DESIGN, ANALYSIS AND IMPLEMENTATION OF ACTIVE MIXER FOR 5 GHz WIRELESS LAN APPLICATIONS

A DISSERTATION

Submitted in partial fulfillment of the requirements for the award of the degree

of

MASTER OF TECHNOLOGY

in

ELECTRONICS AND COMMUNICATION ENGINEERING

(With Specialization in Communication Systems)

By

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CANDIDATE'S DECLARATION

I hereby declare that the work, which is presented in this dissertation report, entitled **"DESIGN, ANALYSIS AND IMPLEMENTATION OF ACTIVE MIXER FOR 5 GHz WIRELESS LAN APPLICATIONS"**, being submitted in partial fulfillment of the requirements for the award of the degree of MASTER OF TECHNOLOGY with specialization in COMMUNICATION SYSTEMS, in the Department of Electronics and Computer Engineering, Indian Institute of Technology, Roorkee is an authentic record of my own work carried out from June 2007 to June 2008, under guidance and supervision of **Dr.N. P. Pathak**, Assistant professor, Department of Electronics and Computer Engineering, Indian Institute of Technology, Roorkee.

The results embodied in this dissertation have not submitted for the award of any other Degree or Diploma.

Date 27th June 2008
Place: Roorkee

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CERTIFICATE

This is to certify that the statement made by the candidate is correct to the best of my knowledge and belief.

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Place: Roorkee

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ABSTRACT

The increased need for data transmission within home and office environment has created the remarkable proliferation of the wireless standards. Some, such as Bluetooth and IEEE 802.11b allocated in the ISM 2.4 GHz band, are a necessary and integral part of every day living. However, this very fact has resulted in an ever increasing number of users and applications. There is an augmenting demand being made on higher data rates which have culminated in the migration of the most powerful communication standards to higher frequency bands and the integration of multiple standards in the modern mobile devices. Wi-Fi in its IEEE 802.11a version WiMax IEEE 802.16a/e, are the leading standards of the new era of reliable broadband-connected wireless networks and both are allocated in the unlicensed 5 GHz U-NII band. The low cost and low power consumption must be key points of reference within the design of receivers and transceivers for these standards.

The development of the FET's mixers, offers FET as the lowest cost and the most effective time-to-market technology. The design of RF mixers allows the challenge of the analog integration aim of the modern receivers and transceivers. In addition, constant advances in the design tools of this process provide better accuracy for the high frequency RF design.

The main objective of this work is the design and implementation of mixer for 5 GHZ WLAN applications. The mixer circuit designed and characterized are the Power Combiner, the RF Matching Section and the IF section. The three blocks composition is explained, analyzed through simulation, final mixer is implemented and then tested.

1. INTRODUCTION

1.1 Mixer Review:

Credit for the invention of the mixer is generally given to Major Edwin Armstrong, one of the most creative inventors in the history of radio communication. Crude mixing process had been employed in the past, mostly in attempts to down-convert received signals directly to baseband. These techniques generally worked poorly, because the stability of the local oscillator was not good enough. Armstrong was the first to use Vaccum tube mixers to shift a received signal to an "intermediate frequency," where it could be amplified with good selectivity and low noise, and finally demodulated. Armstrong is also credited with the invention of frequency modulation, which would be impossible without such receivers, because good quality tunable FM demodulators were nearly impossible to produce. His "superheterodyne" receiver is still the model for communication receivers in use today for applications ranging from cheaper transistor AM radios to the most sophisticated satellite systems.

The subsequent history of mixers [1] is, in fact, the history of low-noise receivers. The major emphasis on high quality microwave mixers arose with the development of radar in World War II, at the MIT Radiation Laboratory. Receiver sensitivity in a radar system is critical in establishing its maximum range capability. Early radars had to operate at least at UHF, preferably microwave frequencies [2], in order to be able to view targets the size of an airplane. At these frequencies, good low-noise amplifiers did not exist, so it was necessary to depend on the mixer as the first stage of the receiver, the one which most strongly affects sensitivity. Very little theoretical work on mixers had been performed until 1940, and the quality of point-contact "crystal" diodes [3] was poor. In less than 10 years, the theoretical underpinnings of mixer design were in place and the quality of diodes – particularly diode packages – had been greatly improved. Conversion losses achievable in the low microwave range dropped from 20 dB in 1940 to 10 dB in 1945. By the early 1950s, mixers with conversion losses around 6 dB were being produced regularly. Today, the theory of diode mixers [4] is well established, and image-enhanced mixers are regularly produced with conversion losses below 4 dB at frequencies around 50 GHz. Though passive mixers [5] (diode mixers) are attractive in terms of linearity and noise performance, active mixers [6] are also preferred mostly in receiver applications where conversion gain is required.

Although single device mixers are entirely practical, and are often preferred for millimeter–wave [7] receiver applications, they have some undeniable faults. The most obvious are the difficulty in applying the LO to the diode without LO leakage into the RF or IF ports, and the need for a filter diplexer or other device to allow LO injection. Balanced mixers [8] overcome these problems and have additional advantages as well. Image reject receiver [9] is one of the types of balanced mixer to eliminate image response since images of the wanted frequency are a problem which interfere with the wanted signal.

Development of high power and inexpensive LO sources are the major challenge faced by researchers. The power available from solid-state sources drops off with the inverse square of frequency due to electronic limitations in the material and hence at the higher frequencies, higher LO powers come at much higher cost. Therefore, one of the main goals of terahertz mixer design has been the reduction of LO power requirements, with emphasis towards receiver configurations that permit harmonic mixing. Therefore, Subharmonic mixers [10] are valuable at these frequencies. The above two mixer design combined to a single design called subharmonic image-reject mixer also performs well in millimeter range.

Devices for Mixers:

Although any nonlinear device can in theory be used as a mixer, only a few devices satisfy the practical requirements of mixer operation. Any device used as a mixer must have a strong nonlinearity, repeatable electrical properties between device and fabrication lots, low noise, and adequate frequency response. The nonlinear device most often employed for mixing is the Schottky-barrier diode, in the past, PN-junction diodes have been used for mixers, although point-contact diodes have been used much more extensively. Schottky-barrier diodes are superior to PN-junction devices for mixer applications because they are majority carrier devices. The successful development of small, high quality schottky junctions has allowed the development of mixers at frequencies above 1000 GHz. In mundane applications, Schottky diodes now used in virtually all mixers, and are available in a wide variety of packages.

FET and bipolar transistor mixers are common at frequencies below 1 GHz, but are uncommon in microwave applications. BJT mixers are occasionally found at low frequencies. One major advantage of the FET mixer over the diode mixer is that it can achieve conversion gain. Since modern FETs are excellent low-noise devices, FET mixers also have good noise figures. FET mixers are practically well suited for use in GaAs monolithic integrated circuits.

A mixer need not be the only thing that works at a certain frequency in order to be useful. Mixers have equally important, if somewhat prosaic, applications in ubiquitous superheterodyne receivers. The primary purpose of a mixer is to translate a signal at one frequency to another where it can be amplified or processed more effectively. The ability to shift a signal in frequency with minimal added noise or distortion is important because the properties of amplifiers, filters, and detectors are all different at different frequencies. In order to perform the functions of filtering, amplification, or detection optimally, it is often necessary to shift signals to the frequencies where these functions can be performed best.

1.2 WLANs:

In 1985, the Federal Communications Commission (FCC) is the U.S. defined the ISM (Industrial, Scientific, and Medical) frequency bands with frequencies at 900 MHz, 2.4 GHZ and 5 GHz which are illustrated in Fig. 1.1.

In recent years, the wireless local-area-network (WLAN) products based on the IEEE 802.11 standards have become widely popular. In fact, products based on this technology—also known as Wi-Fi—have been listed among the most encouraging areas of high-tech growth. To date, most Wi-Fi products worldwide have been limited to spectrum in the 2.4GHz band. Recently, 802.11a products, which utilize the 5-GHz band, have gained a lot of interest in Wireless applications.

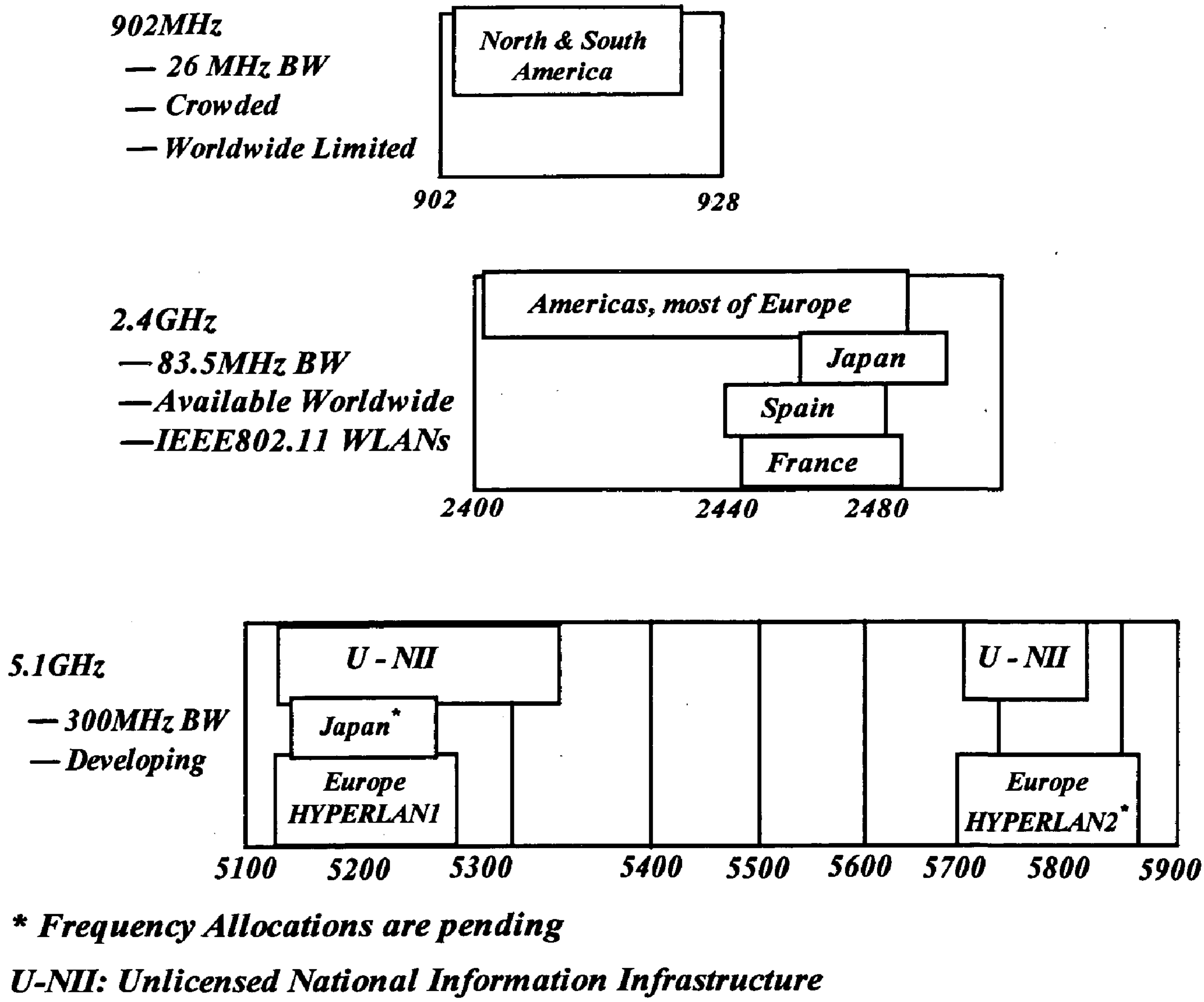


Fig. 1.1 Unlicensed ISM RF Bands

When assessing the pros and cons of 2.4GHz and 5GHz systems, be sure to define requirements. This provides a solid basis for defining all design elements. Without firm requirements, we will be making the choice on flimsy ground.

The following advantages of 5 GHz systems over 2.4 GHz systems are:

- **Performance:** The 5GHz bands have much greater spectrum available in this band there are 12 non-overlapping channels, each with 20MHz of bandwidth. This means significantly better performance as compared to the 2.4 GHz band. The entire 2.4 GHz band is 83.5MHz wide, which only allows three non-overlapping channels. If high performance is an important requirement, then lean toward the 5GHz band.

- **Facility Size:** As frequency increases, range generally decreases. As a result, 5 GHz systems generally have less range than ones operating in the 2.4 GHz band.

- **Radio Frequency (RF) Interference:** 2.4 GHz WLANs can experience interference from cordless phones, microwaves, and other WLANs. The interfering signals degrade the performance of an 802.11b WLAN by periodically blocking users and access points from accessing the shared air medium. A 5GHz system, which is relatively free from interfering sources.

- **Security:** Security of the WLAN is of great concern to most companies. By minimizing the propagation of radio waves outside the physically controlled area of a facility, a wireless network is more secure because of the reduction of the potential for denial of service attacks. As a result, 5 GHz systems can provide enhanced security over 2.4GHz systems because of less range.

Therefore the use of a 5GHz WLAN offers additional benefits (e.g., higher performance, less interference, etc) and this band will become mainstream in the next several years. Japan has started the development of standards for WLAN devices in the 5-GHz band.

A radio consists of a transmitter, receiver, and ancillary functions are needed to support their operation. A complete WLAN radio might look something like the block diagram in Fig. 1.2. Many variants of a simple block diagram of Fig. 1.2 are possible. However, all radios depend on the same set of functional blocks:

- *ADCs* and *DACs* to convert between the analog and digital worlds
- *Amplifiers* to increase the signal power
- ***Mixers*** to convert between frequencies
- *Filters* to select desired frequencies from a multitude of interferers and spurs
- *Switches* to select the required input at the right time

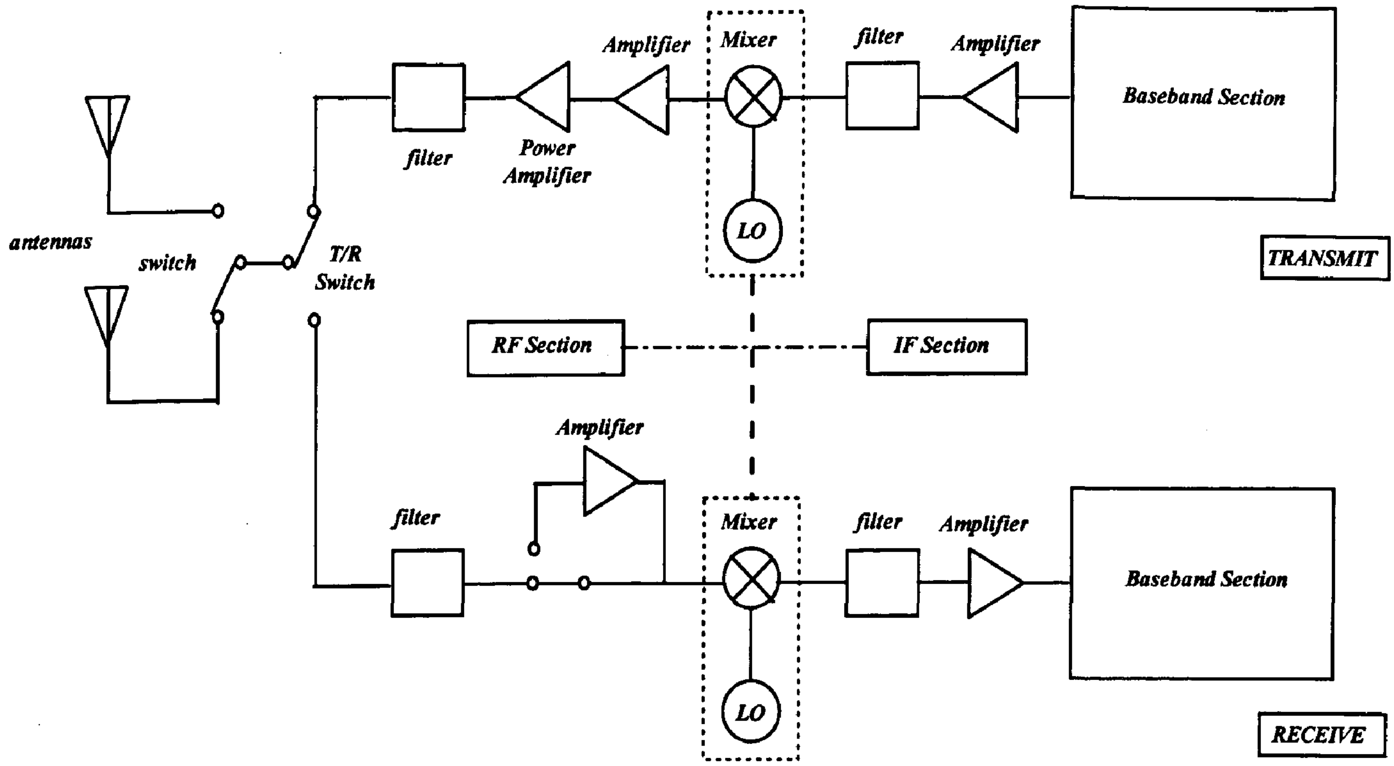


Fig. 1.2 Block Diagram of a typical WLAN Radio Emphasizing Mixer/LO Functions

1.3 Problem Statement:

Since, frequency conversion plays a key role in analog radios which implies mixer plays a vital role in the wireless applications because frequency conversion is possible only through mixers; this thesis is aimed to design the mixer for WLAN operating in 5GHz band.

The objectives of this dissertation are as follows:

- Design of single device Active FET mixer.
- Analyzing the design through simulation.
- Calculation and Measurement of the dielectric constant of the substrate
- Implementation of the designed mixer and comparison of measured and simulated results.

In order to design a RF mixer, firstly the main specifications of the mixer must be obtained according to the system and its working frequency. The values of the main specifications are given in Table 1.1. The mixer is designed to meet these specifications.

Table 1.1 Typical specification values for WLAN (5 GHz) mixer

Conversion Gain	≥ 4 dB
Noise Figure	< 15 dB
P_{1dB} Compression point	~ -15dBm – -5dBm
Isolation	> 25 dB

1.4 Organization of the Report:

This rest of the dissertation report is organized as follows:

Different types and topologies of mixer, their operation, performance specifications, advantages and disadvantages are discussed in *chapter 2*. *Chapter 3* explains the mixer design, measurement of dielectric constant of the substrate, essential components for designing, their design and analysis through simulation. *Chapter 4* discusses the mixer analysis, harmonic balance analysis technique and reliazation of mixer on microstrip.

Chapter 5 provides the simulation results in ADS, measured results and comparison. *Chapter 6* concludes the report. Mixer simulation procedure is discussed in the *Appendix*.

2. MIXER THEORY

Mixers are found in virtually all wireless communications systems. They are frequency translating devices that convert input signals from one frequency to another by mixing these signals from one frequency to another signal of known frequency. One reason frequency translation is a necessary process in wireless transmission is that information signals such as human speech or digital data are usually low frequency signals and are not suitable for a wireless channel. Another is that wireless channels are common channels that are shared by many signals and these signals must be separated into different frequency bins so that electronic circuits (which contain frequency selective components) can keep them from destructively interfering with each other. Among many other properties, frequency is one that is most easily exploited in signal identification. In this chapter we discuss the theory of mixer.

Most mixers are three-port devices, as shown in Fig. 2.1. They take two input signals and the output is a mixing product of these two inputs. There are self-oscillating mixers which provide their own LO signal by having an internal resonating element coupled with the input.

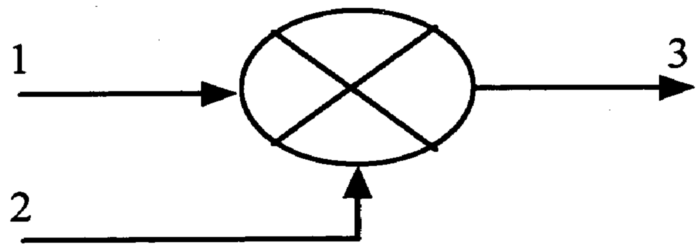


Fig. 2.1 Basic Three Port Mixer

2.1 Mixer Concepts:

As mentioned above, mixer is a frequency converter which produces new frequencies different from those present in the input signal.

Fig.2.2 shows the property of frequency translation using multiplication of two signals. In a transmitter application, the intermediate frequency f_{IF} and the local oscillator signal f_{LO} are mixed together to produce $f_{IF} \pm f_{LO}$, and this is called up-conversion. In a receiver application, an RF signal f_{RF} and a local oscillator signal

f_{LO} are mixed together to produce an intermediate frequency f_{IF}, and much higher frequencies $f_{IF} \pm 2f_{LO}$ are filtered out. This is called down-conversion.

