g values of the low pass filter prototype)

For microstrip, the most important parameters are strip width (w), substrate height (h) and its relative dielectric constant (ϵ_r). Since the structure is not uniform, it supports a

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Quasi-TEM mode. Analysis of microstrip lines can be tedious and hence design equations are made available. Assuming a strip thickness of $t=0$, (for $t/h < 0.005$, effect of strip thickness is negligible), the analysis formulas are the following with given w/h and ϵ_r .

(i) For $w/h \leq 1$

$$Z_o = \frac{60}{\sqrt{\epsilon_{eff}}} \ln \left[\frac{8h}{w} + 0.25 \frac{w}{h} \right] \quad (2.7)$$

$$\epsilon_{eff} = \frac{(\epsilon_r + 1)}{2} + \frac{(\epsilon_r - 1)}{2} \left(1 + 12 \frac{h}{w} \right)^{-0.5} + 0.04 \left(1 - \frac{w}{h} \right)^2$$

(ii) For $w/h \geq 1$

$$Z_o = \frac{120\pi}{\sqrt{\epsilon_{eff}}} \left\{ \frac{w}{h} + 1.393 + 0.667 \ln \left[\frac{w}{h} + 1.444 \right] \right\}^{-1} \quad (2.8)$$

$$\epsilon_{eff} = \frac{(\epsilon_r + 1)}{2} + \frac{(\epsilon_r - 1)}{2} \left(1 + 12 \frac{h}{w} \right)^{-0.5}$$

In this design, characteristic impedance selected for inductance is $Z_h=80$ ohm and for capacitance is $Z_l=20$ ohm. With the impedance value and given relative dielectric constant, the corresponding w/h and effective relative dielectric constant can be calculated using the analysis formulas, equation (2.7) and (2.8). For the Ku band mixer, the substrate is with height $h=25$ mil and relative dielectric constant $\epsilon_r=9.6$. And for the Ka band mixer, h is 10 mil and ϵ_r is 2.22 for the substrate. Except the width of the microstrip line, the length is another parameter that should be designed. The guided quarter wavelength at the cutoff frequency is also calculated using the ready effective relative dielectric constant. The detail calculated values for both the Ku and Ka band are given in Table 2.1.

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Table 2.1 Width and guided wavelength using microstrip

	Z_o	w/h	w (mm)	$\sqrt{\epsilon_{eff}}$	λ_o at 7.5 GHz (Ku) 10 GHz (Ka)	λ_g , mm at 7.5GHz 10 GHz
Ku band $\epsilon_r=9.6$ h = 25 mil	20 ohm	4.316	2.7407	2.7407	40 mm	14.6 mm
	80 ohm	0.306	0.1944	2.4082	40 mm	16.61 mm
	50 ohm	0.987	0.6268			
Ka band $\epsilon_r=2.22$ h = 10 mil	20 ohm	10.2191	2.6	1.4226	30 mm	21.09 mm
	80 ohm	1.4121	0.36	1.3446	30 mm	22.311 mm
	50 ohm	3.0891	0.78			

The following equations are used to find the electrical length of the inductive (characteristic impedance $Z_h=80$ ohm) and capacitive (characteristic impedance $Z_l=20$ ohm) transmission line sections to replace the series inductor and shunt capacitor. R_o is the 50 ohm. The calculated values are given in Table 2.2.

$$\begin{aligned} \beta l_1 &= g_1 R_o / Z_h = 0.3231 & \beta l_4 &= g_4 Z_l / R_o = 0.7728 \\ \beta l_2 &= g_2 Z_l / R_o = 0.5656 & \beta l_5 &= g_5 R_o / Z_h = 0.8838 \\ \beta l_3 &= g_3 R_o / Z_h = 1.2075 & \beta l_6 &= g_6 Z_l / R_o = 0.2068 \end{aligned}$$

Table 2.2 Various length of microstrip section

	l_1 (mm)	l_2 (mm)	l_3 (mm)	l_4 (mm)	l_5 (mm)	l_6 (mm)
Ku band $\epsilon_r=9.6$ h = 25 mil	0.0514λ	0.09λ	0.1922λ	0.123λ	0.1407λ	0.0329λ
	0.8537	1.3135	3.1924	1.7952	2.337	0.48
	$w_1=0.1944$	$w_2=2.7407$	$w_3=0.1944$	$w_4=2.7407$	$w_5=0.1944$	$w_6=2.7407$
Ka band $\epsilon_r=2.22$ h = 10 mil	0.0514λ	0.09λ	0.1922λ	0.123λ	0.1407λ	0.0329λ
	1.147	1.898	4.29	2.6	3.14	0.7
	$w_1=0.36$	$w_2=2.6$	$w_3=0.36$	$w_4=2.6$	$w_5=0.36$	$w_6=2.6$

2.4.3 Simulation of low pass filters at Ku and Ka band

Both the Ku and Ka band mixers need a LPF to extract the desired IF components, while attenuate harmonics at the same time. For the Ku band mixer, the various microstrip sections are designed on alumina substrate with height 25 mil and dielectric constant 9.6. The frequency range is set from 0.01 to 12 GHz for simulation in ADS.

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Ku band

Figure 2.14 The simulation result of the LPF at Ku band using microstrip

As shown in Figure 2.14, the insertion loss, S21 in dB, of the low pass filter was given for both the lump component filter and the microstrip filter. The Lump_S21 (for lump component filter) 3dB cutoff is at 7.5 GHz and has 15.123 dB attenuation at 10 GHz. The Mline_S21 3dB cutoff is at 7 GHz and has 14.977 dB attenuation at 10 GHz. The small difference is due to the approximated transformation from lump components to microstrip lines, which caused the microstrip lines have no exact admittance or reactance as the lump components near and above the cutoff frequency.

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Similarly, for the Ka band mixer, the various microstrip sections, designed in Table 2.2, are based on Duroid RT5880 substrate with height 10 mil and dielectric constant 2.22. The frequency range is set from 0.01 to 20 GHz for simulation in ADS.

Ka band

Figure 2.15 The simulation result of the LPF at Ka band using microstrip

As Figure 2.15 shows, Lump_S21 and Mline_S21 represent the insertion loss of the lump component filter and microstrip line filter respectively. The Lump_S21 3dB cutoff is at 10 GHz and has 21.122 dB attenuation at 15 GHz, which is agreed with the design specification. The Mline_S21 cutoff is at 9.5 GHz and has 16.294 dB attenuation at 15 GHz. The small difference between them is due to the same reason of the Ku band LPF.

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3.1 Finline structure

Finline was originally used for the purpose of orthogonal mode launching in circular waveguides [24]. Over the past decades, the interest in millimeter wave technology has grown significantly and so has the search for suitable transmission media. The principal characteristics required of such media are large bandwidth, low ohmic and radiation losses, compatibility with semiconductor devices with possibility for circuit integration and ease, flexibility and reliability of fabrication and design in large and small quantities. While many of these requirements are individually met by other conventional forms of transmission lines, such as waveguides, microstrip and image lines, finlines offer a number of distinct advantages over most of these media.

Figure 3.1 Cross-section of several finlines (a) unilateral, (b) insulated, (c) bilateral and (d) antipodal

Although waveguides have been available and used for a long time and have relatively low losses, they are not amenable to integrated circuit fabrication. Microstrip, slot-line

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and coplanar lines have been used extensively in integrated circuits. However, these suffer from practical problems of tolerance requirements with the very narrow strip widths required at millimeter wavelengths, are often incompatible with hybrid devices and have high ohmic and dielectric losses. Multimode is also a concern.

Some of the most commonly used finline structures are shown in Figure 3.1 above. The finline can be considered as a shielded slot-line, with the finline being quasiplanar and suspended in the E-plane of a rectangular waveguide. Thus, in a given frequency band, the finline dimensions are identical to that of a commensurate waveguide. The structures shown may be regarded as a dielectric, slab-loaded waveguide, with printed fins on the slab, or as a ridged waveguide with zero-thickness ridges backed by a dielectric slab in the E-plane. A better appreciation of the unilateral, bilateral and antipodal structures can be seen from Figure 3.2. In these structures, the suspended fins concentrate the field energy

Figure 3.2 Three-dimensional views of the three commonly used finlines

in the fin-gap region as in Figure 3.3, leading to the capacitive loading of the dominant HE mode of propagation in the slab-loaded waveguide. This has the effect

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of lowering the cutoff frequency of the fundamental mode to a considerable degree and that of the next higher mode to a very small degree, thus leading to a larger bandwidth of operation for the fundamental mode.