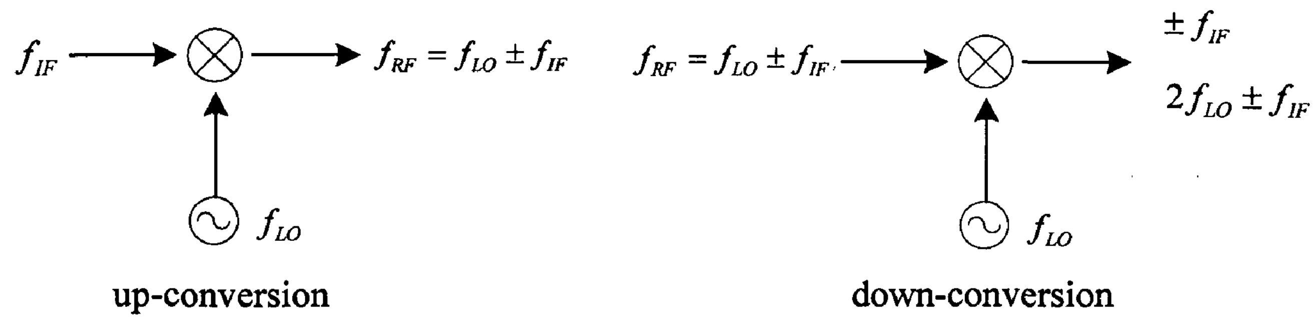


Fig. 2.2 Up-conversion and Down-conversion

Frequency-domain Considerations:

Since mixing involves multiplication in time domain, it involves convolution in the frequency domain. As shown in Fig.2.3, if a baseband signal with the bandwidth f_B is multiplied with the LO signal of $\cos \omega_0 t$, it is split into two signals centered at $\pm f_0$ with the magnitude halved. Again, if the frequency-shifted signal is multiplied with the same LO signal, a half of the original baseband signal is recovered at the baseband, and the unwanted signals are generated at the twice of the LO frequency. Using low pass filter, the desired signal is selected and undesired signals are filtered out.

For example, if the input signal is 10.455 MHz and the LO frequency is 10 MHz in a super-heterodyne receiver given in Fig.2.4, the output of RF mixer will comprise sinusoidal at 455 KHz and 20.455 MHz. If the IF filter has a center frequency of 455 KHz, the 20.455 MHz component will be suppressed. This process has effectively converted the signal down into an IF of 455 KHz from an RF of 10.455 MHz. However, note that while 10.455MHz signal will mix with 10MHz LO signal to produce the IF signal, so will 9.545 MHz signal. 9.545 MHz mixes with 10MHz to 455 KHz and 19.545 MHz. This unwanted RF signal at 9.545 MHz is known as the image of the wanted 10.455 MHz signal.

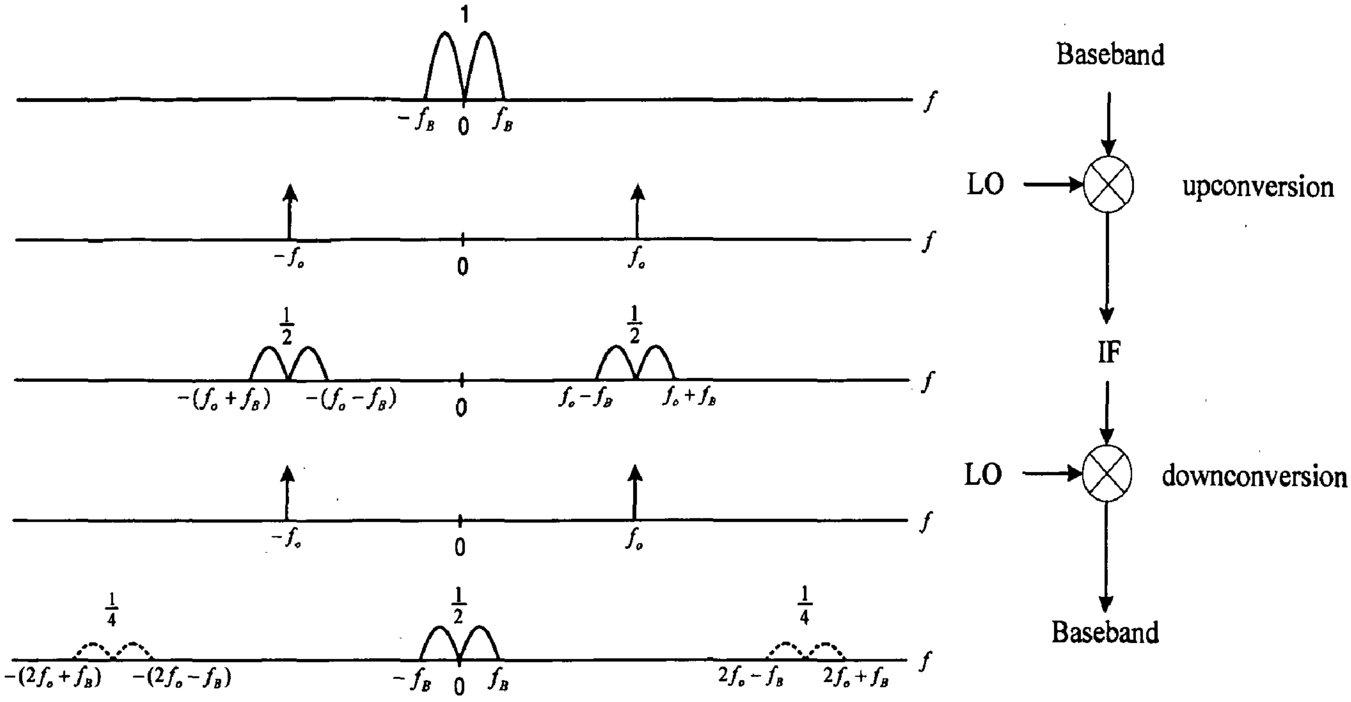


Fig. 2.3 Frequency Spectrum for up-and down-conversion $\pm f_0$

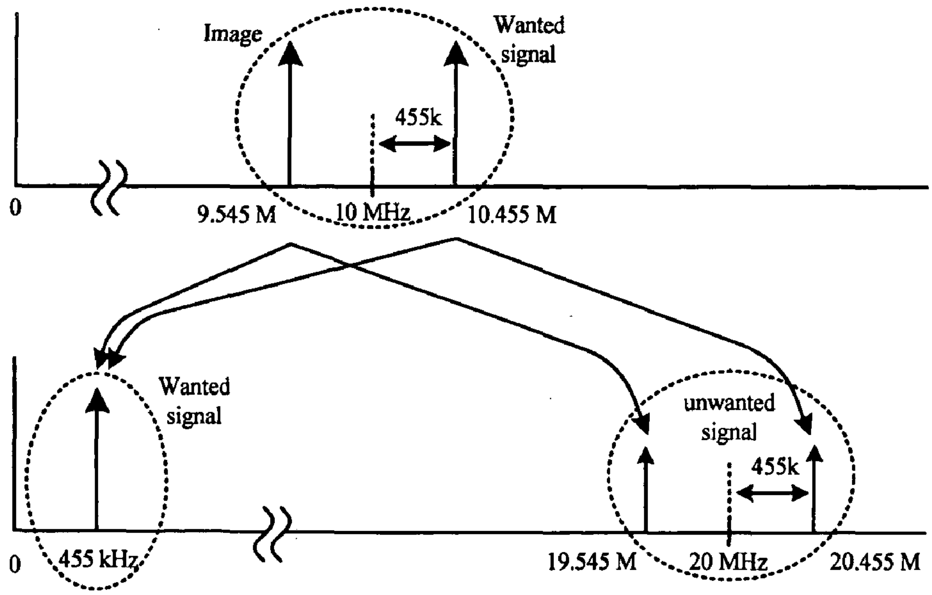


Fig. 2.4 Down-conversion example in Super-heterodyne receiver

Time-domain Considerations:

Mixer changes only the center frequency with the information unchanged. Passing through the mixer, the baseband signal including the information is represented as the envelope and the LO signal as the center frequency as shown in Fig.2.5.

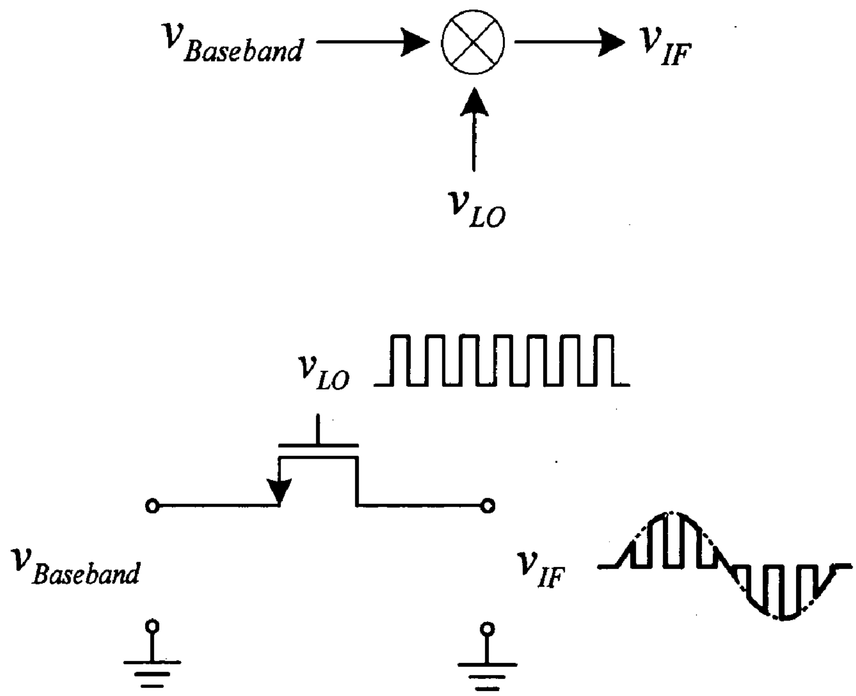


Fig. 2.5 Multiplication of two-signals in time-domain

How to use a nonlinear device as a mixer?

Nonlinear device such as a diode, BJT, or FET can generate multiple harmonics and intermodulated signals, and so can be used as a mixer. Let's consider how a non linear device can generate the sum or difference frequency components of two input signals referring to Fig.2.6.

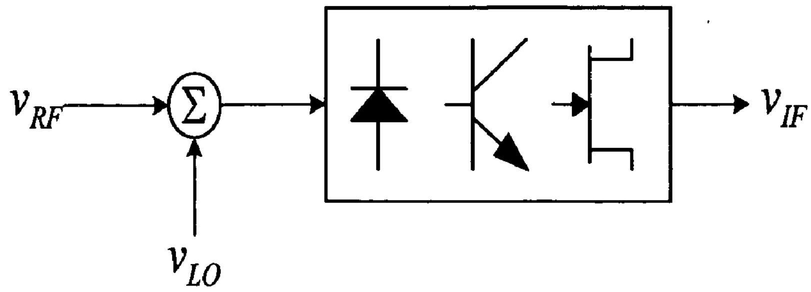


Fig. 2.6 Basic Mixing Concept using nonlinear devices

Both diode and BJT have an exponential characteristic as expressed by the following Shockley diode equation:

$$i_D = I_D + i_d = I_s.(e^{v_D/\eta V_T} - 1) \cong I_s.e^{v_d/\eta V_T}$$, approximated for forward-biased.

If the voltage $v_D = V_D + v_d$ is applied across the diode, the diode current is

$$i_D = I_S.e^{(V_D+v_d)/\eta V_T} = I_S.e^{V_D/\eta V_T}.e^{v_d/\eta V_T} = I_D.e^{v_d/\eta V_T} \tag{2.1}$$

Expanding by Taylors series,

$$\begin{aligned} i_D &= I_D.[1+(\frac{1}{\eta V_T})v_d + \frac{1}{2}(\frac{1}{\eta V_T})^2 v_d^{\,2} + \frac{1}{6}(\frac{1}{\eta V_T})^3 v_d^{\,3} + ...] \\ &= a_0 + a_1.v_d + a_2.v_d^{\,2} + a_3.v_d^{\,3} + ... \end{aligned} \tag{2.2}$$

If the ac signal voltage applied to the diode is $v_d = v_{RF}.\sin\omega_{RF}t + v_{LO}.\sin\omega_{LO}t$, then

$$\begin{aligned} i_d &= a_1.(v_{RF}.\sin\omega_{RF}t + v_{LO}.\sin\omega_{LO}t) + a_2.[\frac{1}{2}v_{RF}^{\,2}(1-\cos 2\omega_{RF}t) \\ &+ v_{LO}v_{RF}\{\cos(\omega_{RF}-\omega_{LO})t - \cos(\omega_{RF}+\omega_{LO})t\} + \frac{1}{2}v_{LO}^{\,2}(1-\cos 2\omega_{LO}t)] + \ldots \end{aligned} \tag{2.3}$$

From this equation, it is clear that the nonlinear devices can generate new frequency components $\omega_{RF} \pm \omega_{LO}$ corresponding to the IF frequency.

2.2 Mixer Performance Parameters

2.2.1 Conversion Gain / Loss:

At the radio frequencies and microwave frequencies, efficiency in transmission of signal power is of great importance. For this reason RF and microwave circuits are optimized for power gain instead of voltage or current gain as commonly found in most low frequency circuits.

The degree to which the frequency shifted signal is attenuated or amplified is an important figure of merit. So "conversion gain" is defined as the ratio of the wanted output signal level to the input, normally expressed in dB. Conventionally negative conversion gain is stated as conversion loss. Passive mixers not using DC power supply cannot amplify the input signal and so exhibit conversion loss. On the other hand, active mixers capable of amplifying the input signal exhibit conversion gain. Conversion loss of 6 to 8dB is usually adequate for passive mixers, and the active mixer gain of unity or few dB is usually optimum. A mixer with high conversion gain reduces the noise contribution from the subsequent stages. But a high conversion gain is not necessarily desirable considering phase margin and distortion.

The "voltage conversion gain" of a mixer is defined as the ratio of the rms voltage of the IF signal to the rms voltage of the RF signal. Note that the two signals are centered around two different frequencies. The voltage conversion gain can be measured by applying a sinusoid at ω_{RF} and examining amplitude of the downconverted component at ω_{IF}.

The "power conversion gain" of a mixer is defined as the IF power delivered to the load divided by the available RF power from the source.

For example, considering just a single sideband in Fig.2.3, conversion loss is 3dB. The reason is clarified as follows. If a RF input signal $A\cos\omega_{RF}t$ is applied to the down-conversion mixer that receives the LO signal $B\cos\omega_{LO}t$, it is given that the conversion gain is B/2 through the following Eq. 2.4.

$$A\cos\omega_{RF}t \cdot B\cos\omega_{LO}t = \frac{AB}{2}\left[\cos(\omega_{RF}-\omega_{LO})t + \cos(\omega_{RF}+\omega_{LO})t\right] \qquad (2.4)$$

2.2.2 Noise:

Noise in electrical systems is defined as random fluctuations in voltage and current. It can be generated internally by components employed in the system or externally by electrical radiation from other systems. RF and microwave oscillators, for example, are very susceptible to external radiation if they are not properly shielded and if they are not sufficiently isolated from physical contacts with nearby objects. Mixers on other hand, most vulnerable to noise that are generated by their own transistors and resistors. Transistors exhibit flicker noise, which is caused by a change in conductance caused by a relatively slow process (e.g. the exchange of charge with surface traps), and shot noise, which is due to random one-way crossings of some barrier by discrete quantities of charge. For mixers at radio and microwave frequencies, flicker noise is negligible since its power spectrum has a 1/f property.

Shot noise power, however, depends on the net total current crossings the pn junctions, and its power spectral density is given by

$$G_i(f) = qI \tag{2.5}$$

where q is the electronic charge and I is the total current. Resistors exhibit thermal noise, which is generated by the random movement of electrons inside the resistive material at a non zero absolute temperature. The thermal noise power (per unit of frequency) of resistors thus depends on temperature and the resistance value of the resistors. However, the available thermal noise power depends solely on temperature. Eq. 2.6 gives the power spectral density of thermal noise

$$G_v(f) = KTR \tag{2.6}$$

where K is the Boltzmann's constant, T is the absolute temperature in Kelvins, and R is the resistance.

The combined effect from the noise sources mentioned above and all other possible noise sources is often treated as though it were caused by only thermal noise. Mixers are specified with noise figure.

Noise figure is noise factor in decibel units (dB) and is an important figure of merit used to characterize the performance of not only a single component but also the entire system. It is one of the factors which determine the system sensitivity. Noise figure is the ratio of the signal-to-noise ratio (SNR) at the input compared to the SNR at the output. When specifying the noise figure of a mixer, distinction must be made as to whether the input is a SSB or DSB signal. Double sideband (DSB) noise figure includes noise and signal contributions at the both RF and image frequencies. On the other hand, considering single sideband (SSB) noise figure, no image signal is included and only the noise power in the IF component is measured. The DSB method assumes that the gain of the DUT (Device Under Test) is the same at both image and intermediate frequencies, so it is not recommended for a narrowband DUT or a high intermediate frequency. The SSB method does not require this assumption but does require an external image frequency filter at the input of the DUT. SSB noise figure is used in most applications where the desired information is contained only in the intermediate frequency and image frequency is rejected.

For simplicity, let us consider a noiseless mixer with unity gain. As shown in Fig.2.7, the spectrum sensed by the RF port consists of a signal component and the thermal noise of R_s in both the signal band and image band. Upon down-conversion, the signal, the noise in signal band and noise in the image band are translated to ω_{IF}. Thus, the output SNR is half the input SNR if the input frequency response of the mixer is the same for the signal band and the image band. In other words, the noise figure of a noiseless mixer is to 3dB.

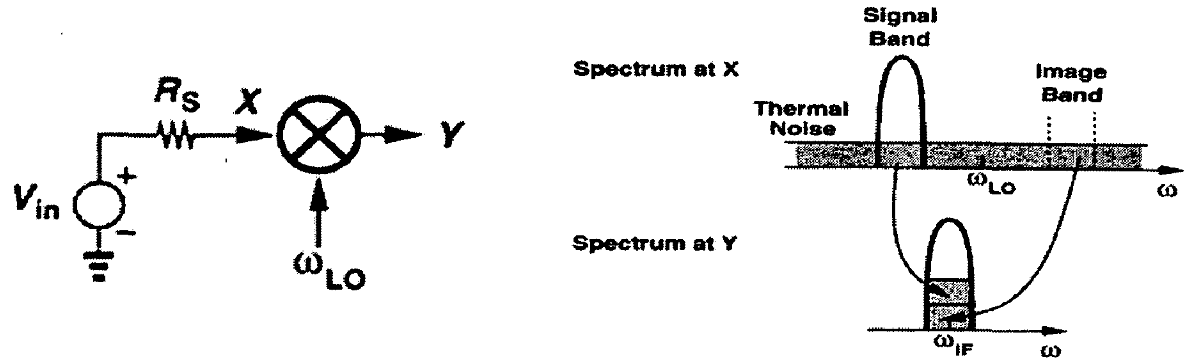


Fig. 2.7 Folding of RF and image noise in to the IF band

The above measurement provides the "single-side band" noise figure (SSB NF) of the mixer. The term SSB indicates that the desired signal spectrum resides on only one side of the LO frequency, a common case in heterodyne systems. Now, consider the homodyne down-conversion of an AM signal by means of a single noiseless mixer Fig.2.8. In this case, the input and output signal-to-noise ratios are equal, giving noise figure of 0dB. This is called the "double-sideband" noise figure (DSBNF) so as to emphasize that the input signal spectrum resides on both sides of ω_{LO}.

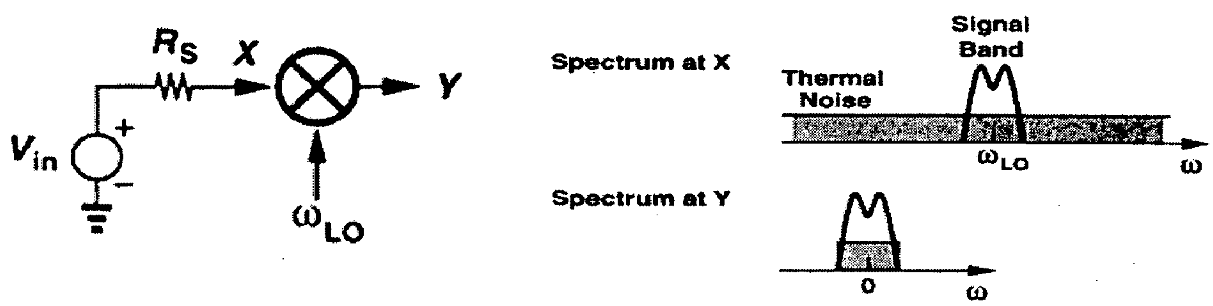


Fig. 2.8 Down-Conversion of an AM signal

In summary, the SSB noise figure of a mixer is 3 dB higher than the DSB noise figure if the signal and image bands experience equal gains at the RF port of a mixer. Typical noise figure meters measure the DSB NF and predict the SSB NF value by simply adding 3 dB.

2.2.3 Port-to-Port Isolation:

Isolation is a measure of how much power is coupled from one port to the next. In the down-conversion mixer, the isolation between LO and RF ports of the mixer is important because LO-to-RF feedthrough results in LO signal leakage to the LNA and eventually the antenna if the mixer is used as first downconverter in a wireless receiver, whereas the RF-LO feedthrough allows strong interferers in the RF path to interact with the local oscillator driving the mixer. Also, large LO and RF feedthrough signals at the IF output port may saturate the IF output port, and decrease the P_{-1dB} of the mixer. LO-to-IF indicates how much LO power leaks through the output IF port. LO appearing at the output IF port can be attenuated easily by a low pass filter since the two frequencies are far apart, but it is more difficult to suppress at the RF port. Finally, the RF-IF isolation determines what fraction of the signal in the RF path directly appears in the IF.