Figure 3.3 Distribution of transverse electric fields (a) unilateral, (b) bilateral and (c) antipodal

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Besides the larger bandwidth that is available, any devices that are connected to the fins for integrated circuit fabrication will be subjected to wider taper impedance selection, resulting in better matching. However, the field concentration will also result in larger conduction and dielectric losses, due to the presence of the dielectric slab and the higher current density near the fin-edges. The attenuation in finline is typically of the order of 0.1 dB/wavelength, and therefore finline is unsuitable for long distance power transmission.

Among the configurations shown in Figure 3.2, the unilateral finline is the simplest and best suited for the fabrication of finline components. The metallization on both sides of the substrate in bilateral finline produces lower transmission loss and greater flexibility in biasing active devices as in mixers with antiparallel switched diodes and matched PIN diode attenuators. Bilateral finline offers characteristic impedance as low as 100 ohms. Antipodal finlines on the other hand offer impedance levels of the order of 10 ohms and are suitable for transitions between a microstrip and a waveguide. The typical impedance transformation ratio is 20:1.

The growing popularity of microwave integrated circuits (MIC) and the advantages offered by finlines in terms of concentration of fields on small diode dimensions, possibility of multiband device fabrication, possibility of high level of integration, greater flexibility than monolithic devices, suitability for small size and mass production and low cost circuit design and development have resulted in realization of virtually all important circuit components in finline, including RF receive/transmit front-ends. In what follows, the analysis techniques for finlines are discussed, together with a synthesis procedure. Tapers' characteristics and design data are presented.

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The technique used to design and simulate the exponential unilateral finline taper is based on the close-form expressions for the power-voltage definition of the characteristics impedance [25]. Although this method is mathematically rigorous, it does not require solutions to complex eigenvalue boundary condition equations. The closed-form expressions enable the reflection coefficients of the exponential taper to be generated and the minimum value at $l/\lambda=2$ will be used to design a unilateral exponential finline taper, where l is the total length of the taper and λ is the wavelength at the selected frequency. More information related to finline taper design can be found in [26-28].

3.2 Finline taper design

3.2.1 Closed-form expressions

The closed-form expressions consist of analysis and synthesis equations and require the impedance schematic representation of the finline taper in order to determine the change in impedance due to an incremental change in distance along the finline. With this approach, reflection coefficient can be generated for finline matching. The details of closed-form expressions can be obtained from literature [25] and are reproduced in the following sections.

The closed-form analysis is by evaluating the equivalent circuit of the finline taper as shown in Figure 3.4. In Figure 3.4(a), the impedance transition of a finline taper is realized by the various fin gap from b to w along the taper and ended at $z = l$ (2λ). When $z=0$ and fin gap $d=b$, the characteristic impedance is z_2 and when $z=l$ and fin gap $d=w$, the characteristic impedance is z_1 , which is normally the device impedance the taper is going to match.

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Figure 3.4 Representation of a taper line transformer

(a) and (b) Taper finline and (c) Impedance schematic representation

Figure 3.4(b) shows the cross section when the taper is inside a waveguide. The width and height of the waveguide are a and b respectively. The thickness of the substrate is s and the fin gap is d . The dimension of the waveguide should be carefully selected to satisfy the design requirements.

The taper design is actually to determine the impedance transition along the taper as shown in Figure 3.4(c). The transition may follow some contour functions.

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Optimization of the taper involves choosing a proper contour function, which offers the minimum reflection coefficient for a given taper length or vice versa. Knowledge of the finline impedance characteristics is necessary in all of the formulas.

In order to get the impedance transition along the taper, the relationship between characteristic impedance and fin gap should be clear first. The characteristic impedance Z_0 and the propagation constant β of the finline taper are defined in [12]. And the parameters using in the following equations can refer to Figure 3.4.

$$Z_0 = \frac{240\pi^2(px + q)\left(\frac{b}{a}\right)}{(0.385x + 1.762)^2 \sqrt{\epsilon_c(f)}} \quad (3.1)$$

$$\beta = \frac{2\pi}{\lambda} \sqrt{\epsilon_c(f)} \quad (3.2)$$

for $d/b > 0.3$

$$p = -0.763\left(\frac{b}{\lambda}\right)^2 + 0.58\left(\frac{b}{\lambda}\right) + 0.0775\left[\ln\left(\frac{a}{s}\right)\right]^2 - 0.668\ln\left(\frac{a}{s}\right) + 1.262 \quad (3.3)$$

$$q = 0.372\left(\frac{b}{\lambda}\right) + 0.914 \quad (3.4)$$

for $d/b \leq 0.3$

$$p = 0.17\left(\frac{b}{\lambda}\right) + 0.0098 \quad (3.5)$$

$$q = 0.138\left(\frac{b}{\lambda}\right) + 0.873 \quad (3.6)$$

$$x = -\ln \sin\left(\frac{0.5\pi d}{b}\right) \quad (3.7)$$

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As given in equation (3.1), the effective dielectric constant is also necessary for the calculation and it is defined in [25] as follow.

$$\epsilon_e(f) = k_e - \left(\frac{\lambda}{\lambda_{ca}}\right)^2 \quad (3.8)$$

where

f = the operating frequency

λ = the operating wavelength

λ_{ca} = the cutoff wavelength of the finline for $\epsilon_r = 1$

ϵ_r = the substrate dielectric constant, and

k_e = the equivalent dielectric constant of the finline at the frequency f , and is given by the following equations

$$k_e = k_c + \frac{k_1 - k_c}{\left(\frac{b}{\lambda_1} - \frac{b}{\lambda_{cf}}\right)} \left[\frac{b}{\lambda} - \frac{b}{\lambda_{cf}} \right] \quad (3.9)$$

$$k_c = 1 + \left(\frac{s}{a}\right)(a_1 x + b_1)(\epsilon_r - 1) \quad (3.10)$$

For $d/b > 0.5$

$$a_1 = -0.5148 \left[\ln\left(\frac{a}{s}\right) \right]^2 + 4.2145 \left[\ln\left(\frac{a}{s}\right) \right] - 5.09 \quad (3.11)$$

$$b_1 = 0.0452 \left[\ln\left(\frac{a}{s}\right) \right] + 1.8257 \quad (3.12)$$

For $d/b \leq 0.5$

$$a_1 = 0.4021 \left[\ln\left(\frac{a}{s}\right) \right]^2 - 0.7685 \left[\ln\left(\frac{a}{s}\right) \right] + 0.3921 \quad (3.13)$$

$$b_1 = 2.42 \sin \left[0.556 \ln\left(\frac{a}{s}\right) \right] \quad (3.14)$$

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$$\lambda_{cf} = \lambda_{ca} \sqrt{k_c} \quad (3.15)$$

$$\lambda_{ca} = 2a \sqrt{1 + \frac{4b}{\pi a} \left[1 + 0.2 \sqrt{\frac{b}{a}} \right] x} \quad (3.16)$$

$$k_1 = 1 + \left(\frac{\lambda_1}{\lambda_{ca}} \right)^2 \quad (3.17)$$

$$\frac{\lambda_1}{\lambda_{ca}} = \pi \sqrt{\frac{x'(3-2x')(\epsilon_r-1)}{12}} \quad (3.18)$$

where

$$x' = s_1 + \frac{q_f}{3} - \sqrt[3]{s_1^3 - \frac{2q_f}{\pi^2}} \quad (3.19)$$

$$s_1 = \sqrt[3]{\sqrt{0.0666 + 0.0466q_f^2 + 0.015q_f^4 - 0.000137q_f^6} + \frac{q_f}{\pi^2}} \quad (3.20)$$

$$q_f = \frac{k_c - 1}{\epsilon_r - 1} \quad (3.21)$$

The previous equations can be used to figure out the characteristic impedance versus fin gap, except that the characteristic impedance change with respect to the length along the taper is also necessary for the design. The taper profiles, which are most commonly used to design the taper impedance transformer, are given below and the exponential taper profile is used for this design.

(a) Exponential

The exponential taper profile is defined by

$$z(z) = z_2 \exp \left[\left(\frac{z}{l} \right) \ln \left(\frac{z_1}{z_2} \right) \right] \quad (3.22)$$

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(b) Parabolic

The Parabolic profile is defined by

$$z(z) = \left\{ \sqrt{z_2} + \frac{z}{1} \left[\sqrt{z_1} - \sqrt{z_2} \right] \right\}^2 \quad (3.23)$$

(c) Cosine taper

The cosine taper profile is defined by

$$z(z) = z_2 \cos \left[\frac{z}{1} \cos \left(\frac{z_1}{z_2} \right) \right] \quad (3.24)$$

(d) Cosine-squared

The cosine-squared taper profile is defined by

$$z(z) = z_2 \cos^2 \left[\frac{z}{1 \cos \left(\frac{z_1}{z_2} \right)} \right] \quad (3.25)$$

The detail design process is very tedious since it involves a lot of equations and mathematics. Therefore, a Matlab program is written to design the finline tapers. The final taper designs are shown in Figure 3.5. The substrate is with dielectric constant 2.22 and height 10 mil. The unilateral finline taper in Figure 3.5(a) is for RF input, which is symmetrical about the x axis. Therefore, only the dimension of taper 1 is given on the right side. The antipodal finline taper in Figure 3.5(b) is for LO input. Taper 3 is solid line while taper 4 is dash line as shown in the figure. The reason is that taper 3 is on the front side of the substrate while taper 4 is on its backside. If the thickness of the substrate is neglected, taper 3 and taper 4 are actually symmetrical about the x axis, too. Therefore, only the dimension of taper 3 is given.

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Figure 3.5 The designed finline tapers (a) unilateral and (b) antipodal

3.3 Waveguide selection — WR28

As discussed at the beginning of this chapter, the finline can be actually considered as a shielded slot-line, with the finline being quasiplanar and suspended in the E-plane of a rectangular waveguide. Therefore the waveguide should be carefully selected to ensure its operating frequency is in the required range.

A metal rectangular waveguide basically consists of a single hollow conductor that can propagate electromagnetic energy above certain cutoff frequency and it allows infinite

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number of propagation modes. The mode of propagation can be divided into transverse electric (TE) and transverse magnetic (TM) mode. For TE mode, it has magnetic field but no electric field in the direction of propagation, whereas the TM mode has an electric field but no magnetic field in the direction of propagation. The signal propagation is usually in single mode propagation, hence the waveguide can be described as a transmission line with a propagation constant (γ) and a characteristic impedance Z_o . The propagation constant of a waveguide is unique and for simplicity the characteristic impedance is assumed to be equal to the wave impedance.

Assume that the waveguide is lossless and filled with a dielectric with a dielectric constant ϵ_r , the guide wavelength λ_g , free space wavelength λ_o , wavelength in dielectric λ_1 , and the cutoff wavelength λ_c can be related as follows [23],

$$\frac{1}{\lambda_1^2} = \frac{\epsilon_r}{\lambda_o^2} = \frac{1}{\lambda_g^2} + \frac{1}{\lambda_c^2} \quad (3.26)$$

Since the characteristic impedance is assumed to be equal to the wave impedance, it can be calculated by using the following equations,

$$Z_o = \begin{cases} \frac{377\lambda_g}{\sqrt{\epsilon_r}\lambda_1} & \text{TE mode} \\ \frac{377\lambda_1}{\sqrt{\epsilon_r}\lambda_g} & \text{TM mode} \end{cases} \quad (3.27)$$

The propagation phase constant β_t is given by

$$\beta_t = \frac{2\pi}{\lambda_g} \quad \text{radians / unit length} \quad (3.28)$$

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By substituting equation (3.26) into (3.27) and making use of the relationship $\lambda = u / f$, where u is the phase velocity and f is the frequency, another form of the characteristic impedance can be obtained.

$$Z_0 = \begin{cases} \frac{377}{\sqrt{\epsilon_r} \sqrt{1 - \left(\frac{f_c}{f}\right)^2}} (\Omega) & \text{TE mode} \\ \frac{377}{\sqrt{\epsilon_r}} \sqrt{1 - \left(\frac{f_c}{f}\right)^2} (\Omega) & \text{TM mode} \end{cases} \quad (3.29)$$

In order to obtain the cutoff frequency of the waveguide, it is required to solve for the electric field and magnetic field equations and with the boundary conditions, the cutoff frequency and wavelength can be found.

$$(f_c)_{mn} = \frac{1}{2\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2} \quad (\text{Hz}) \quad (3.30)$$

$$(\lambda_c)_{mn} = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}} \quad (\text{m}) \quad (3.31)$$

where a and b are width and height of a waveguide respectively, m and n are both integers, μ is permeability and ϵ is dielectric constant.

From the solution of the electric field (TM mode) and magnetic field (TE mode), for TM mode of propagation, neither m nor n can be zero and for TE mode, either m or n (but not both) can be zero. Therefore, the lowest cutoff frequency of the waveguide is the one when $m=1$ and $n=0$. Hence, TE_{10} mode has the lowest cutoff frequency and is known as the dominant mode.

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By using the above equations and conclusions, the calculations for the characteristic impedance and the lowest cutoff frequency of the WR28 waveguide can be found. For WR28, the dimension of waveguide are given as width $a=7.112$ mm and height $b=3.556$ mm. By using equation (3.30), the lowest cutoff frequency for TE_{10} mode is found to be around 21.09 GHz, which is suitable for Ka band operation. By using the calculated cutoff frequency in equation (3.29), the characteristic impedance of the waveguide at 35 GHz is found to be about 472.4Ω .

3.4 Simulation in high frequency simulation structure (HFSS)

The combination of finline and microstrip structure is used for the Ka band mixer design and is simulated by using HFSS. HFSS is an interactive software package developed by Ansoft for calculating the electromagnetic behavior of a structure. It calculates the full 3D electromagnetic field inside a structure by using the finite element method (FEM).

In general, the finite element method divides the full problem space into thousands of smaller regions (elements) and represents the field each element with a local function. In HFSS, the problem space is automatically divided into a large number of tetrahedral, where a single tetrahedron is basically a four-sided pyramid. This collection of tetrahedral is referred to as the finite element mesh.

The value of a vector field quantity (such as the H-field or E-field) at points inside each tetrahedron is interpolated from the values at the vertices of the tetrahedron. At each vertex, the HFSS stores the components of the field that are tangential to the three edges of the tetrahedron. In addition, the system can store the component of

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the field at the midpoint of selected edges that is tangential to a face and normal to the edge. The field inside each tetrahedron is interpolated from these nodal values. By representing field quantities in this way, the system can transform Maxwell's equations into a matrix equation that is solved using traditional numerical methods.

Various interpolation schemes (basic functions) can be used to interpolate field values from nodal values. A 1st order tangential element basis function interpolates field values from both nodal values at vertices and on edges. 1st order tangential elements have 20 unknowns per tetrahedron. A 0th order basis function makes use of nodal values at vertices only and therefore assumes that the field varies linearly inside each tetrahedron. 0th order tangential elements have six unknowns per tetrahedron.

There is a trade-off among the size of the mesh, the desired level of accuracy, and the amount of available computing resources. The accuracy of the solution depends on how small each of the individual elements is. Generating a field solution involves inverting a matrix with approximately as many elements as there are tetrahedral nodes. Therefore, it is desirable to use a mesh fine enough to obtain an accurate field solution but not so fine that it overwhelms the available computer memory and processing power [29].

3.4.1 Unilateral taper for the RF port of the Ka band mixer

The unilateral finline taper structure was designed for the RF input port in Section 3.3. The draft 3D drawing, including two fins, the substrate and WR28, is shown in Figure 3.6. The dielectric constant is 2.22 and thickness is 10mil for the substrate. The dimension of WR28 is 7.112mm x 3.556mm. The structure is basically an impedance transformer from WR28's 480 Ω to 200 Ω of two diodes in series. Therefore the

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insertion loss of the taper should be as small as possible since the RF power is not large normally. Its simulated insertion loss is given in Figure 3.7.

Figure 3.6 Draft 3D drawing of the unilateral taper

Figure 3.7 Insertion loss of the unilateral finline taper

As shown in Figure 3.7, the insertion loss over frequency range, 26.75 to 31.2 GHz, was better than 0.06 dB, which meant most of the input power can reach the diodes from WR28. Therefore, the desired performance was achieved.

3.4.2 Antipodal finline taper for the LO port of the Ka band mixer

The antipodal finline taper together with a microstrip coupler is used for the LO input port. The antipodal taper is also designed as an impedance transformer from WR28's 480Ω to the two parallel diodes' 50Ω. The microstrip coupler is used as an

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IF block and also for external bias of diodes if needed. Since the high frequency LO power is pumped in through this port, the insertion loss of this finline-microstrip structure should be small. Its 3D draft drawing is shown in Figure 3.8.