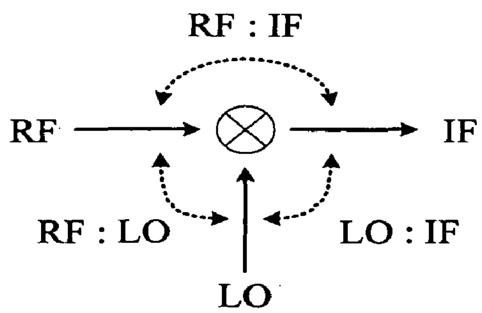


Fig. 2.9 Port-to-Port Isolation

2.2.4 Linearity:

A weakly nonlinear system can be approximated by the following 3rd order polynomial.

$$y(t) = \frac{\alpha_2 A^2}{2} + \left[\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right]\cos\omega t + \frac{\alpha_2 A^2}{2}\cos 2\omega t + \frac{\alpha_3 A^3}{4}\cos 3\omega t$$

$$y(t) = \alpha_1 x(t) + \alpha_2 x^2(t) + \alpha_3 x^3(t) + ... \tag{2.7}$$

Thus the output of such a nonlinear circuit for a sinusoidal input signal $x(t) = A\cos\omega t$ is

$$y(t) = \alpha_1 A\cos\omega t + \alpha_2 A^2 \cos^2\omega t + \alpha_3 A^3 \cos\omega t \tag{2.8}$$

$$y(t) = \frac{\alpha_2 A^2}{2} + \left[\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right]\cos\omega t + \frac{\alpha_2 A^2}{2}\cos 2\omega t + \frac{\alpha_3 A^3}{4}\cos 3\omega t \tag{2.9}$$

In Eq. 2.9, $\cos\omega t$ is the fundamental frequency and rest is higher order harmonics.

An important specification of a mixer is its linearity. Mixers perform frequency translation and realized by either using non-linear or time varying circuit. Thus mixers are inherently non-linear. It is desirable for a mixer to act very linearly with respect to all nonlinearities except the one giving the desired frequency conversion. Two major sources of distortion in the mixer are the device nonlinearities and the nonlinearities coming from the switching devices.

Linearity of a mixer can be estimated with following parameter.

(a).1dB Compression point (P_{1dB}):

A measure of amplitude linearity, 1 dB compression point is the point at which the actual gain is 1 dB below the ideal linear gain. From Eq. 2.9 assuming that all of the fourth and higher order terms are negligible. For a mixer, the fundamental component is the desired output, and it can be written as

$$\alpha_1 A\left[1 + \frac{3}{4}\left(\alpha_3 / \alpha_1\right)A^2\right]$$

This fundamental component is larger than $\alpha_1 A$ (the ideally linear gain) if $\alpha_3 > 0$ and smaller if $\alpha_3 < 0$. For most practical devices, $\alpha_3 < 0$, and the gain compresses as the amplitude A of the input signal gets larger. The 1dB compression point can be expressed in terms of either input power of output power. Measurement of P_{1dB} can be made by increasing the input power while observing the output power until the gain is compressed by 1 dB.

P_{1dB} is an important characteristic of a device and since it indicates the upper limit of the power level of the input signal without saturating the device and generating nonlinear effects. If plotted on a log-log scale as a function of the input level, the output falls below its ideal value by 1 dB at the 1-dB compression point.

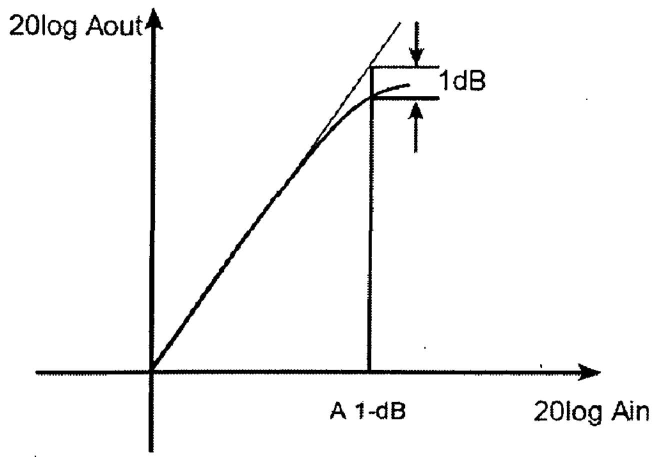


Fig. 2.10 Input 1-dB compression point of the mixer

From the above Fig.2.10 we can see the output level of a mixer fails to track the input level changes proportionally.

2.3 Mixer Topology:

Mixers can be categorized in to passive and active mixers in terms of conversion gain as shown in Fig.2.11. Although passive mixers such as diode mixers are very linear and can operate broadly at very high frequency, they have no conversion gain and so are used in microwave and base station circuits. On the other hand, active mixers provide conversion to reduce the noise contribution from the subsequent stages (for example, the IF stages for down-conversion mixers).

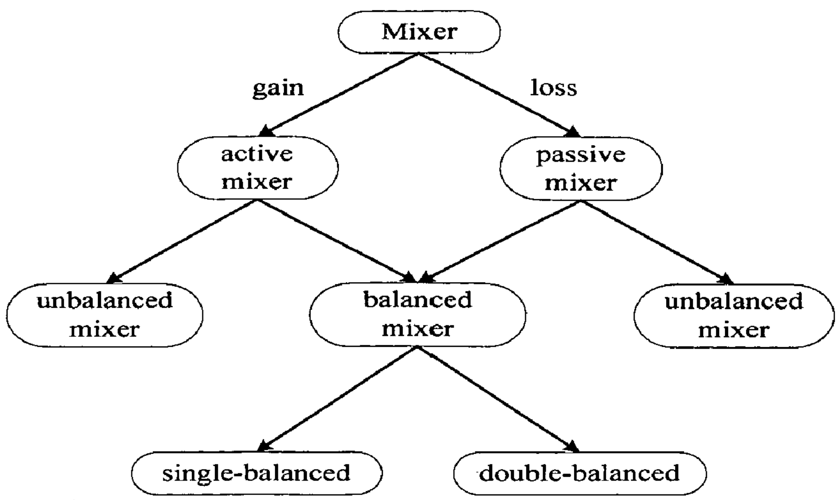


Fig. 2.11 Classification of Mixers

Also mixers are classified in to single-balanced and double-balanced types according to whether input is differential with differential LO input or not.

2.3.1 Passive Mixers:

Diode mixers are one of the examples of passive mixers. They are an example of that does nothing especially well but do most tasks at least well adequately. Such components are the salt of the earth; technology could not exist without them. Diode mixers are low-cost, broadband components that exhibit moderate performance in terms of distortion, conversion loss, port isolation, spurious response rejection, local-oscillator (LO) AM noise rejection, and low frequency noise. Other type of mixers offer better performance for some of these characteristics, but none offers better performance for all of them.

Although mixers are used commonly in communication receivers, they are also used in transmitters and signal generators. Mixers can be employed as modulators and phase detectors as well as frequency converters. The design of mixers for the latter applications is essentially the same as for receivers, except for occasional obvious differences. We examine different types of mixers and their applications in this chapter.

Table 2.1 lists the types of mixers description. It should be possible to find a diode mixer among those listed that is adequate for almost any ordinary application.

Diode Mixer Operation:

Diode mixers are sometimes called resistive mixers, because mixing is caused by a time-varying resistance (or conductance). LO signal drives the mixer and causes device transconductance. Consider a time-varying conductance having the waveform $g(t)$, where

$$g(t) = G_0 + G_1 \cos(\omega_p t) \tag{2.10}$$

We now apply a voltage $V_s(t)$ to this conductance, where $V_s(t) = V_s \cos(\omega_s t)$, a little trigonometry shows the resulting current is $i(t) = g(t) v_s(t)$

$$= G_0 V_s \cos(\omega_s t) + \frac{G_1 V_s}{2}\left[\cos\left((\omega_s - \omega_p)t\right) + \cos\left((\omega_s + \omega_p)t\right)\right] \quad (2.11)$$

The time-varying resistance has generated mixing *products* at the difference frequency $\omega_s - \omega_p$ and the sum frequency $\omega_s + \omega_p$. Usually the difference frequency is the desired output, but occasionally the sum frequency is desired.

In practical diode mixers the situation is not quite so simple. First, the conductance waveform is never perfect sinusoid. In diode mixers it usually is a train of pulses that contain a large number of harmonics. Eq. 2.10 becomes

$$g(t) = G_0 + G_1 \cos(\omega_p t) + G_2 \cos(2\omega_p t) + G_3 \cos(3\omega_p t) + \ldots \quad (2.12)$$

and we obtain mixing products between the RF and all harmonics of ω_p. Thus, the mixing frequencies ω_n, n = 0, ±1, ±2,... are

$$\omega_n = |\omega_s + n\omega_p| \quad (2.13)$$

where n is any integer.

A more common representation for Eq. 2.13 is

$$\omega_n = |\omega_0 + n\omega_p| \quad (2.14)$$

where ω_0 is the frequency $|\omega_s - \omega_p|$. This defines the same set of mixing frequencies as Eq. 2.13, ω_0 is called the *intermediate frequency*, or IF; ω_s is the *radio frequency*, or RF, and ω_p is the *local oscillator* frequency, or LO frequency. The spectrum of mixing frequencies is shown in Fig. 2.11. In deriving Eq. 2.10 we assumed that the voltage across the diode has only an RF (ω_s) component. In fact, the current circulates in the external circuit, generating voltage component at the same frequency as the current components.

Table 2.1 Diode Mixer Description

Mixer Type	Characteristics	Typical Applications
Singly balanced,180-degree ("rat-race") mixer	Approximately 15% of RF and LO bandwidth IF must be less than approximately 15% of the RF/LO frequency. Rejects (2, 1) or (1, 2) spurious response, but not both. Can be dc biased.	Not a general purpose circuit. Best used for noncritical applications in integrated components. A poor choice when a broad-band IF is required.
Singly balanced 90-degree mixer, branch line hybrid	Up to 20% bandwidth. LO and RF port Bandwidths differ. No inherent IMD/spur rejection. Requires a good, broadband source VSWR over both the RF and LO bands at both ports, or imbalance occurs.	Not a good circuit; used more often than it deserves. Can be used for simple, non critical applications.
Double balanced ring mixer; couple line baluns	Multioctave RF and LO bands. Narrowband IF, dc coupled. RF and LO bands can be widely separated.	General-purpose, broad-band applications. The start mixer is better biphase modulator.
Doubly-balanced ring mixer; "horseshoe" balun	Multioctave RF and LO bands. Broadband IF, DC coupled RF and LO bands are widely separated.	Most common type of general purpose, commercial mixer. A very compact circuit.
Star mixer with Marchand-baluns	Octave RF/LO band. Very broadband, DC coupled IF. RF and LO must cover same frequency range.	Where the frequency plan allows it, this is a very good choice. High performance, good balanced, broad band are easily achieved. Good biphase modulator or phase detector.

In general, the diode has both voltages and the current components at all mixing frequencies. These currents and voltages are "coupled" through the mixing process; changing the voltage or current at one mixing frequency affects those at another frequency.

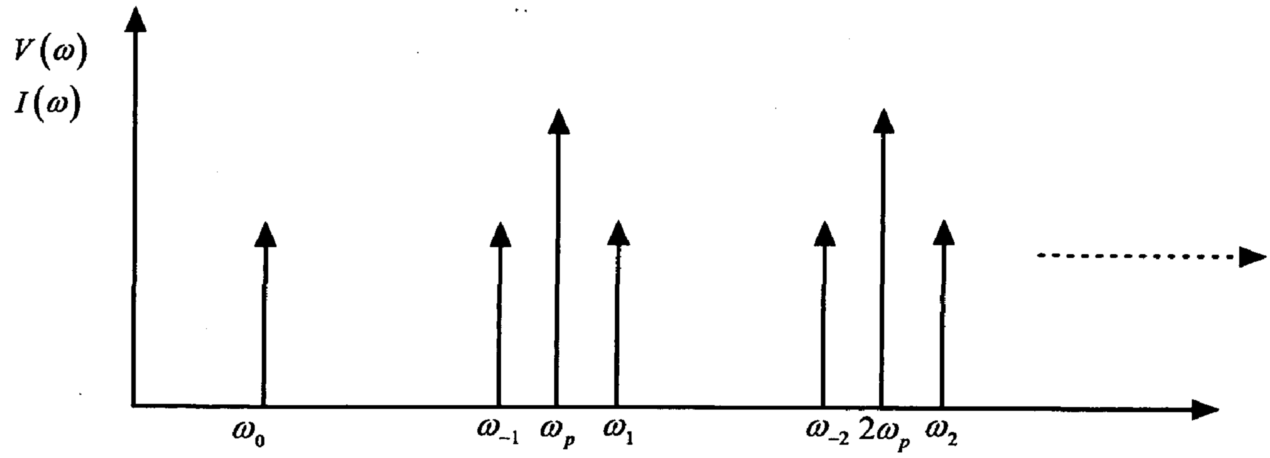


Fig. 2.12 Mixing frequencies ω_n and LO harmonics, $n\omega_p$. The diode junction has voltage and current components at all these frequencies.

2.3.2 Active Mixers:

Diodes and active mixers are very different creatures, so it is difficult to compare their performance. The noise figures of the best FET mixers are only modestly better than those of diode mixers, but the FET mixer's conversion gain improves the overall noise figure and reduces the number of stages in a system. Some higher frequency active mixers exploit the nonlinear characteristics of high gain transistors and can perform the mixing action using only one transistor.

In previous section we saw that a large LO voltage applied to a diode caused its junction conductance to vary that frequency conversion in a diode mixer s caused by this time-varying conductance. In an active mixer we can apply the LO voltage between a transistor's gate and source (or base or emitter) terminals and use the LO to vary its transconductance. Frequency conversion is caused by time varying transconductance. Transistors have number of nonlinearities that can be exploited for frequency conversion. The operation of these mixers is same as diode mixers. Following Table 2.2 will give a brief description about active Mixers.

Table 2.2 Active Mixers Description

Mixer Type	Characteristics	Typical Application
Single device, single-gate FET mixer	Simple, inexpensive. Provides conversion gain, low noise, and low distortion. LO-IF isolation may be poor. Best for moderate bandwidth applications.	Down converters in receivers. The need for LO-RF diplexer and IF filters limits applications.
Single device, dual-gate FET mixer	Good RF-to-LO isolation without filters. Gain and noise figure are somewhat worse than single gate mixer, distortion is about the same. Moderate bandwidth.	A nice mixer for low cost integrated circuits, especially for commercial applications.
Single Balanced FET mixer	Essentially the same characteristics as a single-device mixer, but good LO-RF and LO-IF isolation without a diplexer, 3-dB higher LO power, 3-dB better IP. May require baluns.	High-performance IC applications where the number of FETs and the size of baluns are acceptable. The large number of interconnections makes this a poor choice for hybrid circuits.
Doubly Balanced FET Mixer	Essentially the same characteristics as an single-device mixer, but has all the benefits of a doubly balanced circuit (see diode mixers table),6 dB higher LO power, 6-dB better IP, May require baluns.	Essentially the same application as the singly balanced mixer, but where improved intermodulation rejection justifies the extra complexity and LO power.
Gilbert Cell Mixer	A doubly balanced transistor mixer. Often works well without baluns. Can be used as a "linear" analog multiplier.	Modulators, signal processing as well as mixing.

In active mixer, the RF and the LO voltage, the LO voltage creates the time-varying transconductance, and the RF mixes with it.

The design of an active mixer is based on a single idea; maximize the magnitude of the fundamental-frequency of the transconductance. The reason is similar to the reasoning in previous section 2.3.1. This results in the greatest conversion efficiency and the lowest noise figure.

The drain current, $i_d(t)$, is simply

$$i_d(t) = G_m(t) v_g(t) \tag{2.15}$$

where $G_m(t)$ is the transconductance waveform and $v_g(t)$ is the gate voltage. In general $G_m(t)$ is periodic but not sinusoidal, so we can express it as a Fourier series,

$$G_m(t) = G_{m0} + G_{m1}\cos(\omega_p t) + G_{m2}\cos(2\omega_p t) + G_{m3}\cos(3\omega_p t) + \dots \tag{2.16}$$

where ω_p is the LO frequency and, for simplicity, we have omitted phase terms in each component. If we substitute the RF signal, $v_g(t) = V_s\cos(\omega_s t)$, and (2.16) into (2.15), the drain current becomes

$$\begin{aligned} i_d(t) = G_{m0}V_s\cos(\omega_s t) + \frac{G_{m1}v_s}{2}\left[\cos\left[(\omega_s - \omega_p)t\right] + \cos\left[(\omega_s + \omega_p)t\right]\right] \\ + \frac{G_{m2}V_s}{2}\left[\cos\left[(\omega_s - 2\omega_p)t\right] + \cos\left[(\omega_s + 2\omega_p)t\right]\right] + \dots \end{aligned} \tag{2.17}$$

Our desired output is the $\omega_s - \omega_p$ component of the drain current; the rest is excess baggage and must be eliminated. The only component of the conductance waveform that affects our desired output is G_{m1}. Thus, to maximize the IF output current, we must maximize $G_m(t)$, the fundamental frequency component of the transconductance waveform, $G_m(t)$.

2.3.3 FET Resistive Mixers:

FET resistive mixers [11] are designed for low-distortion applications. They exhibit very low levels of both IMD and spurious responses and their 1-dB compression points are very high. Other performance characteristics – noise, conversion loss and LO power-are similar to those of diode mixers. Low IMD is not their sole purpose, however; these mixers have a number of other delightful properties, including low sensitivity to AM noise and low $1/f$ noise. They also are a convenient solution to the problem of realizing mixers in certain monolithic technologies where Schottky-barrier diode does not exist.

Like diode mixers, FET mixers can be designed as single-device circuits, singly balanced circuits and doubly balanced circuits.

Below Table 2.3 lists the mixers description. They include both single device and balanced circuits and include applications for both RF and microwave.

TABLE 2.3 FET Resistive Mixers Description

Mixer Type	Characteristics	Typical Application
Single-Device FET Resistive Mixer	A good circuit for general use. Requires an LO-IF Diplexer	General applications; lack of an IF balun or hybrid makes this mixer somewhat more practical for ICs. Rejects AM LO noise. A nice circuit for RFICs.
Singly Balanced Mixer	Lower distortion, automatic drain short circuit for the LO. Requires an IF balun or 180 degree hybrid. Requires filtering to separate the RF and IF	Especially useful when the IF is low enough to allow a wire-wound transformer for the output balun. In low frequency mixers using transformers, the IF and RF may overlap.
Doubly balanced ring Mixer	Very low distortion but difficult to design optimally at high frequencies because of large layout parasitics. Bandwidth is limited by hybrids.	A very good circuit for general wireless, RF and lower microwave applications.

FET resistive mixers [12] are an approximation of an ideal known as *linear mixing,* mixing by a linear resistance. Such mixing, if it actually could be achieved, is theoretically capable of distortion less frequency conversion. Although perfection, especially in electronics, is a scarce commodity, the FET resistive mixer comes as close to true linear mixing as anything invented so far.

In Section 2.3.1, Equations (2.10) - (2.14), we saw that pumping a Schottky-barrier diode with a large signal turns its junction in to a time-varying, small-signal conductance. This time-varying then mixes with the applied RF voltage, resulting in mixing products involving the RF frequency and the conductance waveform's fundamental frequency and harmonics. The key equation in this derivation is

$$i(t) = g(t)v_s(t) \tag{2.18}$$

where $g(t)$ is the conductance waveform, $v_s(t)$ is the applied voltage, and $i(t)$ the resulting current.

It is critical to note that Eq. 2.18 is a *linear equation.* The process of frequency conversion, which it describes, is a linear one, and an ideal mixer is a linear component. True, we normally must use nonlinear devices to realize mixers, and this results in distortion, harmonic generation, spurious responses, and other nonlinear phenomena. Nevertheless, Eq. 2.18 shows us that there is no fundamental reason why a nonlinear device should be necessary to produce frequency mixing. We do, however, need the time varying conductance.

In this chapter basic mixer theory is described, operation of different types of mixers are also described. Description on different topologies in various types of mixer are made and mixer performance specifications i.e. mixer figure of merits such as gain/loss, noise figure, isolation, linearity are introduced and described. In the next chapter we explain about the mixer design, determination and measurement of dielectric constant of the substrate, mixer individual components design and analysis.

3. DESIGN OF ACTIVE INTEGRATED CIRCUIT MIXER

3.1 Introduction:

In the previous chapter we discussed theory various types and topologies of mixer. Single gate mixer [13] which involves only single device is chosen for designing the mixer. Single-FET mixer is a prototype for balanced circuits, in this mixer (downconversion), both the RF signal and the LO voltage are applied to the gate. Advantages and disadvantages are discussed before, one difficulty in single-device mixers, is the need for special structures to combine RF and LO. Usually, some type of duplexer is used, although a directional coupler also can be used at the expense of high loss for either the RF and LO.

This mixer requires some type of circuit to combine the RF and LO signals. The most common tool for doing this is a filter diplexer, which consists of two filters, one for the RF and one for the LO, whose passbands do not overlap. The outputs are connected in parallel and the design of both filters must account for this.

Filter diplexer can be tricky to design if the RF and LO bands are closely spaced in frequency. Achieving adequate RF rejection in the LO filter and LO rejection in the RF filter requires complex structures, which usually cause the diplexer to be much larger and more complex than rest of the mixer.

Another approach is to use a directional coupler. Generally, the RF is applied to the "through" port and the LO to the coupled port. This is much simpler than a diplexer but results in several decibels of LO power loss and not insignificant RF loss. For example, if 1 dB of RF loss is acceptable, the coupling—and therefore the LO loss—can be no stronger than -7dB. That much LO loss is usually unacceptable.

A third possibility is to apply the LO to one of the terminals of the mixer. A little thought, however, reveals that a FET has three terminals, and any single excitation voltage can be applied only between a pair of them. The urge to apply the LO to the FET's drain is almost irresistible. This mode of operation has an apparent advantage; the RF and LO are isolated.

To put it simply, drain-pumped FET mixers (sometimes simply called a drain mixer), don't work very well. They have never exhibited noise figures and gain as good as gate-pumped transconductance mixers, and their distortion is about the same. They do not have enough transconductance variation for efficient transconductance mixing, and the finite average drain-to-source resistance dissipates output power. Drain-pumped mixers would have withered from the technological consciousness a long time ago, but for a few early papers [14, 15] indicating that such mixers had unusually low IMD. Although these papers were not really conclusive, the idea that drain-pumped mixers have low distortion somehow became conventional wisdom. So we cruelly disparaged the practice of applying the LO to the drain and weren't terrible enthusiastic about applying it to the source.

Another possibility is to apply the LO to the terminal of the mixer other than the gate and drain. Even though source pumping is done from time to time, it is really not as promising as it might appear. Until we think a little less superficially about this configuration, it sounds like a distinct mode of operation. In source pumped mixer [16], the LO must be applied between the gate and the source. These practical considerations sometimes make a source-pumped mixer worthwhile. For better or worse, source-pumped mixers are used only rarely.

This chapter explains the design of mixer i.e. single-Gate active FET mixer, their individual blocks design and their simulation. Some essential components are implemented and their characteristics are observed. The Fig. 3.1 is idealized single gate FET mixer; the bias sources are not shown. The combiner usually is a diplexer or, an occasionally, a directional coupler. We also explain the determination and measurement of the dielectric constant of the substrate.

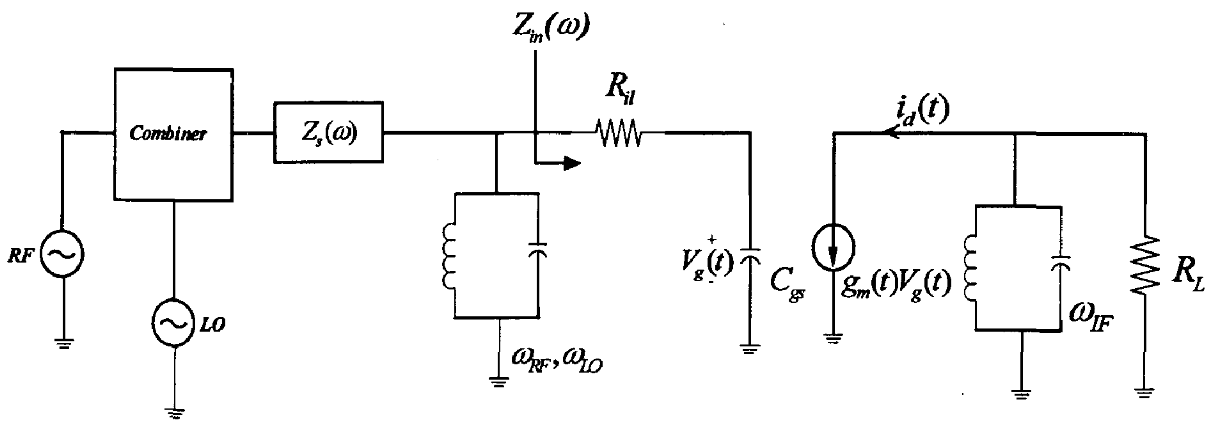


Fig. 3.1 Simplified equivalent circuit of the single-gate FET mixer

3.2 Mixer Description:

The mixer circuit is shown in Fig. 3.1. It is appropriate for use as a downconverter throughout the RF, microwave, and millimeter-wave ranges, as long as the bandwidth is moderate (~20% or less, depends up on diplexer) and the IF frequency is low, no more than 10% of the RF. Because of a wide variety of possible approaches, we have included a design for a combiner, IF filter. In fact, it is best to design the mixer with these components.

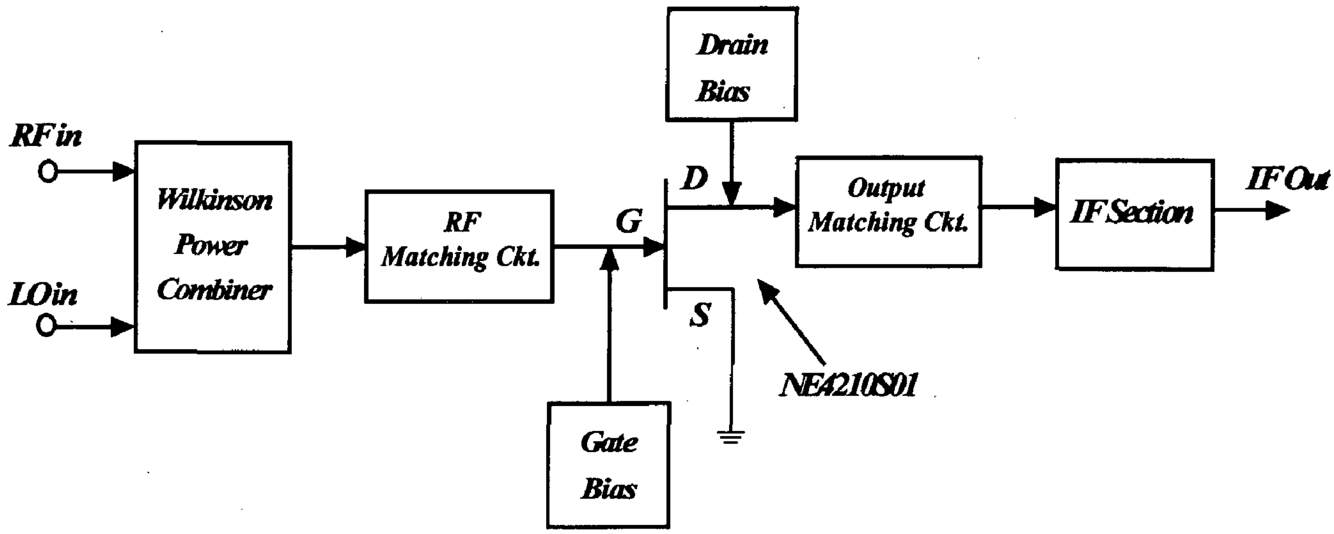


Fig. 3.2 Schematic of Active Single-gate FET mixer

The mixer consists of a Wilkinson power combiner, input matching circuit, output matching circuit, bias circuits and filter circuit. The structures in this design are ones we have found useful.[1] We describe the design of this mixer only, we already added few words of LO injection above in this section.

The RF and LO power combine in Wilkinson power combiner, the RF and output matching circuit will provide maximum gain and minimum reflection of combined power. The transistor will generate harmonics and IF signal and LPF will attenuate the harmonics and pass the IF signal. The bias sources will operate the transistor in saturation region.

3.3 Determination of dielectric constant of the substrate:

The microstrip ring resonator was first proposed by Troughton for the measurements of phase velocity and dispersion of microstrip lines. Compared to the microstrip linear resonator, the microstrip ring resonator does not suffer from open-ended effects and can be used to give more accurate measurements. Since its introduction, in 1969, the microstrip ring resonator [17] has found applications in determining optimum substrate thickness, discontinuity parameters, effective dielectric constant and dispersion and loss and Q-measurements.

This section explains the dielectric constant measurement applications of using ring resonators of the substrate. As we locally purchased the PCB sheets and used in this mixer design, it is essential to find the dielectric constant of the substrate. In the design of mixer and implementation one should know the dielectric constant of the substrate to get the transmission line length at a specified frequency.

3.3.1 Dielectric Constant Measurement:

The ring resonator is a simple circuit. The structure would only support waves that have an integral multiple of the guided wavelength equal to the mean circumference. The circuit is simple and easy to build. For such a simple circuit, however, many more complicated circuits can be created by cutting a slit, adding a notch, cascading two or more rings, implementing some solid-state devices, integrating with multiple input and output lines, and so on. These circuits give various applications.

The ring circuit is an ideal tool for dielectric constant measurements [18]. When Troughton first introduced the idea of a microstrip ring resonator, he described techniques used to measure the phase velocity and dispersive characteristics of a microstrip line by observing the resonant frequency of the ting resonator. The ring resonator, shown in Fig. 3.1 is merely a transmission line formed in a closed loop. The basic circuit consists of the feed lines, coupling gap, and the resonator. The feed lines couple power into and out of the resonator. The feed lines are separated from the resonator by a distance called the coupling gap. The size of the gap should be large enough such that the fields in the resonator are not appreciable perturbed, yet small enough to allow adequate coupling of power. This type of coupling is described in the literature as "loose coupling."

Loose coupling is a manifestation of the negligibly small capacitance of the coupling gap. If the feed lines are moved closer to the resonator, however, the coupling becomes tight and the gap capacitances become appreciable.

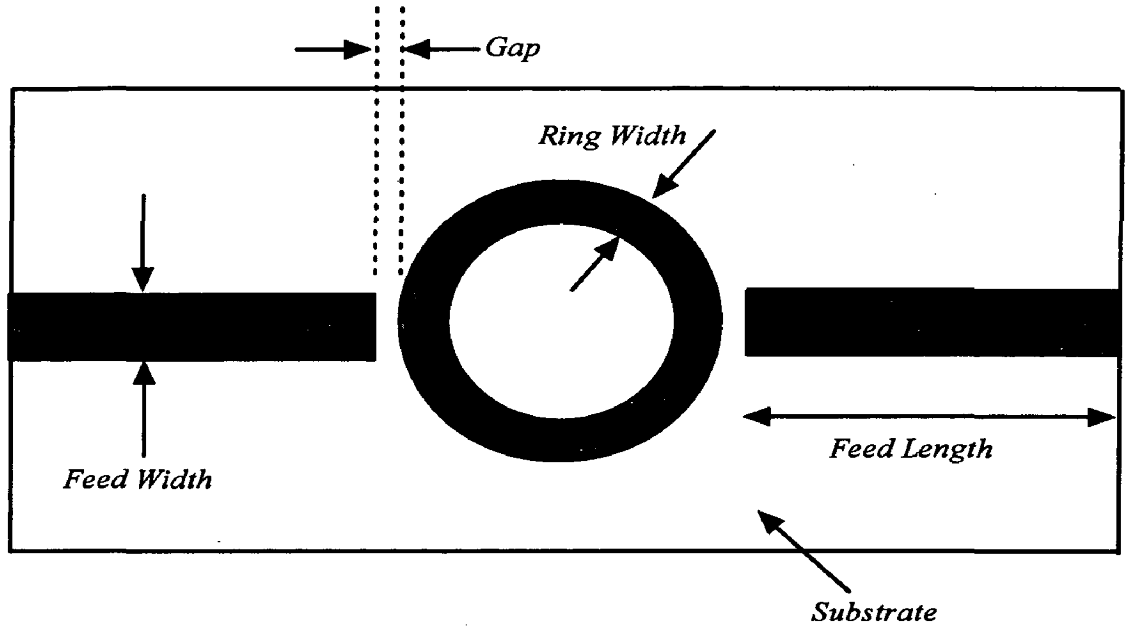


Fig. 3.3 The microstrip ring resonator

This causes the resonant frequencies of the circuit to deviate from the intrinsic resonant frequencies of the ring.

When the mean circumference of the ring resonator is equal to an integral multiple of a guided wavelength, resonance is established. This may be expressed as

$$2\pi r = n\lambda_g \text{, for n} = 1,2,3,\ldots \tag{3.1}$$

where r is the mean radius of the ring that equals the average of the outer radius and inner radii, λ_g is the guided wavelength, and n is the mode number or number of wavelengths on the ring. This relation is valid for the loose coupling case, as it does not take into account the coupling gap effects. From Eq. 3.1, the resonant frequencies for different modes can be calculated since λ_g is frequency dependent. For the first mode, the maxima of field occur at the coupling gap locations, and nulls occur 90^0 from the coupling gap locations.

There exists in a nondispersive medium a linear relationship between the frequency and the phase constant or wavenumber, β where

$$\beta = {}^{2\pi}\!/_{\lambda_g} \tag{3.2}$$

If the frequency doubles, then likewise the wavenumber doubles. In a dispersive medium this is not true. The microstrip line is a dispersive medium.

The dispersion in a microstrip line can be explained by examining the effective permittivity, ε_{eff}. In microstrip the effective permittivity is a measure of the fields confined in the region beneath the strip. In the case of very narrow lines or a very low frequency the field is almost equally shared by the air ($\varepsilon_r = 1$) and the substrate so that, at this extreme,

$$\varepsilon_{eff} \approx \frac{1}{2}(\varepsilon_r + 1) \text{ as } f \to 0 \tag{3.3}$$

where ε_r is the relative dielectric of the substrate. For very wide lines or a very high frequency nearly all of the field is confined to the substrate dielectric, and therefore at this extreme,

$$\varepsilon_{eff} \approx \varepsilon_r \; as \; f \to \infty \tag{3.4}$$

It is therefore obvious that the effective permittivity is frequency dependent, increasing as the frequency increases.

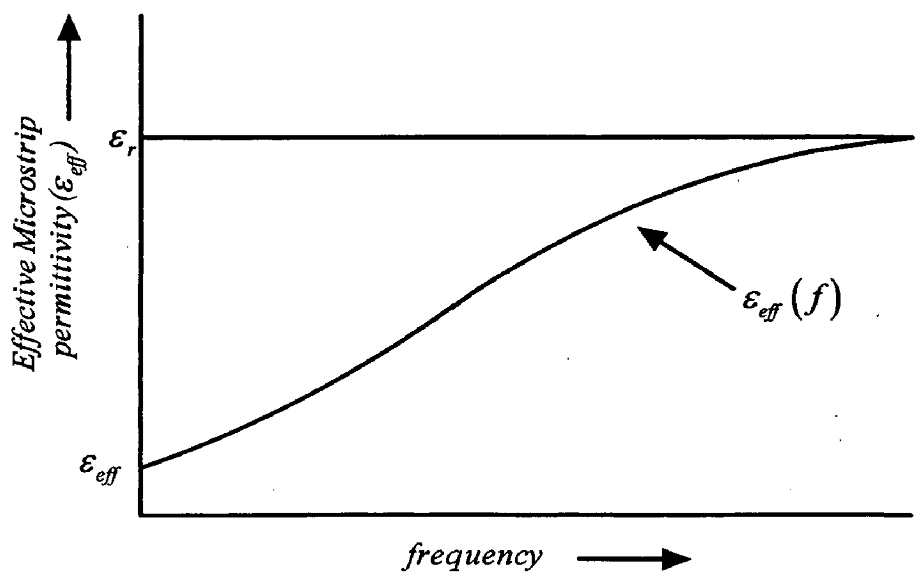


Fig. 3.4 $\varepsilon_{eff}(f)$ dispersion vs microstrip frequency

The effective permittivity is defined as the square of the ration of the velocity in free space, c, to the phase velocity, v_p, in microstrip, or

$$\varepsilon_{eff}(f)=\left(\frac{c}{v_p}\right)^2 \tag{3.5}$$

For any propagating wave, the velocity is given by the appropriate frequency wavelength product. In the microstrip line, the velocity is $v_p = f\lambda_g$. Substituting for v_p in Eq. 3.5 results in the equation

$$\varepsilon_{eff}(f)=\left(\frac{c}{f\lambda_g}\right)^2 \tag{3.6}$$

If we assume that, as in Eq. 3.1, any microstrip resonator will only support wavelengths that are an integral multiple of the total length, then

$$l_t = n\lambda_g \tag{3.7}$$

where l_t is the total length of the resonator. Substituting for λ_g in Eq. 3.4 yields the equation

$$\varepsilon_{eff}(f)=\left(\frac{nc}{fl_t}\right)^2 \tag{3.8}$$

If the total length of a resonator, the resonance order n and the resonant frequency are known, then ε_{eff} can be calculated from Eq. (3.8).

$$\varepsilon_{eff}=\frac{\varepsilon_r+1}{2}+\frac{\varepsilon_r-1}{2}\left(1+10\frac{h}{w}\right)^{-0.55} \tag{3.9}$$

where h and w is the thickness of the substrate and width of the microstrip line. If the static effective dielectric constant ε_{eff} is obtained, the relative dielectric constant ε_r of the substrate can be extracted from Eq. 3.9. Thus from the above description ring resonator procedure is the best technique to find the dielectric constant of the substrate.

3.3.2 Procedural Steps to find the dielectric constant of the substrate:

- First and foremost step to start is to find the thickness of the substrate; we found the thickness of the unknown substrate with the help of Vernier Caliper and measured as 1.6mm.
- Next two assumptions are made to calculate the mean radius of the ring.
  - Initial good guess, assumed the dielectric constant as 4.
  - Impedance of the ring, gap between the ring and field lines are assumed as 50 ohm and 0.5μm respectively.
- From the Eq. 3.1, 3.9 and 3.10

$$Z_o = \begin{cases} \dfrac{60}{\sqrt{\varepsilon_r}} \ln\left(\dfrac{8H}{W} + \dfrac{W}{4H}\right) & \textbf{for } \dfrac{W}{H} \le 1 \\[2ex] \dfrac{120\pi}{\sqrt{\varepsilon_r}\left(\frac{W}{H} + 1.393 + 0.667\ln\left(\frac{W}{H} + 1.444\right)\right)} & \textbf{for } \dfrac{W}{H} > 1 \end{cases} \qquad (3.10)$$

  The radius of the ring is found by assuming the resonant frequency at 2.4GHz and mode number n = 2.

- From the above step , find the following parameters:

  R = 11.2759 mm

  W = 3.33 mm

  r_{out} = 12.9409 mm,

  r_{in} = 9.6109 mm,

  ε_{eff} = 3.115

- Fabricated the ring resonator as shown in Fig. 3.1 using above parameters on the unknown substrate and tested on Hewlett-Packard Vector Network Analyzer (VNA).
- After test it is found that the circuit resonates at 2.3525 GHz.
- Now using this resonant frequency in Eq. 3.1 to calculate ε_{eff} and found as ε_{eff} = 3.25.

- Once again the dimensions of the ring resonator are found with the new resonant frequency obtained after test.
- Fabricated once again with new dimensions and tested the ring resonator on VNA, now this circuit resonated at 2.3509 GHz.
- From the Eq. 3.9 and $\varepsilon_{eff} = 3.25$, it is found that $\varepsilon_r = 4.21$.

From the above procedure and from the obtained results it concludes that the dielectric constant of the substrate is 4.21. The measurement results are shown in Fig.3.3 and 3.4.

3.3.3 Results:

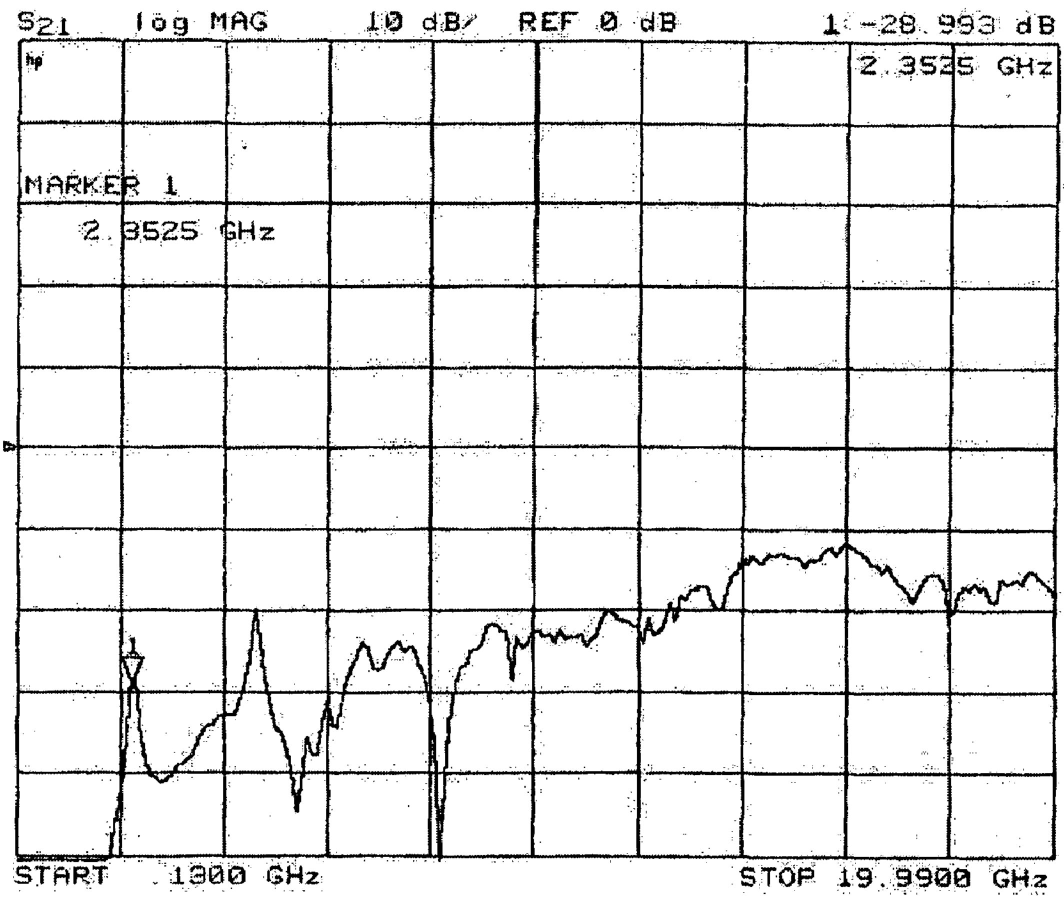


Fig. 3.5 Measured Response on HP-VNA (Initial assumption)