Figure 3.8 Draft 3D drawing of the antipodal finline taper and microstrip coupler

Similar to the case of the unilateral finline taper, the substrate has a dielectric constant of 2.22 and a height of 10 mil. The waveguide WR28 is also used. However, for the antipodal taper, one fin is on the front side of the substrate, while the other is on the backside. For the microstrip coupler, the ground plane should be there.

The insertion loss of this LO port is shown in Figure 3.9 below. The largest insertion loss was around 2.05 dB at the lower end of the band 26.55 to 31.2 GHz. It may be due to the narrow band property of the microstrip coupler.

Figure 3.9 Insertion loss of the antipodal finline taper with microstrip coupler

Chapter 4 Structure and Performance Analysis of the Mixers

4.1 Introduction

There are several factors that should be taken care of for the performance of mixers, including RF to IF conversion loss, standing wave ratio (SWR) at LO and RF inputs, LO/RF isolation, spurious rejection and harmonic suppression [15].

Conversion loss is the ratio of the IF output power to the signal input power. Some of the input signal power is converted in the mixer to the sum frequency, some of the input power remains at its original frequency, and part of the input signal power is converted to the IF. If one third of the input signal power is converted to each of these output frequencies, the IF output power is only one third of the input signal power, so the conversion loss of the mixer is 5 dB.

However, in this project, there are no suitable diodes for the Ku band mixer in the ADS data bank to do the integrated simulation. Therefore, only individual components are simulated. HFSS is used for the Ka band mixer simulation. In the structure manner, individual finline structure analysis is done.

If the LO and RF ports are mismatched for the inputs, reflection process will appear at both ports which cause reflections. In Chapter 2 and 3, the small insertion losses of these ports prove that the reflections of them are actually very little.

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The LO-to-RF isolation specifies how much of the LO signal will leak out through the RF port. The LO signal leaking into the RF port can interfere the RF signal or be transmitted out of the system antenna and can interfere with other systems.

The suppression of the single-tone and two-tone intermodulation products represent the spurious rejection. This parameter is also called intermodulation distortion (IMD). For a mixer inside a downconverter, a low pass filter is used to extract the desired IF components while rejecting the spurious. Similarly, a band pass filter is used if a mixer is used as an upconverter.

Harmonic suppression refers to the suppression of the local oscillator and signal harmonics. Low pass filter is used for harmonics suppression for both the Ku and Ka band mixers.

Unfortunately, no mixer meets all of the above specifications in the optimal way. Usually, only some factors, which are more important for the designed system, attract special attention while the remaining factors are sacrificed. Various structures are used to realize the special performance requirements. In this project, both the Ku and Ka band mixers require high LO/RF isolation and high RF leakage rejection and low conversion loss. The high RF and LO leakage rejection is achieved by an RF short followed by a LPF for the Ku band mixer. For the Ka band mixer, a special structure, called “swallow tail”, is used to adjust the flatness of RF/LO and LO/RF isolations across the whole Ka band. Further discussion on the structure and performance of the mixers will be shown in the rest of this chapter.

4.2 Structure analysis of the Ku and Ka band mixers

4.2.1 Structure analysis of the Ku band mixer

This section focuses on the structure analysis of the Ku band mixer. In order to allow broadband RF and LO input signals as required, Lange coupler has been used for both RF and LO input ports. Its high port-to-port isolation also ensures good RF/LO and LO/RF isolation. Two open radial stubs, or RF short, together with a LPF are designed for high RF and LO leakage rejection and harmonics suppression. Two Schottky barrier diodes are put between the Lange coupler and the RF short in opposite polarities to form a single balanced mixer. Schottky barrier diodes are selected due to their low impedance at high frequency and low transit time. Single balanced mixer structure is used since it is able to suppress LO noise and reject spurious responses comparing to single-ended mixer and requires less driving power and has lower conversion loss comparing to double balanced mixer. The main contribution of the Ku band mixer is its high RF leakage rejection. Normally, the RF leakage rejection of low pass filter is enough for a single balanced mixer. However, for a transceiver of a satellite ground station, antenna leakage is large and LO power leakage is also significant due to the common reference crystal extra RF short is needed for further RF rejection, especially for wideband operation.

The detailed circuit structure of the Ku band mixer is shown in Figure 4.1. As shown, RF and LO are both fed to the mixer through two ports of the Lange coupler, which is capable of broadband inputs and high port-to-port isolation. Two open radial stubs connect to 50 ohm line after the diodes for broadband RF leakage rejection. A LPF follows the RF short for further reduction of RF leakage and harmonics.

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Figure 4.1 The Ku band mixer structure

The structure can be simplified to the following block diagram in Figure 4.2. 90° phase shift is resulted for either RF or LO signal after the Lange coupler. Then combined signal, either shifted RF plus original LO or shifted LO plus original RF, is fed to Schottky barrier diodes.

Figure 4.2 Block diagram of the Ku band single balanced mixer

The principle of the Ku band mixer operation can be expressed using the following equations. The current and voltage directions are indicated by the arrows in Figure 4.2.

For inputs

$$RF = V_{RF}\cos(\omega_{RF}t)$$

$$LO = V_{LO}\cos(\omega_{LO}t)$$

After the Lange coupler, the voltages across two diodes are

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$$V1 = V_{RF}\cos(\omega_{RF}t) + V_{LO}\cos(\omega_{LO}t-\pi/2) \quad (4.1)$$

$$V2 = -V_{LO}\cos(\omega_{LO}t) - V_{RF}\cos(\omega_{RF}t-\pi/2) \quad (4.2)$$

The currents through two diodes are

$$I1 = aV1 + b(V1)^2 + c(V1)^3 + d(V1)^4 + \dots \quad (4.3)$$

$$I2 = aV2 + b(V2)^2 + c(V2)^3 + d(V2)^4 + \dots \quad (4.4)$$

$$I1 = a(V_{RF}\cos(\omega_{RF}t) + V_{LO}\sin(\omega_{LO}t)) + b(V_{RF}\cos(\omega_{RF}t) + V_{LO}\sin(\omega_{LO}t))^2 + \dots \quad (4.5)$$

$$I2 = -a(V_{RF}\cos(\omega_{LO}t) + V_{LO}\sin(\omega_{RF}t)) + b(V_{RF}\cos(\omega_{LO}t) + V_{LO}\sin(\omega_{RF}t))^2 + \dots \quad (4.6)$$

Normally, the 2nd order terms in equations (4.5) and (4.6) are considered since they provide the desired intermediate component, $\omega_{IF} = (\omega_{RF} - \omega_{LO})$ or $(\omega_{RF} + \omega_{LO})$.

4.2.2 Structure analysis of the Ka band mixer

The circuit pattern of finline-microstrip mixer in Figure 4.3, including a unilateral finline taper, an antipodal finline taper, a microstrip coupler, a LPF, swallow tail structure and two Schottky barrier diodes. The finline tapers are able to provide broadband RF and LO inputs. RF is fed to Schottky diodes with opposite polarities through finline. The two diodes are connected in series across the broadwall of the waveguide as seen from this finline RF input port. The LO pumped power is fed to Schottky diodes through antipodal finline and a coupled line DC-block (it also serves as IF-block). The diodes are thus in parallel with respect to the LO signal and the IF signal is extracted via a low pass filter. Microstrip coupler is used to avoid IF short and also can be used for external bias of diodes in case of self-biased down. A LPF is to take out the desired IF component while attenuating unwanted ones. Swallow tail structure is for flatness adjustment ability on RF/LO and LO/RF isolations.

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The unilateral finline taper for RF input is designed to a suitable slot width, with its characteristic impedance matching to the serial impedance of two diodes. It is beneficial to the size reduction when comparing with conventional reduced-height waveguide transformer or multi-section matching circuit. The LO input antipodal finline taper is also designed to a slot width with its characteristic impedance matching to parallel impedance of two diodes. Therefore, the impedance level of unilateral finline taper, diode and antipodal finline taper must be in ratio 4:2:1.

Figure 4.3 The Ka band swallow tail finline/MIC mixer structure

If the back metallization of the substrate is ended at the end point of front microstrip, the diodes will be poorly grounded through fins of finline due to lack of ground metallization (back metallization), therefore conventionally, backside sliding short is required for compensation.

By extending the back metallization to the edge of the ended finlines, and adding two metallized triangles that forms a swallow-tailed pattern, the grounding effect of diodes

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can be greatly improved and the sliding short is not necessarily required. This makes the matching between the microstrip and parallel diodes simpler and easier.