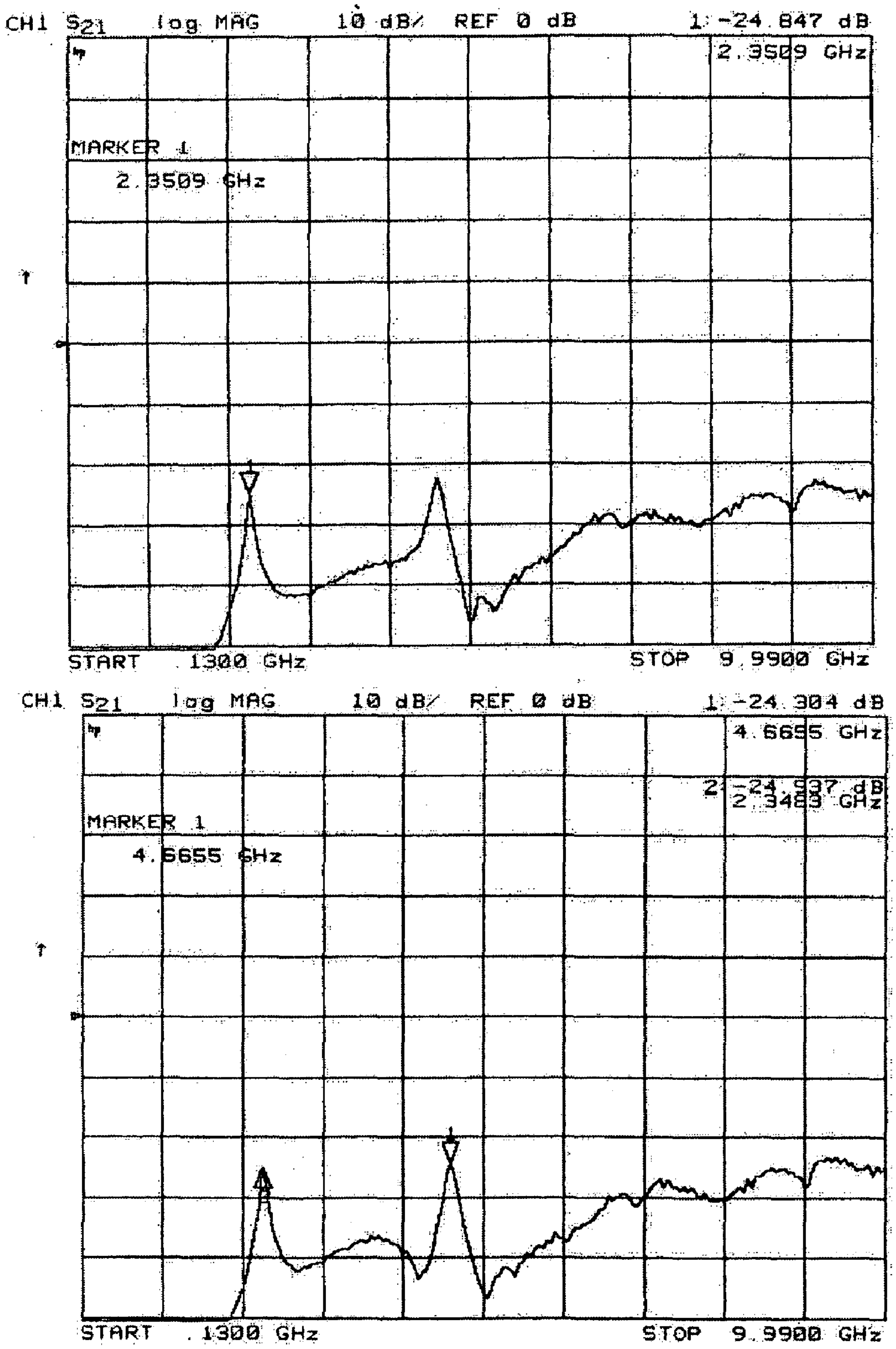


Fig. 3.6 Measured Responses on HP-VNA (Actual Value of ε_r)

Thus with solid measurement technique we measured the dielectric constant of the substrate by using a ring resonator. We fabricated and measured the circuit twice; first with initial assumption and then with actual dielectric constant. Results obtained from VNA are shown. Finally we conclude that the substrate used in the mixer design will have a dielectric constant of 4.21 with thickness 1.6mm.

3.4 RF Section:

In RF section, we design the RF circuit of the mixer, and observe its characteristics as an amplifier.

3.4.1 Wilkinson power divider/combiner:

Power combiners/dividers are used to combine power or divide the power, works as a diplexer. Wilkinson power divider [19] which splits an input signal in to two equal phases output signals (a three port circuit), the most common in practice by far, (but Wilkinson described N-way divider). Power dividers can also be used as power combiners because they are bi-directional.

With respect to the concept Wilkinson made, we can also use those concepts or ideas for combining two equal-phase signals in to one. Power divider consists of two quarter wave transformers to match the port to the output ports. Resistor was added to the circuit this not only matches all three ports, and it fully isolated the ports at the center frequency. Because a lossless reciprocal three-port network cannot have all ports simultaneously matched.

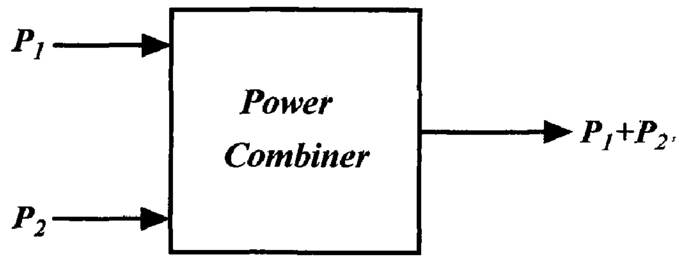


Fig. 3.7 Power Combiner

The circuit design of Wilkinson power divider shown in Fig. 3.8:

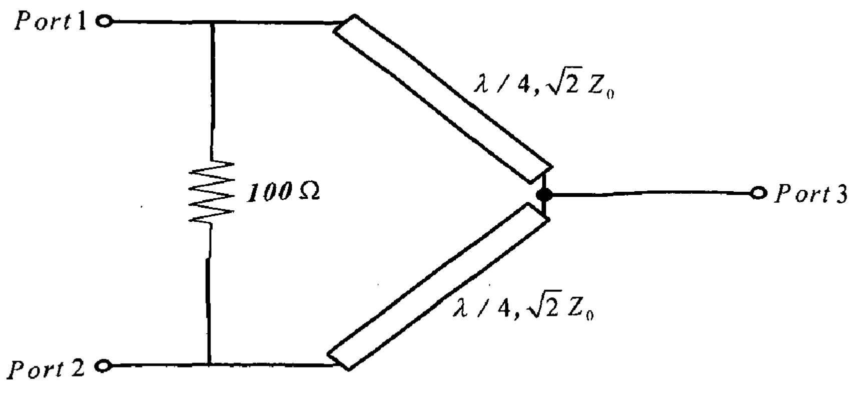


Fig. 3.8 Ideal two-port Wilkinson Power Combiner

It has useful property of being lossless when the output ports are matched; that is only reflected power is dissipated. Wilkinson relied on quarter-wave transformers to match the split ports to the common port. The analysis of this circuit can be done by reducing it to two simpler circuits driven by symmetric and asymmetric sources at the output ports of the Wilkinson power divider.

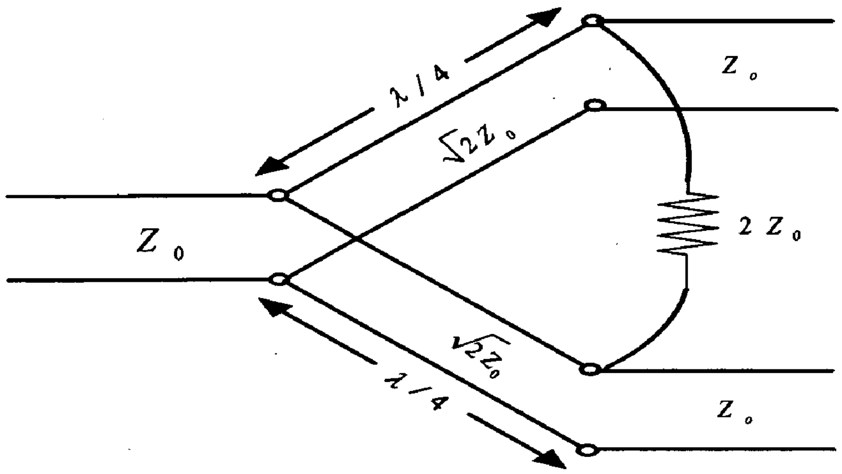


Fig. 3.9 Transmission line circuit of Wilkinson power divider

Normalizing all impedances to the characteristic impedance Z_0, the Fig. 3.9 with voltage generators at the output ports is as shown in Fig. 3.10. This network has been drawn in a form that is symmetric across the midplane; the two source resistors of normalized value 2 combine in parallel to give a resistor of normalize value 1, representing the impedance of matched source. The quarter-wave lines have normalized characteristic impedance Z, and the shunt resistor of normalized value of r . We will find the values of r and Z as in Fig. 3.10.

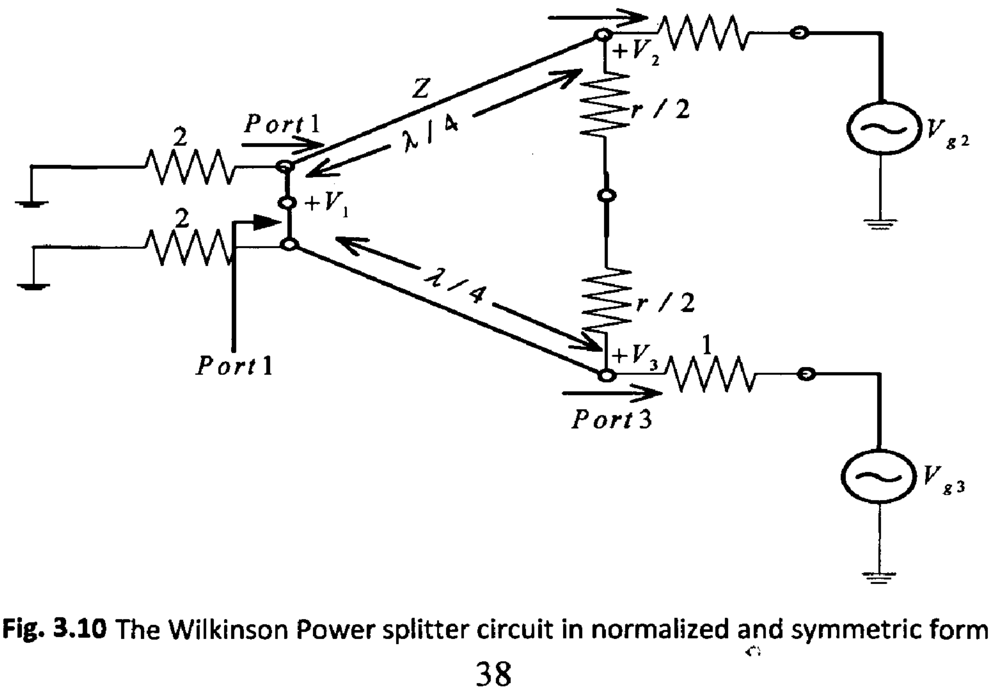


Fig. 3.10 The Wilkinson Power splitter circuit in normalized and symmetric form

The power divider is excited by two separate modes, the even mode, where $V_{g2} = V_{g3} = V$, and the odd mode, where $V_{g2} = V_{g3} = V$.

For even mode excitation, $V_{g2} = V_{g3} = V$, and so $V_2^e = V_3^e$ and there is no current flow through the resistors $r/2$ or the short circuit between the inputs of the two transmission lines at port 1. Thus we can bisect the network of Fig. 3.10 with open circuits at these points to obtain the network of Fig. 3.11 (the grounded side of the λ/4 line is not shown). Then, looking into port 2, we see impedance

$$Z_{in}^{\ 2} = \frac{Z^2}{2}, \tag{3.11}$$

since the transmission line looks like a quarter-wave transformer. Thus, if $Z = \sqrt{2}$, port 2 will be matched for even mode excitation.

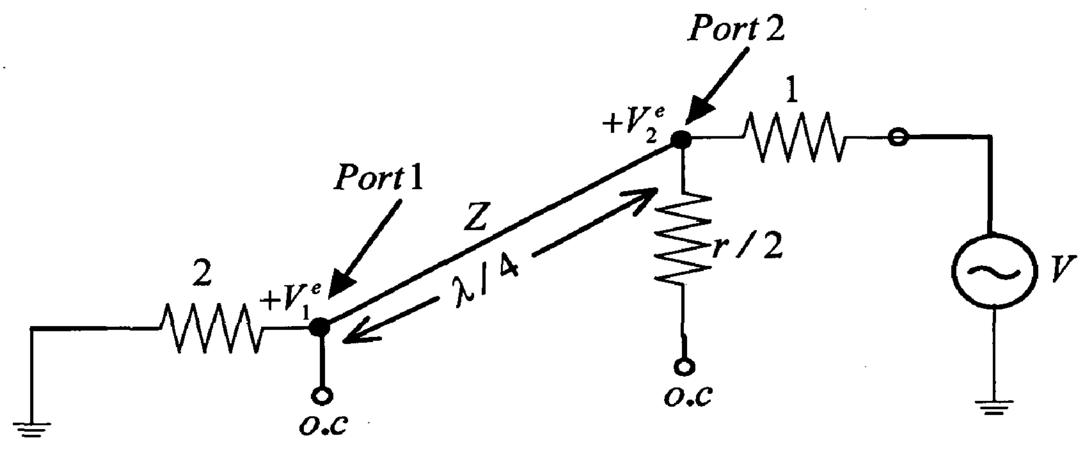


Fig. 3.11 Even mode Excitation

In odd-mode excitation, $V_{g2} = - V_{g3} = V$, and so on $V_2^o = -V_3^o$, and there is a voltage null along the middle of the circuit in Fig 3.10. Thus , bisecting this circuit by grounding it at two points on its midplane to give the network of Fig. 3.12, Looking into port 2, we see an impedance $r/2$, since the parallel connected transmission line is λ/4 long and shorted at port 1, and so looks like an open circuit at port 2. Thus, port 2 will be matched for odd mode excitation if $r = 2$. Then $V_2^0 = V$ and $V_1^0 = 0$; for this mode of excitation all power is delivered to the $r/2$ resistors, with none going to port 1.

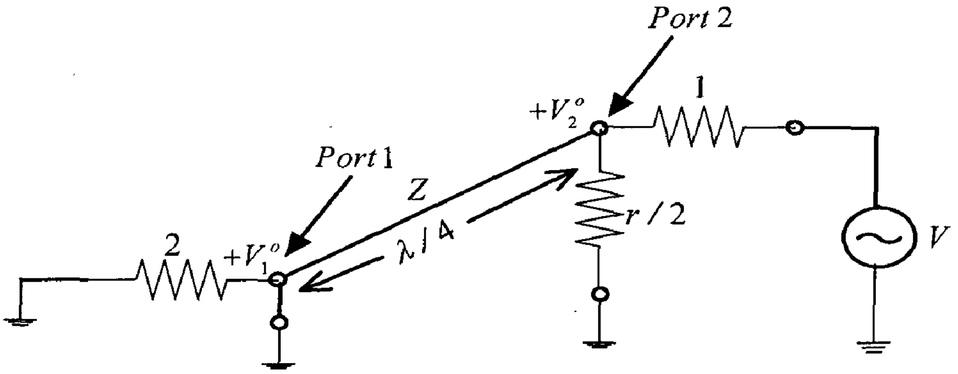


Fig. 3.12 Odd mode excitation

To practically built powder combiner at 5 GHz on the substrate with substrate thickness 1.6 mm and dielectric constant 4.21, use the formulae for finding width, length of the transmission line and effective dielectric constant of the substrate from the following equations.

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \cdot \frac{1}{\sqrt{1 + \frac{12H}{W}}} \tag{3.12}$$

$$Z_o = \begin{cases} \dfrac{60}{\sqrt{\varepsilon_r}} \ln\left(\dfrac{8H}{W} + \dfrac{W}{4H}\right) & \textbf{for } \dfrac{W}{H} \le 1 \\ \dfrac{120\pi}{\sqrt{\varepsilon_r}\left(\dfrac{W}{H} + 1.393 + 0.667 \ln\left(\dfrac{W}{H} + 1.444\right)\right)} & \textbf{for } \dfrac{W}{H} > 1 \end{cases} \tag{3.13}$$

From Electrical length (βl) in degrees to physical length (l) in mm we have

$$\beta l = \frac{2\pi}{\lambda_g}, \quad \lambda_g = \frac{c}{f\sqrt{\varepsilon_{eff}}} \tag{3.14}$$

Table 4.3 Dimensions of Microstrip line characteristic impedance

	Width (mm)	Length(mm)	Effective Dielectric
50Ω	3.2	8.2	3.24
70.7Ω	1.7	8.45	3.1

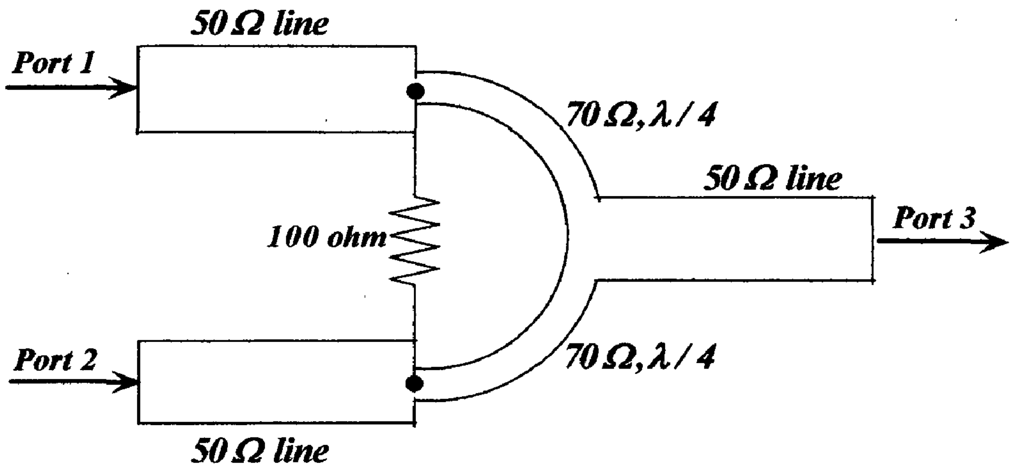


Fig. 3.13 The top view of the layout for the Wilkinson Power Combiner Simulated in ADS

Analyzing of the Wilkinson power combiner was done using Agilent Advance Design System (ADS) [20] tool which is an EDA (Electronic Design and Automation) tool for RF/MW designers, and the results are shown in Fig. 3.15. The power which was fed at ports 1 and 2 appear at port 3 at the center frequency of 5 GHz. There is enough isolation of more than 40dB between the input ports because of resistor. The RF and LO signals fed to the port 1 and 2 will combine and appear at the port 3 of the Wilkinson Power Combiner. Fig. 3.14 shows the combiner simulation in ADS tool.

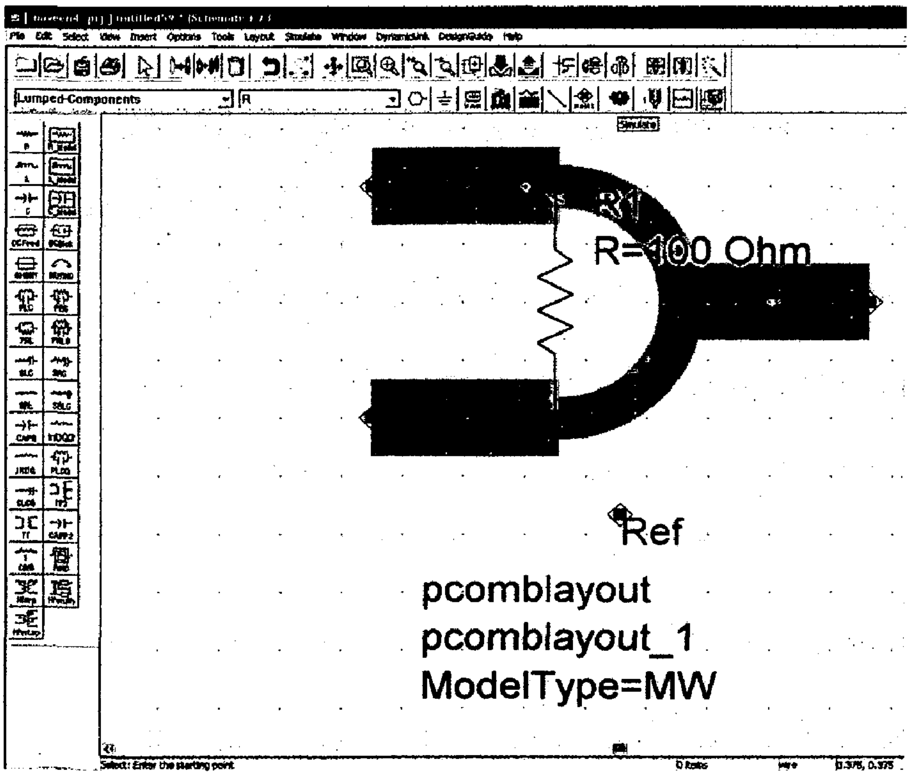


Fig. 3.14 Wilkinson Power Divider with ADS software

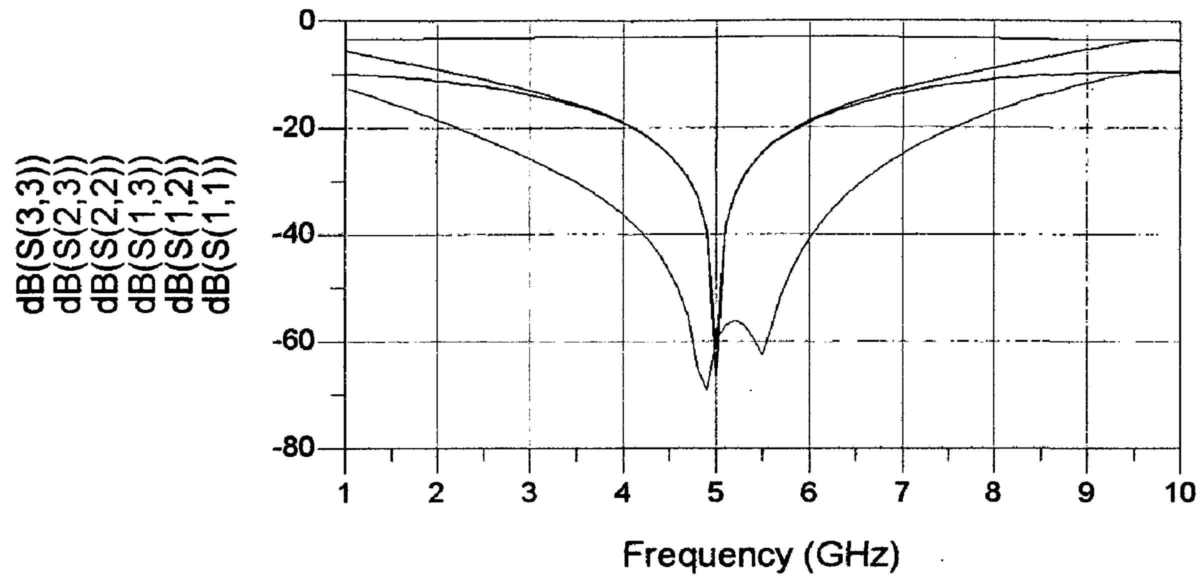


Fig. 3.15 Results from the Data Display window of ADS, illustrating the S-Parameters of the Wilkinson Power Combiner.