The Ka band mixer is a single balanced mixer too. However, it is 180° phase shift, unlike the 90° phase shift of the Ku band. The 180° hybrid section from the unilateral

Figure 4.4 E-field propagation direction (a) E-plane magic-T and (b) 180° hybrid

finline taper to microstrip is the critical part to form a single balanced mixer. The explanation for 180° phase shift can go back to magic-T description in [30]. The E-field of E-plane Magic-T is shown in Figure 4.4(a) when input signal is in port 3. The two output signals from port 1 and 2 have the same magnitude but with 180° phase shift due to the E field propagation direction as indicated in Figure 4.4(a). Similarly, the 180° hybrid can also be explained using E-field propagation direction in Figure 4.4(b). If two diodes are put inside this hybrid with opposite polarities, the RF signals through them will be in phase. However, the LO signals, which are fed to the diodes through microstrip from the right side, will be 180° out of phase due to the E-field propagation direction. Therefore, they form a 180° phase shift single balanced mixer. The simple block diagram is shown in Figure 4.5.

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Figure 4.5 Block diagram of the Ka band single balanced mixer

The operating principle of Ka band mixer can be explained using the following equations. The current and voltage directions are both indicated by the arrows in Figure 4.5. The voltages across two diodes at the 180° hybrid are

$$V_3 = V_{RF}\cos(\omega_{RF}t) - V_{LO}\cos(\omega_{LO}t) \quad (4.7)$$

$$V_4 = V_{RF}\cos(\omega_{RF}t) + V_{LO}\cos(\omega_{LO}t) \quad (4.8)$$

The currents through two diodes are

$$I_3 = aV_3 + b(V_3)^2 + c(V_3)^3 + d(V_3)^4 + \dots \quad (4.9)$$

$$I_4 = aV_4 + b(V_4)^2 + c(V_4)^3 + d(V_4)^4 + \dots \quad (4.10)$$

$$I_3 = a(V_{RF}\cos(\omega_{RF}t) - V_{LO}\cos(\omega_{LO}t)) + b(V_{RF}\cos(\omega_{RF}t) - V_{LO}\cos(\omega_{LO}t))^2 + \dots \quad (4.11)$$

$$I_4 = a(V_{RF}\cos(\omega_{RF}t) + V_{LO}\cos(\omega_{LO}t)) + b(V_{RF}\cos(\omega_{RF}t) + V_{LO}\cos(\omega_{LO}t))^2 + \dots \quad (4.12)$$

The 2nd order terms in equations (4.11) and (4.12) are considered since they provide the desired intermediate component, $\omega_{IF} = (\omega_{RF} - \omega_{LO})$ or $(\omega_{RF} + \omega_{LO})$.

Chapter 4 Structure and Performance Analysis of the Mixers

4.3 Performance analysis of the Ku and Ka band mixers

The RF/LO and LO/RF isolation, RF and LO leakage rejection and conversion loss are the parameters needed to pay special attention to for both the Ku and Ka band mixers. In this section, their performance analysis will be discussed.

4.3.1 RF/LO and LO/RF isolation

The RF/LO and LO/RF isolation of the Ku band mixer are realized by the good port-to-port isolation of the Lange coupler. Therefore, by simulating the port-to-port isolation of Lange couplers in ADS, the RF/LO and LO/RF isolation can be achieved. As Figure 4.6 shows, port 1 and 3 are used for LO and RF port respectively. Therefore the simulation results of S13 and S31 of the Lange coupler give the LO/RF and RF/LO isolation.

Figure 4.6 RF/LO and LO/RF isolation of the Ku band mixer

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Obviously, the Lange coupler's port-to-port isolation, S_{31} and S_{13} , were the same and greater than 20dB over the required frequency range, which ensured good RF/LO and LO/RF isolations.

The RF/LO and LO/RF isolation of the Ka band mixer is due to the field orthogonality of RF and LO signals. In addition, the gap, where two diodes are put, can further improve the isolation. Also the special structure, swallow tail, can be used to adjust the flatness of RF/LO and LO/RF isolation. This will be further discussed in Section 4.3.3.

4.3.2 Broadband high RF leakage rejection of the Ku band mixer

Broadband high RF and LO leakage rejection is the main advantage of the Ku band mixer, which is realized by a RF short and a low pass filter. RF short initially comes

Figure 4.7 Open stubs (a) rectangular and (b) radial

from quarter wavelength stub, as in Figure 4.7(a). However, this is only suitable for narrow band operation. The wideband RF rejection is realized by the open radial stub shown in Figure 4.7(b). The two stubs are both connected to 50 ohm line. For the open radial stub, the distance from its open edge to the connecting edge is various, unlike the one in Figure 4.7(a), the distance from its open edge to the connecting edge

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is the constant of a quarter wavelength at the center frequency. Therefore, the open radial stub is capable of attenuating more broadband signal. The design of this RF short is based on quarter wavelength over the whole Ku band and then uses the trial and error method. Numerical methods may be used for further optimization [31].

Figure 4.8 presents the two different open stubs, rectangular and radial. Both of them are connected to a 50 ohm line. Advance Design System (ADS) is used for simulating the performance of both types of stubs. The superiority of open radial stub comparing to rectangular one is shown in Figure 4.9. The lower curve represented the radial stub and the upper one was for the rectangular stub. Obviously, the open radial stub has a larger bandwidth than the rectangular one. Over the specified frequency range, the leakage rejection of the open radial stubs was no less than 10 dB. It's mainly due to the various distances from its open edge to the connecting edge. This feature resulted in its potential wideband operation. The simulation results confirmed it.

Figure 4.8 Simulated rectangular and radial open stubs

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Figure 4.9 Rejection comparison of the two open stubs

In order to achieve better than 30 dB RF leakage rejection over the required frequency range, RF short followed by a LPF structure was used with excellent harmonic suppression and leakage rejection as shown in Figure 4.10. The sharp transition around 18.35 GHz was due to the addition of the LPF and RF short.

Figure 4.10 Total leakage rejection of the open stubs and LPF

As confirmed by the simulation results, the present design exploits the broadband properties of the Lange coupler and radial stubs to achieve broadband operation. The structure with two open radial stubs followed by a LPF has achieved better than 30 dB RF leakage rejection over the required frequency range. The broadband and high

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leakage rejection features prove that the designed mixer is suitable for satellite transceivers, which have large antenna leakage and LO power leakage.

4.3.3 Broadband RF/LO isolation adjustment of the Ka band mixer

The main advantage of the Ka band mixer design is its adjustment capability for RF/LO and LO/RF isolation using the swallow tail structure. Finite element method (FEM) is used for performance analysis of the Ka band mixer in this section. It aims to study the RF/LO isolation properties of the swallow tail finline mixer we did previously [6], which has backside metal at the end of the unilateral finline taper.

Two different structures were simulated by using HFSS and the only difference between them was the swallow tail structure, which was the backside metal at the end of unilateral finline taper. Figure 4.11 redraws part of Figure 4.3 to emphasis the difference between two structures. One has swallow tail structure and the other one doesn't have.

Figure 4.11 Swallow tail structure

The results in Figure 4.12 showed high RF/LO isolation over Ka band because of orthogonality of RF and LO fields. However, there were some sharp transitions near

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26GHz and 34GHz in Figure 4.11(a), which were not good for wideband operation. However, for the structure with swallow tail, the isolation was around 28 to 38dB over the whole Ka band as shown in Figure 4.11(b), which ensured wideband operation stability of the finline-microstrip mixer. Together with the experiment result in [1], the simulation results confirmed the adjustment capability of the swallow tail structure on RF/LO isolation.

The swallow-tailed structure on the backside of finline-microstrip joint of the finline-microstrip mixer could reduce the hardware size, structure complexity, and mostly provide adjustment capability on RF/LO isolation flatness across the specified frequency band.