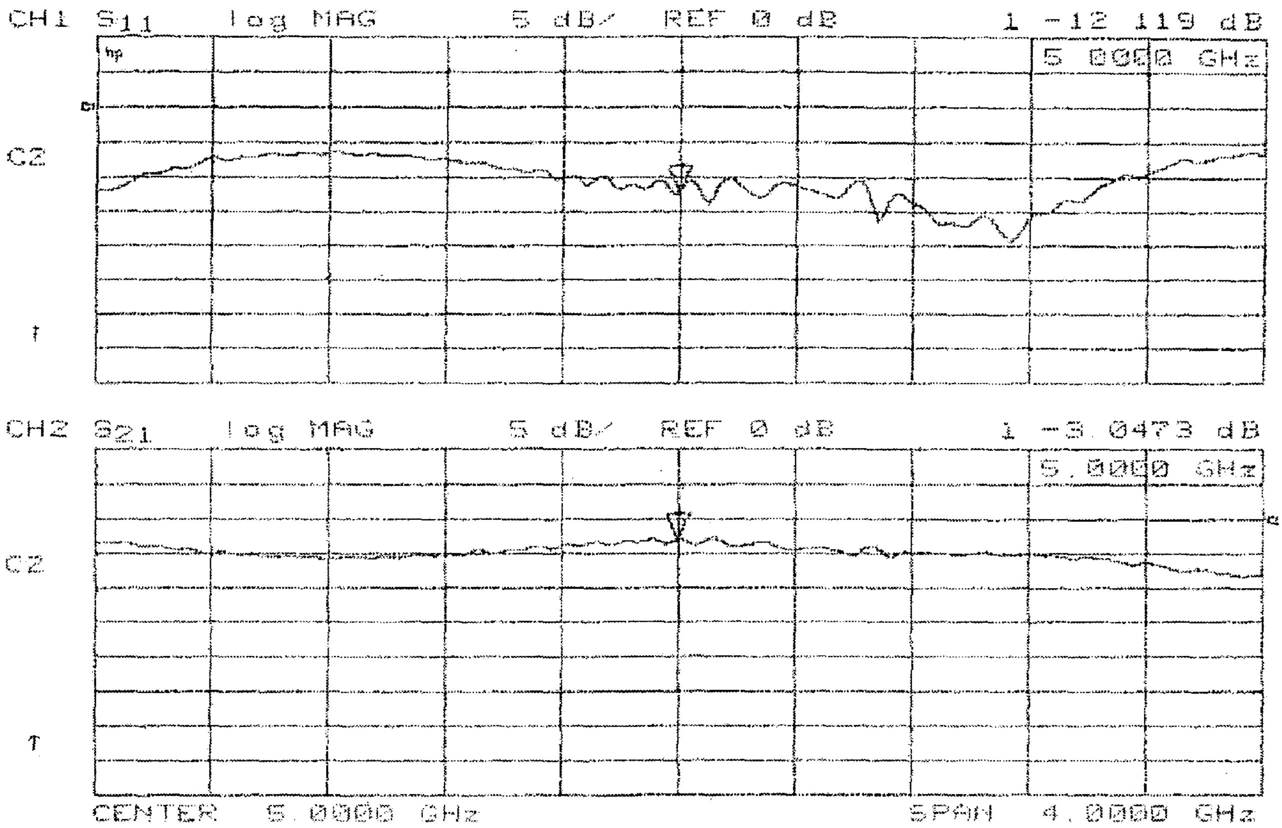
From the S-parameters in Fig. 3.15 it can observe that the power applied at port 1 and port 2 combines appears at the output with a loss in the actual signal power. Since the circuit is lossy, remaining power is dissipated in the 100 ohm resistor. The only disadvantage of Wilkinson Power divider/combiner limited bandwidth [21].

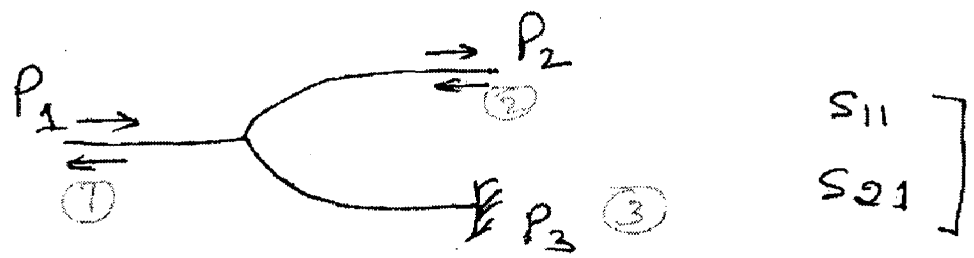
Wilkinson power combiner is implemented and tested on HP-VNA. Fig. 3.16 (a), (b), (c) shown is the S-parameters of the three port circuit. There is a power loss as shown in Fig. 3.16 which is dissipated in the 100 ohm resistor in the circuit. The power loss is 3 dB and the isolation is 15 dB between the input ports the power combiner according to measurements made on HP-Network analyzer. During the measurements, since our Network analyzer is two port and the power combiner is three port circuit, a matched load is connected to the one of the port of the power combiner.

3.4.2 Impedance Matching and Bias Network Design:

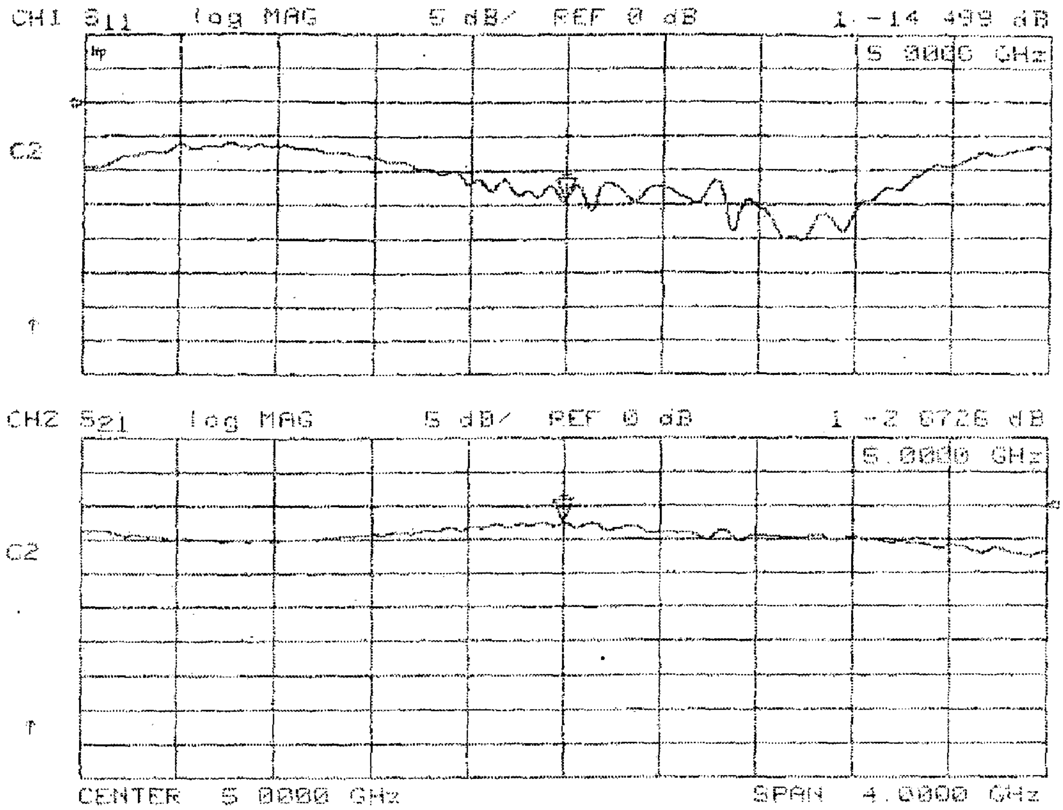
In order to design bias circuit and impedance matching circuits, first requirement is to select the proper non-linear device for mixing action. Virtually all RF and microwave electronics circuits use one or more of three general types of devices: Schottky-barrier diodes, junction transistors, or the field-effect transistors (FETs). Optimized for either low cost on high performance; BJTs; HBTs; and various types of FETs, including metal-epitaxial semiconductor FETs

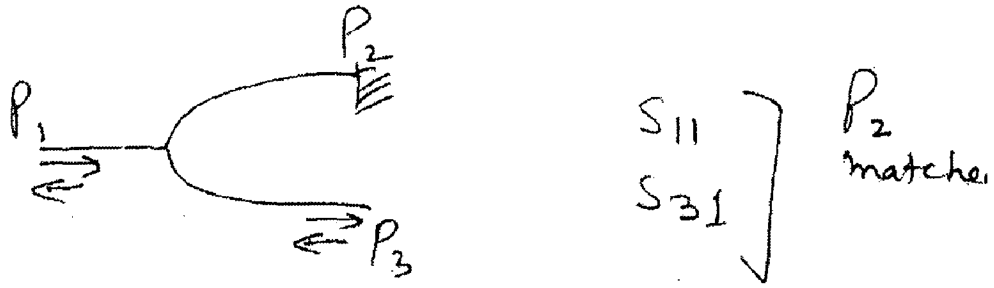
Measurement Results of Power Combiner:





(a)


CH1 S11 log MAG 5 dB/ REF 0 dB 1 -14 499 dB
5 0000 GHz
C2
CH2 S21 log MAG 5 dB/ REF 0 dB 1 -2 6726 dB
5.0000 GHz
C2
CENTER 5 0000 GHz
SPAN 4.0000 GHz


P1
P2
P3
S11
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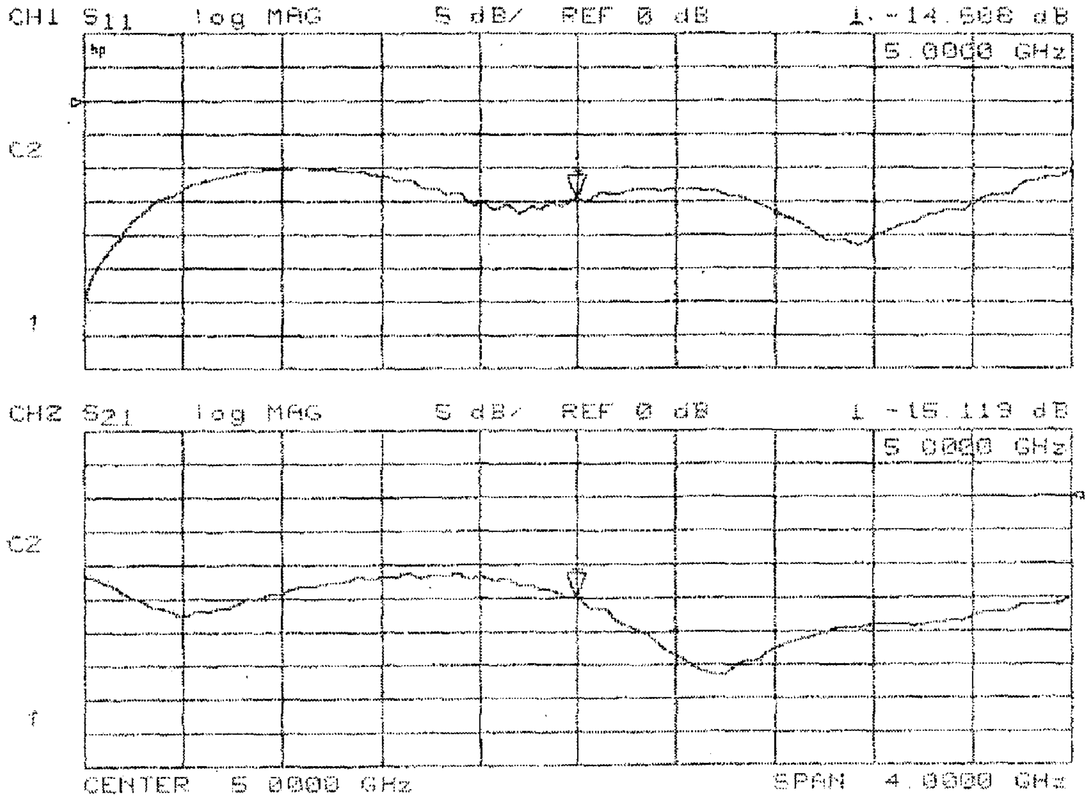
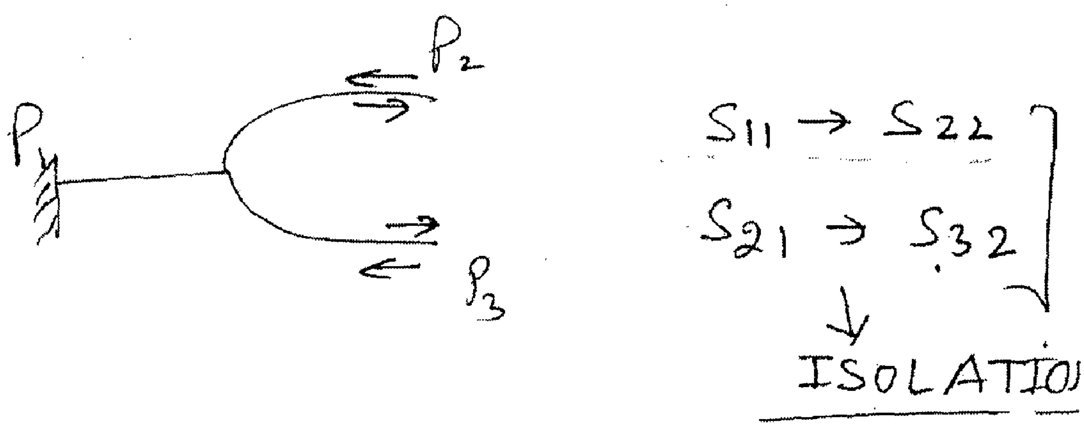
(b)





(c)

Fig. 3.16 Measurement Results of Power Combiner on HP-NA, (a),(b) Transmission coefficient and Reflection coefficient between ports , (c) Input Port isolation.

(MESFETs), high-electron-mobility transistors (HEMTs), metal-oxide semiconductor FETs (MOSFETs), and junction FETs (JFETs).

These devices have distinctly different characteristics, so in most cases the appropriate device for a particular circuit is obvious. The choice of a device also may be colored by the available technologies and, above all, cost.

Table 3.1 suggests the solid state devices description. The suggestions for applications and frequency ranges are weak; often there are good reasons to use a device outside its optimum frequency range or for applications where it might not, at first inspection, seem appropriate.

In the present mixer design, JFET is chosen as non-linear mixing device because of low cost, low noise, well suited for applications in amplifiers, mixers, oscillators and switches as described in Table 3.1. NECs NE4210S01 [22], a HJFET searched in the web which has excellent low noise figure, high associated gain and make suitable for Commercial systems used in our present mixer design. Frequency range of this transistor is up to 26 GHz.

Table 3.1 Non-Linear Devices Description

Device	**Frequency range**	**Uses and Characteristics**
Schottky-barrier diode	The RF to submilli-meter wave region.	Mixers, modulators, and detectors; occasionally used for frequency multipliers and switches.
BJT	Usually X band and below; millimeter-wave BJTs have been made.	Small-signal amplifiers; not low noise. Fast digital circuits. Good power devices. Low 1/f noise makes them ideal for low-noise oscillators
HBT	Some types may have gain at 60 GHz; practical limits are around 30 GHz	Power amplifiers, low-noise oscillators. Fast analog circuits, MESFETs and HEMTs have much better noise figure, but HBTs have lower 1/f noise, making them preferable for oscillators

JFET	Up to 30GHz	Low-cost, moderately, low-noise applications in amplifiers, mixers, oscillators, and switches
MESFET	Microwave workhorse up to 26GHz. Can go higher, but HEMTs are usually preferred.	Amplifiers, oscillators, mixers, modulators, frequency multipliers, control components; in short, everything. Bipolar devices have lower 1/f noise and are preferred for oscillators
HEMT	Highest frequency device available; over 200 GHz.	Much the same as MESFETS; best suited for small-signal, low-noise uses, but power devices are possible
MOSFET	Up to -10 GHz for advanced technologies; 5-7 GHz more common	Analog, digital, and RF Si IC applications. MESFETs and HEMTs have much lower noise figures at microwave frequencies

3.4.2.1 Input (RF) Impedance Matching Network:

Matching and microwave circuit design are to some designers synonymous. It is the act of making the source and load impedances matched to achieve the desired amount of power reflected and power transferred. Matching is required if the circuit is to yield optimum gain and return loss. Poorly matched devices can cause large amount of reflected power, poor noise performance, and low gain. Poor input match can also reduce the gain of the mixer and cause the system to have non-optimum noise performance.

Two input matching circuits have been used at the gate and drain of the transistor. Stub impedance matching circuits at the input and output ports generate a very good signal sensitivity and a low noise figure in the RF, LO and IF ports. In addition, the matching circuit has been designed to have a good isolation among RF, LO and IF ports. The input impedance stub matching circuit matches the input reflection coefficient of the HJFET (NE4210S01) to 50 ohm of the RF input port.

To maximize conversion efficiency, gate is conjugate-matched at the RF. There appears to be little value in mismatching the input to improve the noise figure of an active mixer. There are a couple of good reasons for this. First, in an amplifier, the input mismatch is a trade-off between the need for gain, which dictates a conjugate input match, and low noise, in which the source impedance is chosen to force the noise currents to cancel. The gain of an FET mixer inherently is much lower than an amplifier, so the trade-off of noise against gain must be biased more toward gain.

The conversion gain of the mixer will be the maximum value when the source and load impedances is matched to the complex conjugate impedance of input reflection coefficient and output refection coefficient respectively. Ideally, the input matching circuit should match the input impedance of the FET at both the RF and the LO frequencies. In many cases, however, the RF and LO frequencies are significantly different, and it is impossible to match the device successfully at both frequencies. When this conflict exists, it is better to match the device at the RF frequency and to accept a mismatch at the LO frequency. A poor RF match degrades conversion performance, but the only consequence of a poor LO match is to waste a little LO power.

Because the gate circuit of a FET is not strongly nonlinear (as long as it is driven into conduction), the input impedance of a pumped FET can be estimated from S-parameters [23]. It simply is the gate-input impedance of the device, biased near pinch-off, with a short-circuit on the drain. So the matching circuits can be designed at 5 GHz from the following S-parameters of HJFET (NE4210S01) (these S-parameters are obtained from the data sheet)

$$S_{11} = 0.825\angle -58.15^{o}$$
$$S_{21} = 4.565\angle 112.41^{o}$$
$$S_{12} = 0.055\angle 48.65^{o}$$
$$S_{22} = 0.549\angle -42.36^{o}$$

For maximum gain, we should design the matching sections for a conjugate match of the transistor. Thus $\Gamma_S = \Gamma_{in}^*$ and $\Gamma_L = \Gamma_{out}^*$ can be determined by

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1^2|}}{2C_1} \tag{3.15}$$

$$\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2^2|}}{2C_2} \tag{3.16}$$

where the variable B_1, C_1, B_2, C_2 are defined as

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2, \tag{3.17}$$

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2, \tag{3.18}$$

$$C_1 = S_{11} - \Delta S_{22}^*, \tag{3.19}$$

$$C_2 = S_{22} - \Delta S_{11}^*. \tag{3.20}$$

$$\Delta = S_{11}S_{22} - S_{12}S_{21}. \tag{3.21}$$

By above Eq. (3.15) - (3.21) we get

$$\Gamma_S = 0.98\angle 91.67^0$$

$$\Gamma_L = 0.985\angle 46.64^0.$$

The matching networks can easily be determined using the smith chart. For the input matching section, we plot Γ_S. The impedance, Z_S, represented by this reflection coefficient is the impedance seen looking into the matching section toward the source impedance, Z_0. Thus, the matching section must transform Z_0 to the impedance Z_S. There are several ways of doing this, but we will use a shunt stub followed by a length of line. Thus we convert to the normalized admittance y_S and work backward to find the length of the line. Similar procedure gives the length of the line and stub length for the output matching circuit.