Figure 4.12 RF/LO isolation (a) without swallow tail (b) with swallow tail

Chapter 5 Scaled S Band Mixer Measurement for Design Evaluation

5.1 S band mixer design

Due to time limitation, both the Ku and Ka band mixers are still in fabrication. Therefore the ready S band mixer was measured up to now. The block diagram of S band mixer is similar to that of the Ku band downconverter as shown in Figure 4.2, but without the RF short circuit. The Lange coupler and low pass filter designs for the S band mixer are the same as those of the Ku band downconverter, except that the substrate height is 30mil and its dielectric constant is 2.45. Therefore based on the similar procedures, the design values of the Lange coupler are as follow.

$$W/h = 0.5$$

$$S/h = 0.052$$

$$W = 0.5 \times 30 = 15 \text{ mil}$$

$$S = 0.052 \times 30 = 1.56 \text{ mil}$$

The even- and odd-mode velocities are

$$V_e = 2.16 \times 10^8 \text{ m/s}$$

$$V_o = 2.32 \times 10^8 \text{ m/s}$$

The even- and odd-mode effective relative dielectric constants are

$$\begin{aligned} \epsilon_{re} &= (c/V_e)^2 \\ &= [(3 \times 10^8)/(2.16 \times 10^8)]^2 \\ &= 1.929 \end{aligned}$$

$$\begin{aligned} \epsilon_{ro} &= (c/V_o)^2 \\ &= [(3 \times 10^8)/(2.32 \times 10^8)]^2 \\ &= 1.672 \end{aligned}$$

The effective relative dielectric constant is $\epsilon_{eff} = [\epsilon_{re}\epsilon_{ro}]^{0.5} = 1.796$.

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The quarter wavelength for S band and the required substrate is

$$c = 3 \times 10^8 \text{ m/s and } f = 2 \text{ GHz}$$

$$\lambda = c/f/(\epsilon_{\text{eff}})^{0.5} = 11.19 \text{ cm}$$

$$1 \text{ inch} = 1000 \text{ mil} = 2.54 \text{ cm}$$

Therefore $\lambda = 4406.61 \text{ mil}$ and $\lambda/4 = 1101.65 \text{ mil}$.

For the low pass filter, the design requirements are as follow.

Filter type: Chebyshev (Equi-ripple)

Ripple height, L_{ar} : 0.1 dB (set by integer)

Cutoff frequency (f_c): $f_c = 1 \text{ GHz}$

Minimum stopband, L_{as} : 15 dB at stopband frequency, f_s , 1.5 GHz

The filter order can be obtained using the following formula with the corresponding given values.

$$n \geq [1/\cosh^{-1}(\omega_s/\omega_c)] \cosh^{-1}\{(10^{L_{\text{as}}/10}-1)/(10^{L_{\text{ar}}/10}-1)\}^{1/2}$$

The value of n is determined as $n \geq 4.4509$.

The filter prototype element values are $g_0 = 1$, $g_1 = 1.1468$, $g_2 = 1.3712$, $g_3 = 1.9750$, $g_4 = 1.3712$, $g_5 = 1.1468$ and $g_0 = 1$ with $n = 5$, $g_0 = 1$, $w'_1 = 1$ and 0.1 dB ripple.

The corresponding capacitance and inductance can be got using the following formulae,

$$C_n = g_n/w_c Z_o$$

$$L_n = g_n Z_o/w_c$$

Therefore

$$C1 = 3.65 \text{ pF}, L1 = 10.912 \text{ nH}, C2 = 6.287 \text{ pF}, L2 = 10.912 \text{ nH}, C3 = 3.65 \text{ pF}$$

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The prototype for the designed LPF is shown in Figure 5.1.

Figure 5.1 The prototype of the low pass filter using LC components

5.2 Simulation and optimization of S band mixer’s components

Lange Coupler

Lange coupler is the most critical part to realize the single balanced mixer, which is supposed to separate the input power from port 1 and 3 equally but with 90° phase shift.

Figure 5.2 Simulation result of the Lange coupler based on calculated values at S band

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The simulation results of S21 and S41 of the designed Lange coupler are given in Figure 5.2. The coupling ratio, S21, and directivity, S41, were both close to desired 3 dB value. Optimization can further improve the circuit performance. It helped to remove the uncertainty and inaccuracy of the results after the simulation. The optimized values were updated until the desired performance was obtained.

In this simulation, the width, spacing and length of the Lange coupler were given a small range for adjustment during optimization so that the appropriate changes can be made in the design in order to meet the required specifications. To obtain optimum performance, goals were set for S21 and S41 in the frequency range of 1.6 to 2.2 GHz. Minor changes resulted for the design after optimization. The design dataset was updated until the desired performance was obtained.

The optimization setup was very similar to that of the Ku band mixer in Chapter 2. The optimized results are shown in Figure 5.3. The initial values of Lange coupler were $W = 15$ mil, $S = 1.56$ mil and $L = 1101.65$ mil. After optimization simulation, W became 14 mil, S was 1.5536 mil and L was 1135.98 mil.

Figure 5.3 (a) S21 and S41 of Lange coupler before and after optimization

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Low Pass Filter

The low pass filter using lump components was simulated according to the calculated values. Figure 5.4 shows the simulation setup and result. Obviously this LPF worked as expected. The cutoff frequency was around 1 GHz and the attenuation was about 20 dB at the stop frequency 1.5 GHz.

Figure 5.4 Low pass filter for S band

The low pass filter realized by microstrip line was simulated in ADS too. The simulation result shown in Figure 5.5 is slightly shifted to the left at the edge of the cutoff frequency comparing to the one given in Figure 5.4. The small difference is due to some approximation in the process of transforming lump components to microstrip lines, which makes the microstrip lines not provide the same admittance and reactance as the lump components at the edge of the cutoff frequency.

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Figure 5.5. Simulation result of LPF using microstrip

Integration Simulation

Conversion loss is an important performance measurement of a mixer. The integration simulation was done to measure it. The RF signal was set to -20 dBm and .85 GHz. The LO signal was set to 10 dBm and 1.8 GHz. The diode, HSMS286C, was selected from the ADS data bank. The IF component would be at 50 MHz based on operation principle of a mixer. By measuring the output power at 50 MHz from the IF port, the difference between the output IF power and the input RF power would be the conversion loss.

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Figure 5.6(a) Integration simulation of the S band mixer

However, the RF power going to the Lange coupler was not -20 dBm because of the loss in source. Therefore, the measured RF input power, V_{in} as indicated in Figure 5.6, was used as the RF input power instead. In this case, it was -30.653 dBm. The

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measured IF power was -37.693 dBm at 50 MHz. Then the conversion loss was around 7 dB. Other settings for RF or LO sources can be used for more simulations.

Except the conversion loss, the LO/RF isolation was simulated too. The result is given in Figure 5.6(b). It is greater than 30dB over the frequency range with good flatness.

Figure 5.6(b) S band mixer's LO/RF isolation

5.3 Scaled S band mixer evaluation results

The photograph of the fabricated S band mixer is shown below. The RF short structure in the middle was put there for photograph only. In the real measurement, the RF short structure was not there since the LPF can reject the RF and LO leakage in this S band mixer. The conversion loss, LO/RF isolation, RF and LO leakage rejection, SWR at RF port, spurious rejection and harmonic suppression have been measured.

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5.3.1 Conversion loss

An important specification for a mixer is conversion loss, which is the ratio of the IF output power to the signal input power. Some of the input signal power is converted to the sum frequency in the mixer, some of the input power remains at its original frequency, and part of the input signal power is converted to the IF. If one third of the input signal power is converted to each of these output frequencies, the IF output power is only one third of the input signal power, so the conversion loss of the mixer is 5 dB. As shown in the simulation result of the S band mixer, the conversion loss was around 7 dB. In mathematics, conversion loss = input power (dB) – output power (dB).

Figure 5.7(a) Measurement setup for conversion loss

The measurement setup is shown in Figure 5.7(a). Two Agilent signal generators were used to provide both RF and LO input signals. A spectrum analyzer was used to see the IF output signal at the desired intermediate frequency. The conversion loss can be obtained from the difference of RF input power and IF output power. In this measurement, there were five sets of parameters for RF and LO signals used to measure the conversion loss. The only variable was the LO power and its values were 0, 5, 10, 15 and 20 dBm separately. And its frequency range was 1.55 to 1.95 GHz.

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The RF power level was -20 dBm and frequency range was from 1.6 to 2 GHz. The IF product was at 50 MHz. The measurement results of the conversion loss are shown in Figure 5.7(b)

Figure 5.7(b) Measured conversion loss of the S band mixer

As shown in the above figure, the trend of conversion loss was to increase with frequency although some variations. That means the performance of this mixer was better at lower end of the frequency range. Other than frequency, LO power was another factor that affected the conversion loss. The better conversion loss was obtained when the LO power increased from 0 to 10 dBm. However, the trend was no longer true when the LO power reaches 15 and 20 dBm. As the previous simulation result showed, the conversion loss was around 7 dB when RF power was set at -20 dBm and 1.85 GHz and LO power was set at 10 dBm and 1.8 GHz. However the measured result was around 8.5 dB for the same settings, which was a bit higher than the simulated one. The reason may be that some ideal conditions are often assumed in the simulation, for example, zero metal loss.