Impedance matching circuits have been used at the gate of the transistor. The input impedance stub matching circuit matches Γ_{in} to 50 ohm of the RF input port and output impedance stub matching circuit matches Γ_{out} to 50 ohm of the LPF input port. The conversion gain of the mixer will be the maximum value when the source impedance is matched to the impedance of Γ_{in}^{*} and output impedance is matched to the impedance of Γ_{out}^{*}.

We consider a matching technique [24] that uses a single open-circuited length of transmission line (a "stub"), connected in parallel with the transmission feed line at a certain distance from the load. The shunt tuning stub is especially easy to fabricate in microstrip form. The proper length of open or shorted transmission line can provide any desired value of reactance or susceptance. For transmission line media such as microstrip or stripline, open-circuited stubs are easier to fabricate since a via hole through the substrate to the ground plane is not needed. For finding the length of the stub and the distance to place the stub, we have a well known procedure is by Smith chart. This is a simple job; from the smith chart following length of the stub and line are obtained as shown in Fig.3.17. Open circuited Stub and transmission line length calculation is done using smith chart is shown in Fig. 3.18.

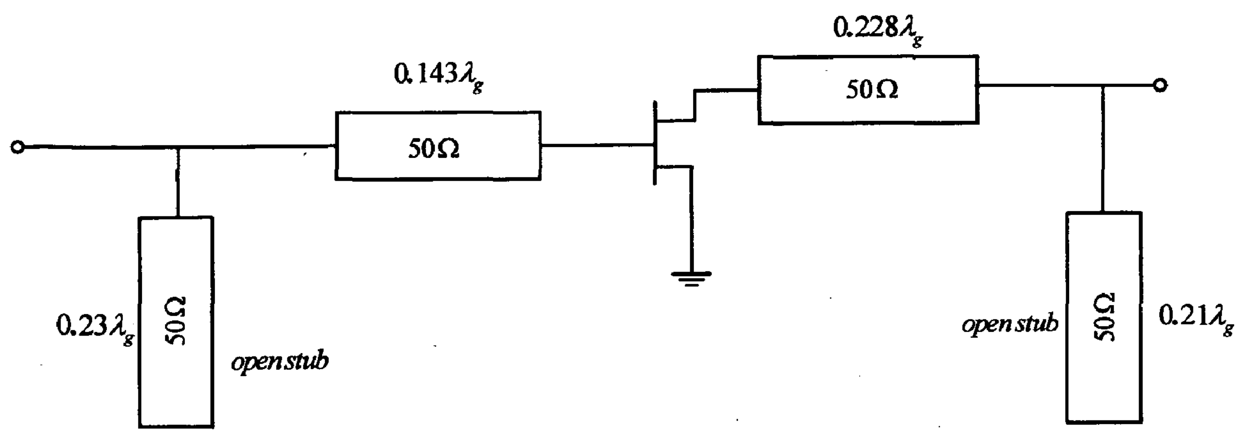


Fig. 3.17 RF Circuit

where $\lambda_g = \dfrac{c}{f\sqrt{\varepsilon_{eff}}}$, f = 5 GHz, from Eq.(4.2) and (4.3) we get the widths , lengths of the line and open circuit stub.

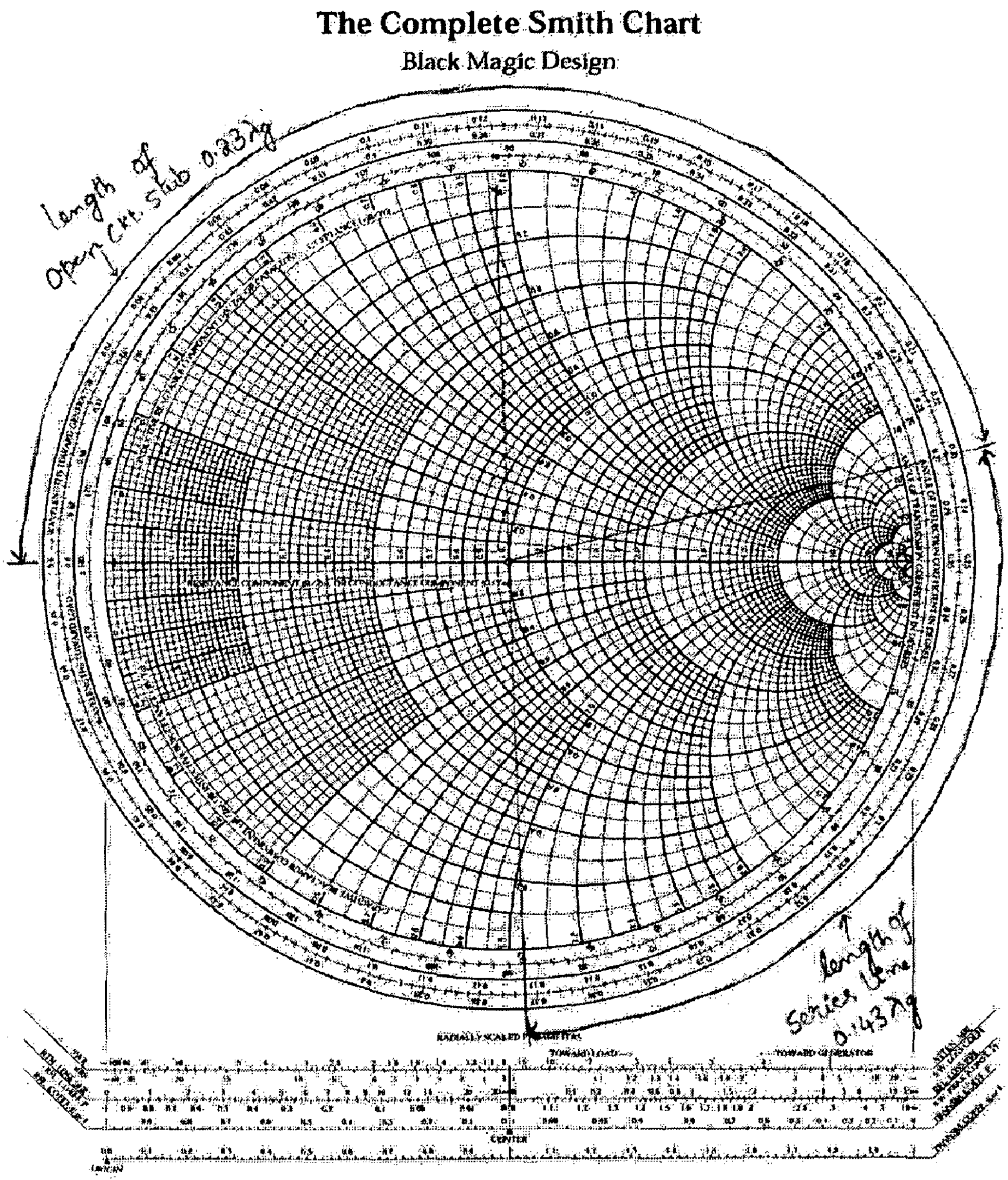


Fig. 3.18 Smith chart for the design of input matching network

The reflection coefficient matching section is shown in Fig.3.19.

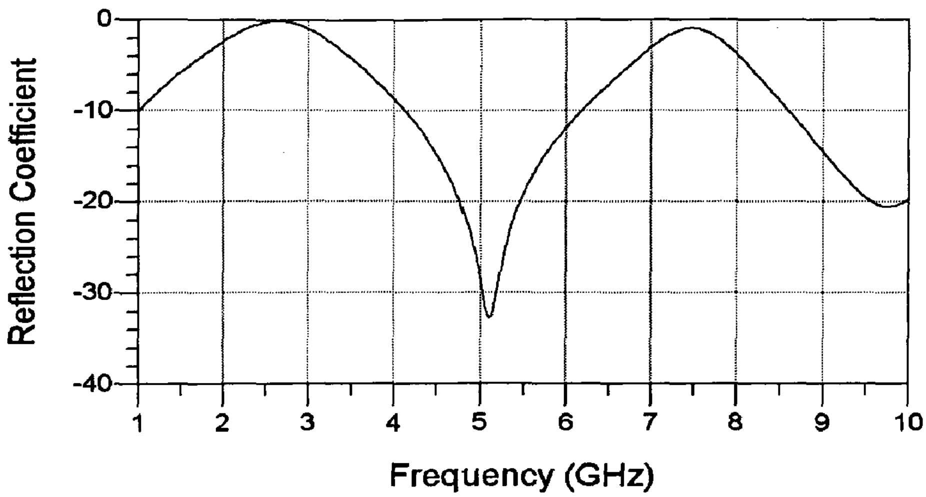


Fig. 3.19 *Reflection Coefficient vs. Frequency of the stub*

This open stub matches 50Ω to the input impedance of the FET at 5 GHz which is RF frequency , as already mentioned that we are trying to match at RF frequency only since matching at both RF and LO frequencies is not possible as long as both RF and LO frequencies are very close to each other.

3.4.2.2 Bias Network Design:

In all FETs, including HEMTs, the transconducatance is maximum when the device is operated in current saturation. In this case the transconductance can be varied from zero (at pinch-off) to some maximum value. In MESFETs and other junction FETs, the maximum transconductance occurs at maximum channel current. Clearly, the LO must vary the transconductance between these two extremes, Fig. 3.20 tells at what options are there for biasing.

The JFET is usually (but by no means exclusively) operated in saturation region as discussed earlier. When gate-to-source voltage $V_{gs} \leq V_p$, gate pinch-off voltage, the device is saturated and the current is almost constant with increase in V_d, drain-to-source voltage. The drain current of JFET is best expressed as

$$I_d\left(V_g,V_d\right)=I_{dss}\left(1-\frac{V_g}{V_p}\right)^2 \tag{3.22}$$

where I_{dss} is the drain current at $V_g = 0$. The square-law *I/V* characteristic described by Eq. 4.15 probably should not be taken too linearly, Eq. 3.22 implies that JFET has little or no third-order distortion and thus it can generate new frequency components.

If the transconductance is idealized as a linear function of gate voltage, and enough LO power is available to drive the FET to maximum transconductance, biasing the FET at pinch-off provides the greatest $G_m(t)$, where $G_m(t)$ is the transconductance wave form. In real devices, the optimum gate bias may not be precisely at pinch-off, but will be close.

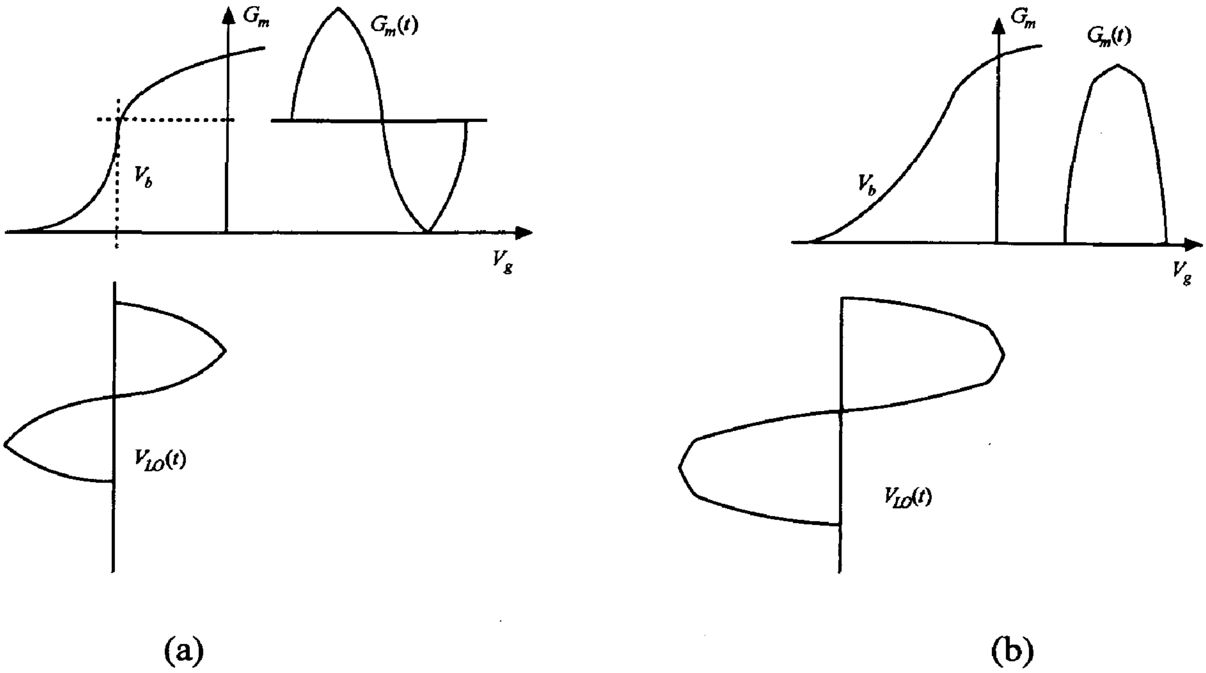


Fig. 3.20 For high gain and low noise, the gate bias must be chosen to provide the greatest fundamental-frequency component of the transconductance waveform. In (a), the FET is biased in the active region; in (b) near pinch-off. The latter provides optimum performance.

At this point, we have several rules for operating transconductance mixers:

1. Bias the device at pinch-off.
2. Drive the gate with LO power so the peak gate voltage provides maximum tranconductance.
3. Keep the FET in current saturation throughout the LO cycle.

The methods for achieving these goals should be clear:

1. Conjugate-match the gate at the LO frequency.
2. Provide enough LO power.
3. Short-circuit the drain at all LO harmonics, including the fundamental. This will "pin" the drain voltage at its dc value, keeping the device current saturation throughout the LO cycle. It also will prevent LO leakage into the IF circuit.

RF/Microwave transistors/ FET require some form of circuit to set correct bias conditions for a particular RF performance. Bias network design for the mixer is dependent on the frequency range in which mixer needs to be designed, that is to say if mixer needs to be design at low frequency application then a choke (inductor) is used but getting discrete inductors at microwave frequencies is difficult. What is required is a low DC resistance but a high resistance to ensure that the RF circuit is not loaded and RF signals do not flow on to the supply lines. One probable and commonly used method is to place low impedance stub immediately after $\lambda/4$ high impedance bias line at frequency of RF will achieve proper isolation at desired RF frequency, no matter what component added after $\lambda/4$ long bias line. Fig. 3.21 shows the circuit design for bias circuit where if could be seen that high impedance $\lambda/4$ bias line immediately followed by open circuit stub ($\lambda/4$ long @ RF) and the stub impedance should be as low as possible. The impedances here for the bias line are chosen within constraints of size and technology. A quarter wave high impedance short-circuited stub at RF frequency to short circuit the IF at the input is also used for the gate bias circuit.

Low impedance is chosen as 25 Ω and high impedance as 100 Ω for this bias network design. A chip capacitor is connected between stub and ground. It should be large enough to short circuit the gate over the entire IF band. A resistor (on the order of 1000Ω) and large-value capacitor complete the bias circuit. These provide additional filtering and reduce the danger of gate damage from electrical transients. We have chosen resistor and capacitor of 2.2kΩ and 10pf farads respectively.

Same bias network was used at the drain side and the bias voltages which should be applied to the transistor at gate and drain side was supplied through this network. The bias voltages obtained for the mixer design was V_{gs} = - 0.4V and V_{ds} = 3V so that the mixer operates in the current saturation region.

Based on the frequency range and gain requirement, the NE4210S01 HJFET device was selected for the present mixer design. The first analysis that needs to be performed is the DC simulation to find out the right biasing points for the mixer. The datasheet of the device provides the DC analysis. Biasing point is found through simulation. Based on the analysis bias points are selected as V_{gs}= - 0.4V and V_{ds}= 3V, which provides the drain current of 30mA.

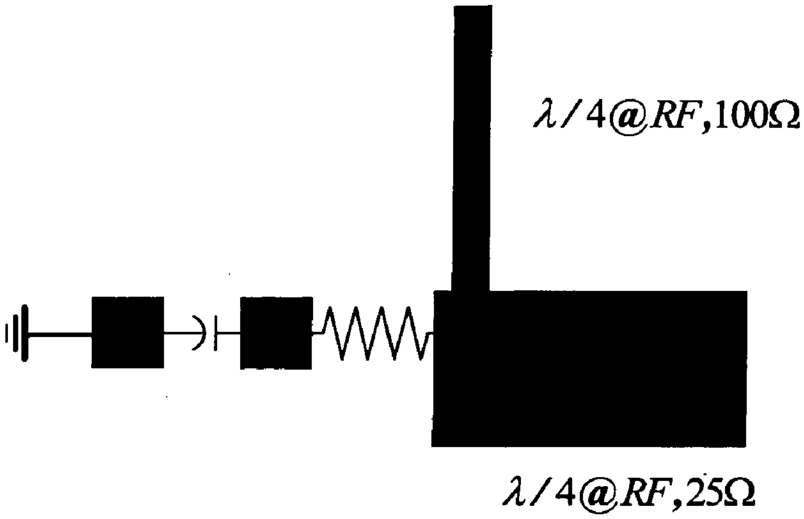


Fig. 3.21 Layout of Bias Network

3.4.3 RF Performance:

A single-stage microwave transistor amplifier can be modeled by the circuit of Fig. 3.22, with a matching network on the both sides of the transistor to transform input and output impedance to the source and load impedance respectively; the matching network is shown in Fig. 3.17. The performance of this RF circuit (amplifier) is observed by simulating it in ADS with gate and drain bias as shown in Fig. 3.23. The S-parameters of the circuit, mainly S_{21} in dB, represent gain of the amplifier.