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5.3.2 LO/RF isolations

The LO-to-RF isolation specifies how much of the LO signal will leak out through the RF port. The LO signal leaking into the RF port can interfere with the RF signal or be transmitted out of the system antenna and can interfere with other systems. The measurement setup for isolation is shown in Figure 5.8.

Figure 5.8 Measurement setup of RF/LO isolation

The HP network analyzer was used for the measurement. The IF port of the mixer was terminated with a 50 ohm load. RF and LO ports were connected to two ports of the network analyzer. By measuring the S parameters of these two ports, the isolation can be determined. Before the S-parameter measurement, a full 2-port calibration should be done for the network analyzer to obtain accurate measurement result. The LO was set to 10 dBm from 1.55 to 1.95 GHz for the LO/RF isolation measurement. The measurement result is shown in Figure 5.9.

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Figure 5.9 Measured LO/Rf isolation of the S band mixer

Over the frequency range of interest, the LO/Rf isolation was greater than 25 dB. It means there was less than 0.3% LO signal leaked to RF port from 1.55 GHz to 1.95 GHz. Comparing to the simulation result in Figure 5.6(b), its flatness was not as good as the simulated one. At the lower end of the frequency, the isolation was worse than the simulated one while it's better at the higher end. This was mainly due to the fabrication tolerance on the width and the gap of the Lange coupler.

5.3.3 RF and LO leakage rejection

RF and LO leakage rejection measures how much the RF and LO power will leak to the IF port. If the mixer is used as a downconverter, the IF signal will be very small since it is the weak received signal minus the conversion loss. Any high RF and LO signal leakage will interfere with the signal and make it hard to detect. Therefore high RF and LO leakage rejection is required. Normally a LPF is used for leakage rejection and sometimes, additional RF short is necessary. In this measurement, no RF short was used.

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The measurement setup for RF and LO leakage rejection in Figure 5.10 was the same as the one for conversion loss except that the leaked RF and LO signals were measured instead of IF signal in the spectral analyzer. The RF was set to -20 dBm from 1.6 to 2.0 GHz. The LO power was set to 0, 5 and 10 dBm and its frequency range was from 1.55 to 1.95GHz.

Figure 5.10 Measurement setup of RF leakage rejection

For this S band mixer, the LPF was designed to reject the RF and LO leakage. Its performance was measured using Network Analyzer HP8753E before connecting to the IF port of the mixer. The measurement result is given in Figure 5.11. The 3dB cutoff frequency was around 1GHz. The attenuation was around 24 to 40 dB from 1.5 to 2 GHz, which included both the LO and RF frequency ranges. Beyond 2 GHz, the attenuation became smaller. It was due to the nature of the microstrip line filters, which is periodic. For example, there is a microstrip line with length, λ_1 , corresponding to frequency, f_1 . The signal at frequency, $f_2=2 \times f_1$, can go through this microstrip line with the same characteristics as the signal at f_1 since λ_1 is $2 \times \lambda_2$. Comparing to the simulation result in Figure 5.5, the cutoff point was slightly shifted to the left. One reason was due to neglect of the metal loss in the simulation. Another was due to the fabrication. The etching process affected the width of the

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microstrip line and the real dielectric constant and height of the substrate may not be the same as the simulated ones.

Figure 5.11 Measurement result of the fabricated LPF

The leakage rejection of the mixer with and without the LPF was measured for LO power at 0, 5 and 10 dBm. The result is shown in Figure 5.12(a). The upper portion represented the LO leakage rejection with the LPF and the lower portion represented the LO leakage without the LPF. The exact improvement of the LO leakage rejection achieved by the LPF is given in Figure 5.12(b), which was around 25 to 34 dB from 1.55 to 1.95 GHz. Comparing to the LPF's response from frequency 1.5 to 2 GHz in Figure 5.11, the improvement was agreed with it. No matter with or without the LPF, the higher LO power achieved better LO leakage rejection.

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Figure 5.12(a) LO leakage rejection with/without the LPF

Figure 5.12(b) LO leakage rejection difference with and without the LPF

Figure 5.13 shows the measurement result of the RF leakage rejection with and without the LPF for LO power at 0, 5, 10 dBm. The improvement of the leakage rejection by the LPF was from 26 to 40 dB from 1.6 to 2.0 GHz, which was similar to the result of the LO leakage rejection. However, the LO power doesn't have much effect on the RF leakage rejection.

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Figure 5.13(a) RF leakage rejection with/without the LPF

Figure 5.13(b) RF leakage rejection improvement with the LPF

5.3.4 Spurious rejection

Spurious rejection is a very important performance factor of a mixer, especially for upconverters. The measurement setup for spurious rejection is given in Figure 5.14, which is exactly the same as the one for conversion loss. However, this measurement

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focuses on the spurious rejection, which is the difference between the desired IF signal and the unwanted spurious. The RF signal was set to -20 dBm from 1.6 to 2 GHz. The LO signal was set to 0, 5 and 10 dBm to measure the variation of spurious rejection capability with different LO power. The LO signal was from 1.55 to 1.95GHz. The spurious, which are close to the desired IF signal, should be paid attention to since they cannot be filtered out by the LPF. For the measurement setup here, the intermodulation products at $2(RF-LO)$, $3(RF-LO)$ and so on, were very close to the desired IF signal at 50MHz. The spurious at 100 MHz, 150 MHz and so on, were possible to appear. Within 1 GHz span, the spurious at $2(RF-LO)$, 100MHz, appeared and was measured. The measurement results with and without the LPF are both given in Figure 5.15.

Figure 5.14 Spurious rejection measurement setup

Without the LPF, the spurious rejection was between 35 and 50 dB and higher LO power achieved better rejection. With the LPF, the spurious rejection was improved more than 25 dB as shown in the upper portion of Figure 5.15. However, the LO power no longer has much effect on the rejection.

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Figure 5.15 Spurious rejection with/without LPF under various LO power

5.3.5 Harmonic suppression

This parameter is a measure of the capability of a mixer on suppression of input signal harmonics. The measurement was the same as the one for conversion loss except that the harmonics, 2LO and 2RF, were measured on the spectral analyzer since they are the closest harmonics to the desired IF signal. The RF signal was set to -20 dBm from 1.6 to 2 GHz. And the LO signal was set to three different values, 0, 5 and 10 dBm, for the same frequency range 1.55 to 1.95 GHz to measure the variation of harmonic suppression with LO power. . During the measurement, only the 2nd harmonic of the LO signal (2LO) appeared on the spectral analyzer. The 2nd harmonic of the RF signal had the same power level as the noise floor. The measurement result of 2LO is shown in Figure 5.16. The x axis is the frequency of the 2LO component and the y axis is the output power of the 2LO component. The upper portion is the result without the LPF. The lower portion represents the result with the LPF. Obviously, the 2nd harmonic of the LO signal had higher power before the LPF was connected. After the LPF was connected, the 2nd harmonic of LO disappeared for LO power at 0 dBm and only appeared at the higher end frequency

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for for LO power at 5 and 10 dBm. As the upper portion of Figure 5.16 shown, the higher LO power resulted in larger 2nd harmonic of LO.

Figure 5.16 2nd harmonic of LO signal

5.3.6 SWR at RF input

The measurement setup is easy since it is basically the reflection of RF port. For RF port’s reflection, LO port was connected to the signal generator and IF port was terminated with a 50 ohm load. The setup for RF port’s reflection is shown in Figure 5.17. The SWR of RF port can be obtained by measured S parameters. Before doing that, the 1-port calibration is necessary for accurate measurement results.

Figure 5.17 Reflection measurement setup of RF port

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Figure 5.18 S11 of RF port

The reflection coefficient at RF port was acceptable at the lower end of the frequency range since it is more than 9 dB. However, the mixer has bad SWR at the higher end frequency.

5.4 RF short

The RF short structure was not used in the S band mixer design since the LPF can reject the RF and LO leakage. However, it has been used in the Ku band mixer design. In this chapter, a RF short has been designed, fabricated and measured to show its performance. The simulation result and measurement result in Figure 5.19 and 5.20 showed the high attenuation feature of this structure in the designed frequency range, which was a proof of the feasibility for the RF short using in the Ku band mixer. The small difference between the simulation result and the measurement result is mainly due to fabrication, for example, etching.

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Figure 5.19 Simulation result of the RF short

Figure 5.20 Measurement result of the RF short

Chapter 6 Conclusion and Future Work

6.1 Conclusion

Throughout this project, the objectives specified in Chapter 1 were met. The operation principle of single balanced mixers has been understood. The 3 dB Lange coupler for broadband RF and LO inputs of the Ku band mixer has been designed. The unilateral finline taper for impedance transformation from the waveguide impedance of $480\ \Omega$ to beam lead diode impedance of $180\ \Omega$ and the antipodal finline taper for impedance transformation from the waveguide impedance $480\ \Omega$ to $50\ \Omega$ microstrip line have been designed. Microstrip LPFs with cutoff at 7.5 GHz and 10 GHz for the Ku and Ka band mixers respectively have been designed too. In addition, special structures, such as RF short of the Ku band mixer and swallow tail of the Ka band mixer, have been theoretically studied.

For the Ku band mixers, a 3 dB Lange coupler is used for RF and LO inputs due to its wideband operation and high port-to-port isolation. The Lange coupler design is based on Presser's work and optimized in ADS. Beam lead Schottky diodes follow the designed Lange coupler to achieve mixing operation. A low pass filter with cutoff frequency at 7.5 GHz is used to take out the desired IF component. In order to satisfy high RF leakage rejection requirement for some special transceiver sets at satellite ground station, whose RF leakage may be originated from antenna coupling and common shared synthesizer crystal, an additional RF short is used before the LPF. Its performance has been discussed in Chapter 4.

For the Ka band mixer, it consists of an exponential unilateral finline taper, an exponential antipodal finline taper, a microstrip coupler and a microstrip low pass filter. The exponential finline taper is designed to transform the waveguide

Chapter 6 Conclusion and Future Work

impedance of 480Ω to the beam lead diode impedance of 180Ω . The exponential antipodal finline taper is designed to transform the waveguide impedance of 480Ω to the microstrip coupler of impedance 50Ω . The microstrip coupler is designed to pass signal from 26.55 to 31.2GHz with small insertion loss. And the microstrip loss pass filter is designed to cut off at 10GHz.

The exponential unilateral finline taper is designed at minimum reflection coefficient at a length of 17.142mm with a slot width of 3.556mm, i.e., width of WR28 waveguide with 480Ω impedance, and narrowing down to 0.3mm, i.e., 180Ω for matching beam lead diode impedance. The exponential antipodal finline taper is designed at minimum reflection coefficient at a length of 9.0909mm with a slot width of 3.556mm, i.e., width of WR28 waveguide with 480Ω impedance, and narrowing down to 0.8mm, i.e., 50Ω for matching the microstrip coupler impedance. The transition design of the tapers and their performance are shown in Chapter 4.

Software ADS is used for simulation and optimization of microstrip components using in the mixers. HFSS is used for finline and finline-microstrip combined components simulation. Since no suitable diodes in ADS and no diodes in HFSS at all, there are no integrated simulations for both Ku and Ka band mixers. Due to the time limitation, both Ku and Ka band mixers are still in fabrication now. The scaled S band mixer measurement is feasible evaluation of the designed Ku band mixer. The conversion loss measurement result over the frequency range 1.6 to 2.0 GHz proves the broadband property of the Lange coupler. The broadband high LO/RF isolation of the Lange coupler has been verified by the measurement result too. The measurement results of the LO and RF leakage rejection, spurious rejection and harmonic suppression, have shown the improvement of the LPF on these parameters. The RF short structure, which has greater than 10 dB attenuation from 1.5 to 3.0

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GHz, shows the broadband RF short in Ku band is feasible. The similar structure of the Ku band mixer was given in [33]. The feasibility of the Ka band mixer design, however, had been shown in previous work [6] and it is quite agreeable.

6.2 Future work

During the whole project, the following areas can be further studied.

- Bonding wires of Lange coupler.

For high frequency application, the width and spacing of Lange couple are quite small normally. The performance will be affected much due to fabrication error. Bonding wires can be used to adjust the performance. Therefore, equivalent circuit of bonding wires can be further studied to find out the rule of their adjustment ability.

- Momentum optimization of Lange coupler's performance

In this project, all optimization has been done using SCHEMATICS. ADS can also do optimization using LAYOUTS, which is supposed to achieve performance closer to real products.

- Broad band RF short design method

In this project, the RF short design is based on try and error method and its attenuation is more than 10dB. Further research can be done to find a better way for higher attenuation.

- Further study on swallow tail structure

Chapter 6 Conclusion and Future Work

According to previous research work, the swallow tail structure will affect the isolation/rejection performance. Further analysis is suggested on its equivalent circuit modeling.

- Upconverter design

The designed Ka band mixer will be pumped in the LO and IF powers to observe the RF output as an upconverter. Alternatively, RF and IF ports can be internally exchanged, or, keep the original design but put the diodes in the same polarity instead. These works can be done in the future. In order to achieve broadband operation, tunable band pass filter may be suggested.

Published Papers

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2. J. Y. Zhou, W. X. Liu, Y. L. Lu and J. Fu, "Study of Power Allocation in Multiple-Input-Multiple-Output (MIMO) Systems," Accepted for publication in *IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, Barcelona, Spain, Sep 2004.
3. J. Fu, J. Y. Zhou, Y. L. Lu, J. G. Ma and Z. X. Shen, "Finline Technology for Millimeter Wave MEMS," Accepted by *International Symposium on Microtechnologies*, Spain, May 2003.
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Appendix A — Drawings of the Ku Band Mixer

(I) Substrate

1. Substrate: Gold plated Alumina. DK = 9.6 and thickness = 25 mil
2. Unit of the diagram is in **mm**
3. Cutting to the designed size and etching tolerance ± 0.1 mil

Enlarge portion unit is in mil

A Lange coupler
unit: mil

B

C

(II) Chip carrier

- 1. Silver coated Aluminum Alloy
- 2. Unit is in mm

(III) Housing

- 1. Silver coated Aluminum Alloy
- 2. Flatness is ± 0.01 mm
- 3. Tolerance is ± 0.05 mm
- 4. Unit is in mm

Housing-body
unit: mm

(IV) Draft 3D drawings of Substrate and Housing

Housing

Chip Carrier

Substrate

Appendix B — Drawings of the Ka Band Mixer

(I) Circuit Pattern

- 1. Material: **gold plated Duroid** soft substrate (RT/Duroid 5880, DK=2.22 and Height = 10 mil)
- 2. In Figure 11, the upper one is the front side of the substrate and the lower one is the backside.
- 3. The shaded area is metal.

SECTION A in circuit pattern is shown below.

COUPLER (unit: mm)

DIODE (MA4E2037)

The two diodes must be put in **opposite polarity**.

(II) Housing

The housing is composed of **two** parts and screw them together

1. Material is gold plated brass.
2. Flatness is ± 0.01 mm
3. Tolerance is ± 0.05 mm

(III) Housing – Upper part

(IV) Housing – Lower part

Appendix C — Vendor List

The vendor list for various materials, components and fabrication are given in the following table.

	Individual suppliers	Contact information	E-mail address	Model number
Substrate	Rogers Corporation Microwave Material Division	Charles HP: 97945370	charles@meds-tech.com	Ultralam 2000 DK 24 - 2.6 may be used for the second design.
Diode	M/A-COM			MA4E2037
Housing	Distinct Engineering	Mr. Ken Soh HP: 96284546	distinct@singnet.com.sg	
SMA connector	MuTECH Pte Ltd	Dennis Kum HP: 97320973	denkws@pacific.net.sg	Coaxial Components Corp. 3244-6-*
GPO connectors (optional)	Amdus	Chan Bin HP: 96890623		
Gold etching factory	Satellite Lab in S2.2	A/Prof. Chua Tai Wei Tel: 67906859	etwchua@ntu.edu.sg	
System OEM	MEDs Technologies Pte Ltd	Charles HP: 97945370	charles@meds-tech.com	
PCB	Circuit Image Manufacturing Pte Ltd	Royston Tan (Manager) Susannah LEE (engineer) Tel: 65454338 ext. 103		General PCB
Instrumentation Grade Adapter	ANRITSU	Chen Xuesong Tel: 65687430		35WR28KF
Screws	Farnell	Tel: 63803105 Fax: 67880300	sg-accounts@farnell- networkinone.com	