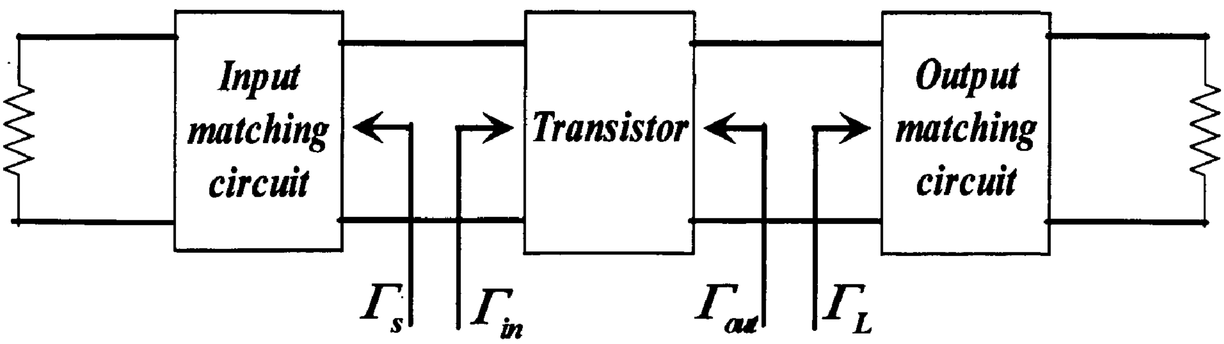


Fig. 3.22 The general transistor amplifier ckt

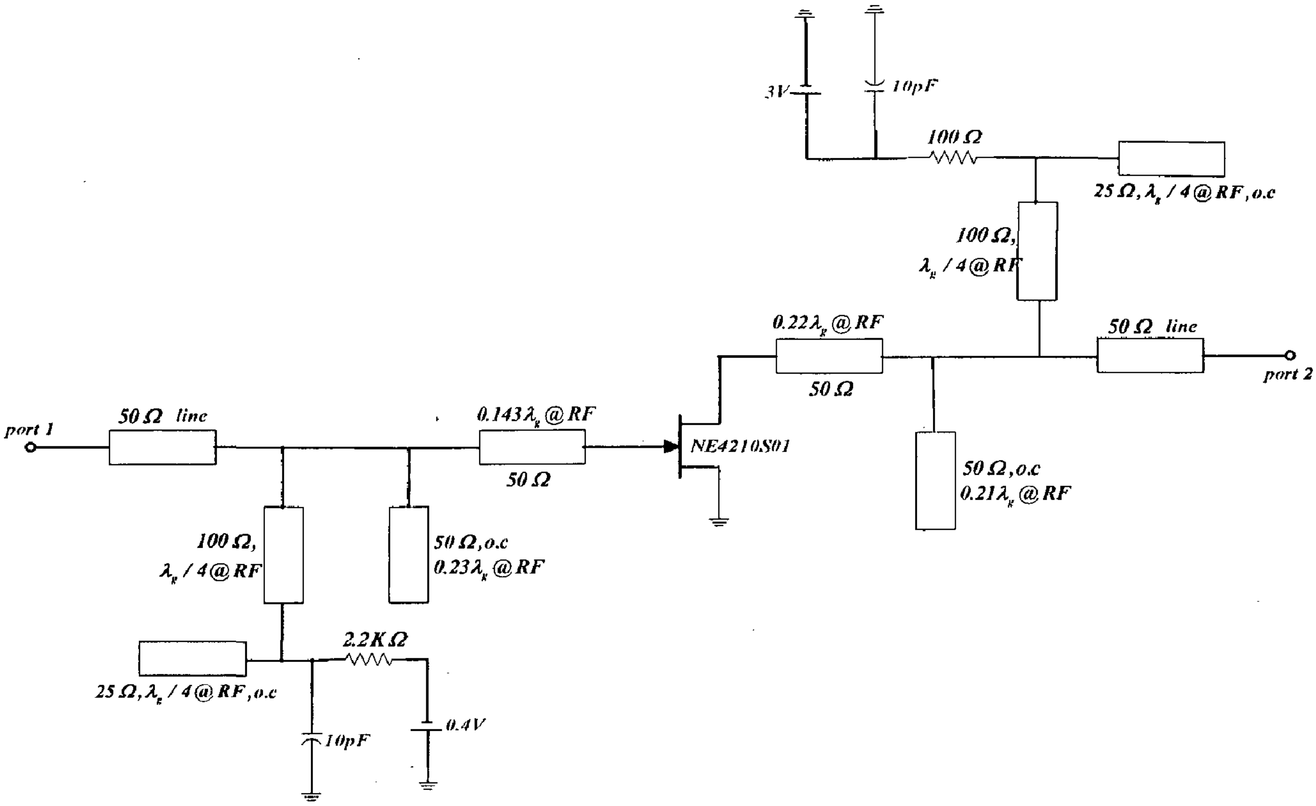


Fig. 3.23 Amplifier (RF circuit) Model simulated in ADS

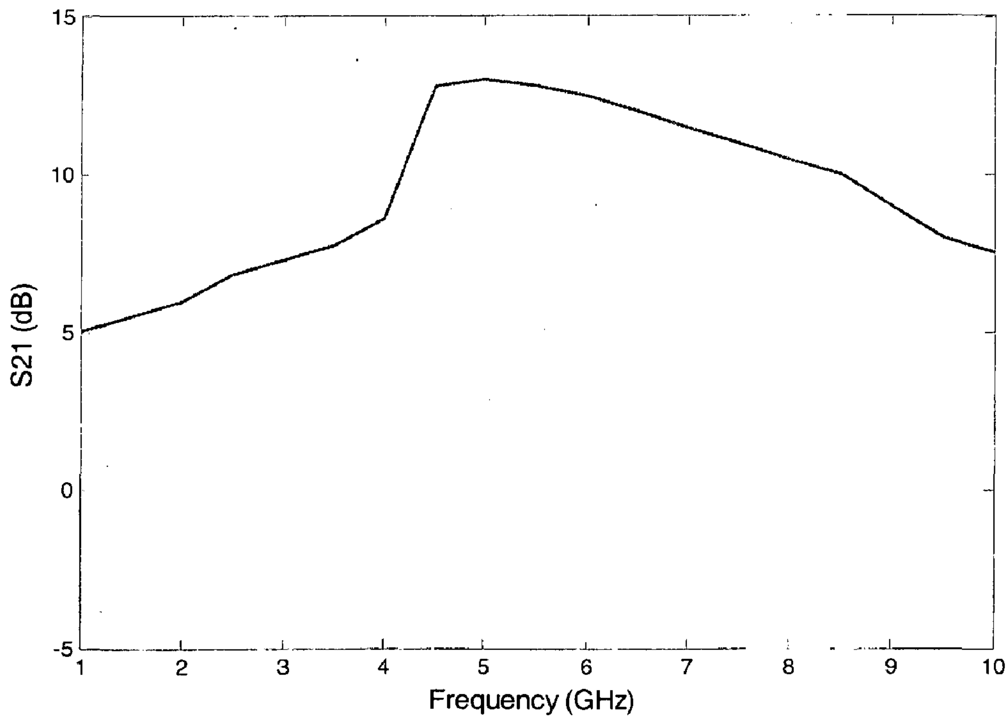


Fig. 3.25 RF Characteristics (Gain Response)

The response of the circuit is shown in Fig.3.25. The amplifier produces maximum gain of 13 db at 5 GHz, since the stub and line lengths are calculated at 5 GHz. With this performance we are able to state that matching was done. Thus the

The response of the circuit is shown in Fig.3.25. The amplifier produces maximum gain of 13 db at 5 GHz, since the stub and line lengths are calculated at 5 GHz. With this performance we are able to state that matching was done. Thus the design part of RF section was done. We move further to the design and analysis of IF section.

3.5 IF Section:

The output circuit needs to provide a path for the IF component of the drain current to the load, while short circuiting the RF and LO frequencies. Here, this function is provided by an IF filter. The purpose to introduce the filter here in this design is to suppress all other harmonics of LO, RF and image frequency. As the FET mixer will also amplify the LO signal because of insufficient isolation over the ports. One of the common ways to design a filter section is to use a low pass filter.

A good approach is to use a filter structure that effectively filters or attenuates the RF and LO bands. We select a filter consisting of a series of alternating low (Z_L) and high (Z_H) impedance transmission lines sections, usually called as stepped impedance low pass filter, which is popular because it is easy to design and takes less space than similar low-pass filters using stubs. However due to the approximation involved, the performance is not as good and it is limited to application where a sharp cutoff is not required.

3.5.1 LPF Design:

The design [25] starts with choosing high and low impedances values, the cutoff frequency and necessary attenuation at particular frequency are mandatory to design. The ratio of Z_H/Z_L should be as high as possible, limited by the practical values that can be fabricated on a printed circuit board. Typical values are Z_H =100 to 150Ω and Z_L=10 to 15Ω. Cutoff frequency of 2 GHz and it is necessary to have more than 20 dB insertion loss at 3 GHz, the filter impedance is 50Ω. First lumped Model of the filter was designed.

Here f_c= 2 GHz and f = 3 GHz, calculating the ratio $\left|\frac{\omega}{\omega_c}\right| - 1 = 0.5$, then from Fig. 3.26 indicates N=5 (5 lumped elements) give necessary attenuation at 3GHz.

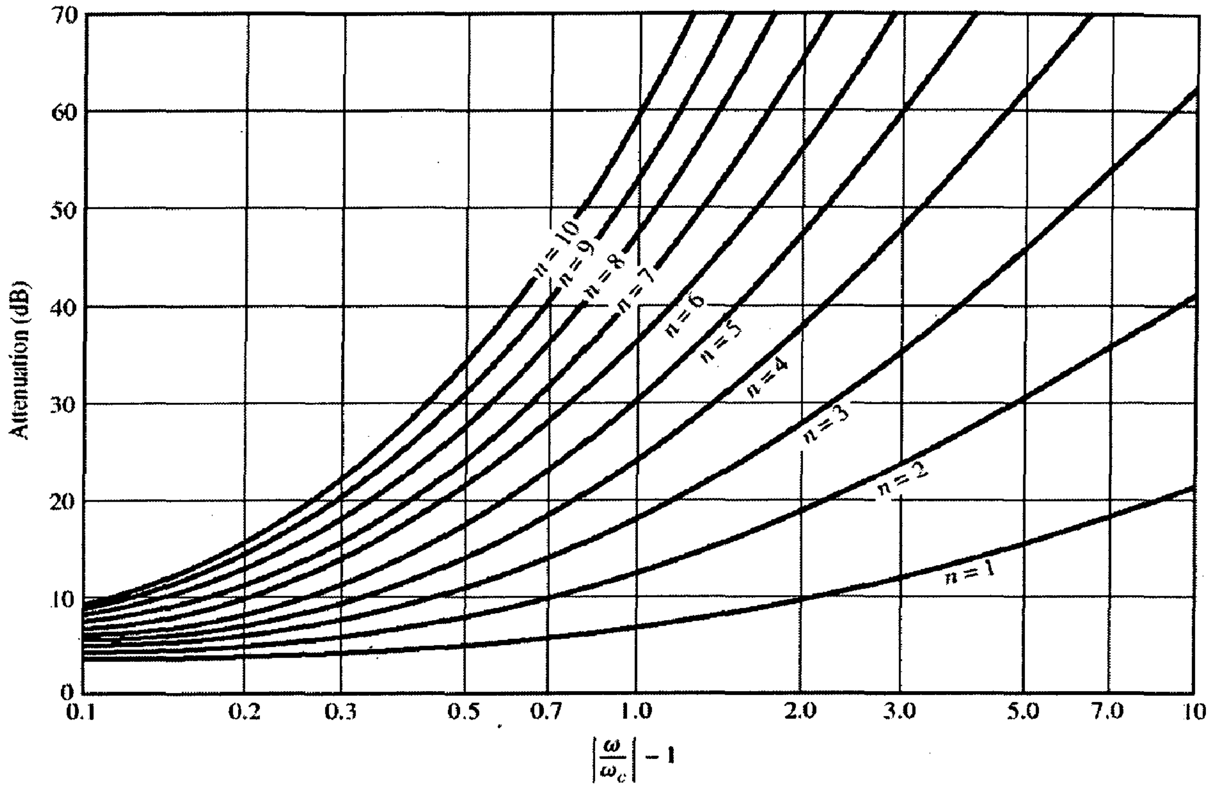


Fig. 3.26 Attenuation vs normalized frequency for maximally flat filter prototypes. Adapted from G.L. Matthaei, L. Young, and E.M.T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*.

The table 3.1 gives the low-pass prototype values for the given specifications of the filter are

$$g_1 = 0.6180 = C_1,$$

$$g_2 = 1.6180 = L_2,$$

$$g_3 = 2 = C_3,$$

$$g_4 = 1.6180 = L_4,$$

$$g_5 = 0.6180 = C_5$$

Lumped Model of the Filter

The lumped values of the filter are given by:

$$C_k' = \frac{g_k}{R_0 \omega_c} \quad \text{and} \quad L_k' = \frac{R_0}{g_k \omega_c} \quad \text{where } R_0 = 50\Omega \tag{3.23}$$

TABLE 3.1 Element Values for Maximally Flat Low-Pass Filter Prototypes

N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8
1	2.00	1.00						
2	1.414	1.414	1.00					
3	1.00	2.00	1.00	1.00				
4	0.765	1.847	1.847	0.765	1.00			
5	0.6180	1.618	2.00	1.618	0.618	1.00		
6	0.5176	1.4142	1.9318	1.9318	1.4142	0.5176	1.00	
7	0.4450	1.247	1.8019	2.00	1.8019	1.2470	0.4450	1.0

By the above Eq. 3.23 we can calculate the lumped values of the low pass filter; the resulting values are given by

$C_1' = 0.984$ pF, $L_2' = 6.44$ nH, $C_3' = 3.1847$ pF, $L_4' = 6.44$ nH, $C_5' = 0.984$ pF.

The resultant lumped model low pass filter is shown in Fig. 3.27.

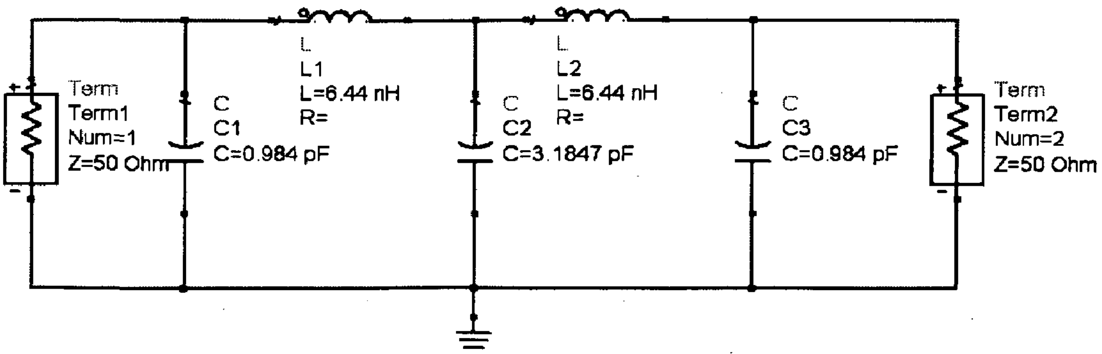


Fig. 3.27 Low-Pass Maximally flat filter Circuit

Frequency response of the above filter circuit is shown Fig. 3.28.

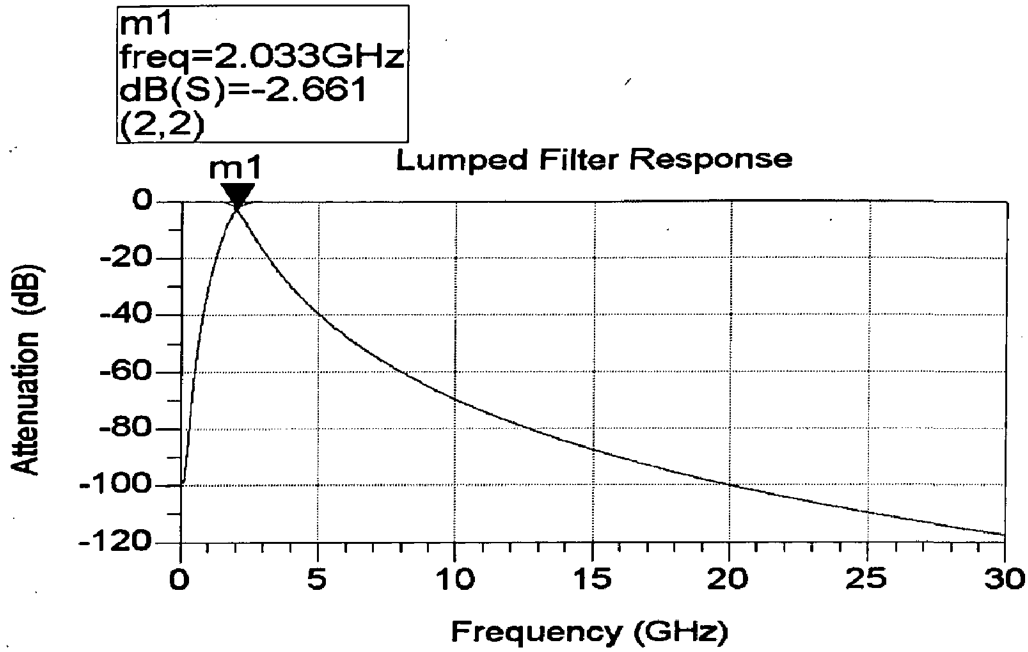


Fig. 3.28 Frequency response of the above filter design

Since a typical low-pass filter consists of alternating series conductors and shunt capacitors in a ladder configuration, we can implement the filter on a printed circuit board by using alternating high and low characteristic impedance section transmission lines.

For distributed design the electrical length is given by:

Length of capacitance section (βL_c) = $C_k Z_L / R_0$

Length of inductance section (βL_i) = $L_k R_0 / Z_H$,

Where Z_L is the low impedance value,

Z_H is the high impedance value,

R_0 is the source and load impedance,

ω_c is the desired cutoff frequency.

Considering Z_L=10Ω and Z_H = 150Ω and then finding the electrical lengths of the hi-Z, low-Z transmission line sections to replace the series inductors and shunt capacitors:

$\beta l_1 = 7.085°$, $\beta l_2 = 30.917°$, $\beta l_3 = 22.93°$, $\beta l_4 = 30.917°$, $\beta l_5 = 7.085°$,

Those are electrical lengths of the transmission lines for the corresponding high and low impedances, since $\beta = 2*\pi / \lambda_g$, where λ_g is the guided wavelength.

A typical substrate PCB with ε_r = 4.21 and H = 1.6mm is used; substituting the corresponding values we find the physical lengths of the lines.

L_{c1} = 1.4914 mm = l_1, L_{i2} = 7.68727 mm,= l_2

L_{c3} = 4.8268 mm,= l_3, L_{c4} = 7.68727 mm,= l_4, L_{c5} = 1.4914 mm.=l_5

The corresponding width of the transmission lines with characteristic impedance of 10 and 150Ω are calculated by below formulae:

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \cdot \frac{1}{\sqrt{1 + \frac{12H}{W}}} \tag{3.24}$$

$$Z_o = \begin{cases} \dfrac{60}{\sqrt{\varepsilon_r}} \ln\left(\dfrac{8H}{W} + \dfrac{W}{4H}\right) & \textbf{for } \dfrac{W}{H} \le 1 \\ \dfrac{120\pi}{\sqrt{\varepsilon_r}\left(\frac{W}{H} + 1.393 + 0.667\ln\left(\frac{W}{H} + 1.444\right)\right)} & \textbf{for } \dfrac{W}{H} > 1 \end{cases} \tag{3.25}$$

TABLE 3.2 *Dimension of microstrip line characteristic impedances.*

	W/H	H/mm	W/mm	ε_{eff}
Z_L = 10Ω	16.2	1.6	25.9136	3.91
Z_O = 50Ω	1.97	1.6	3.1573	3.243
Z_H = 150Ω	0.12	1.6	0.186	2.8

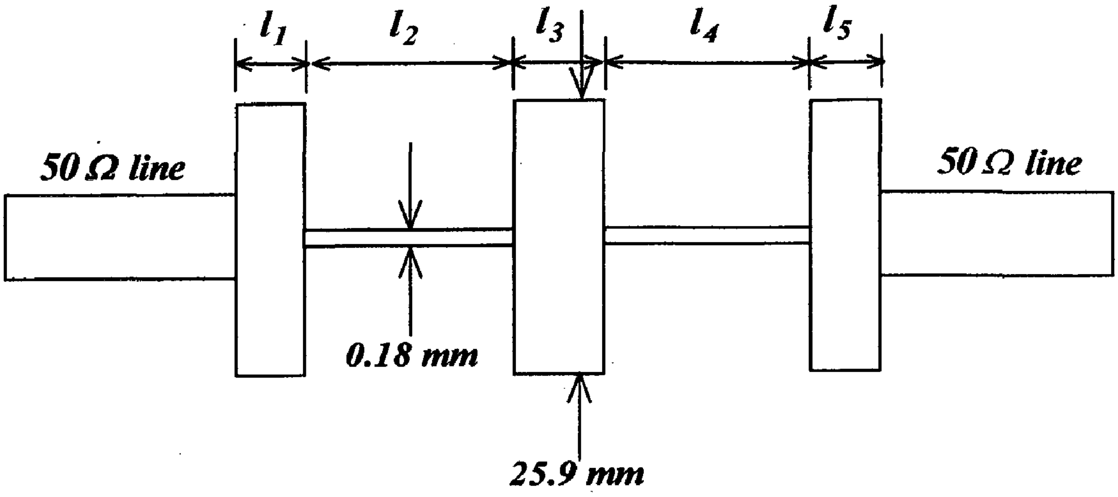


Fig. 3.29 The top view of the layout for the Low pass filter on PCB

The layout of the filter shown in Fig. 3.29 is simulated in ADS tool and the window performing simulation in ADS is shown in Fig. 3.30.

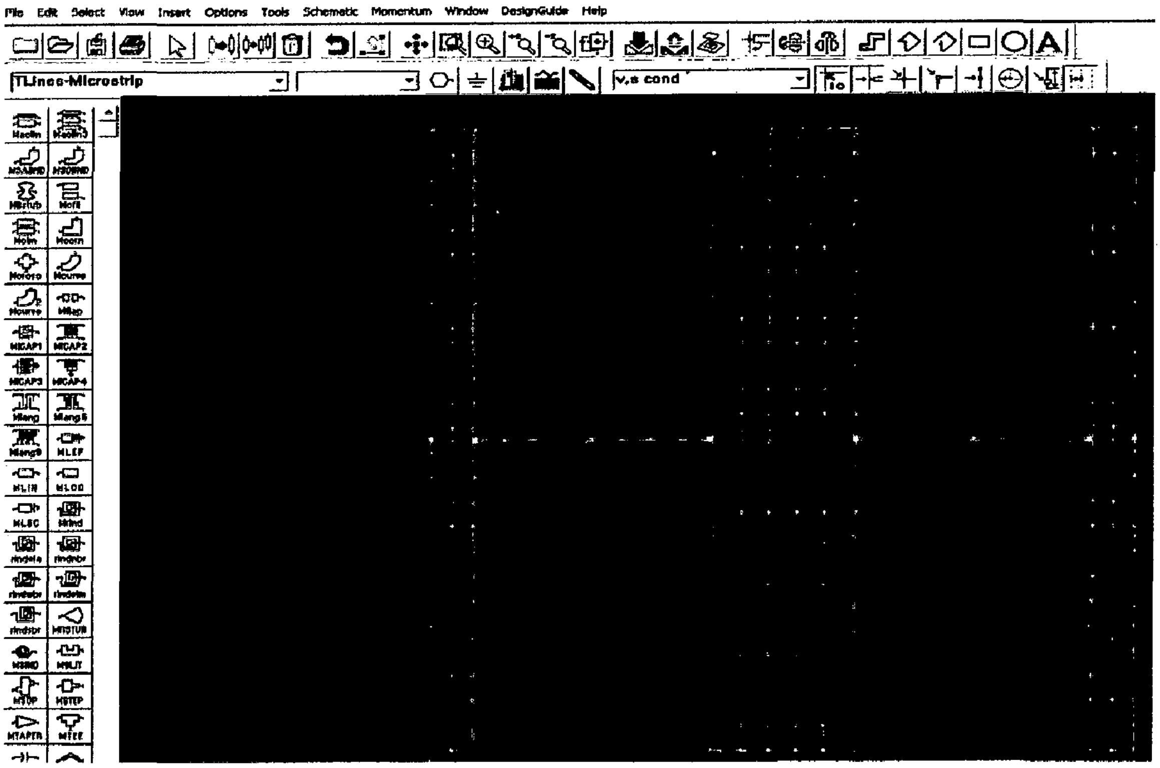


Fig. 3.30 Stepped Impedance LPF simulation with ADS software

Analysis of the step-impedance low pass filter was done using Agilent ADS tool and the results from ADS are shown Fig. 3.31.

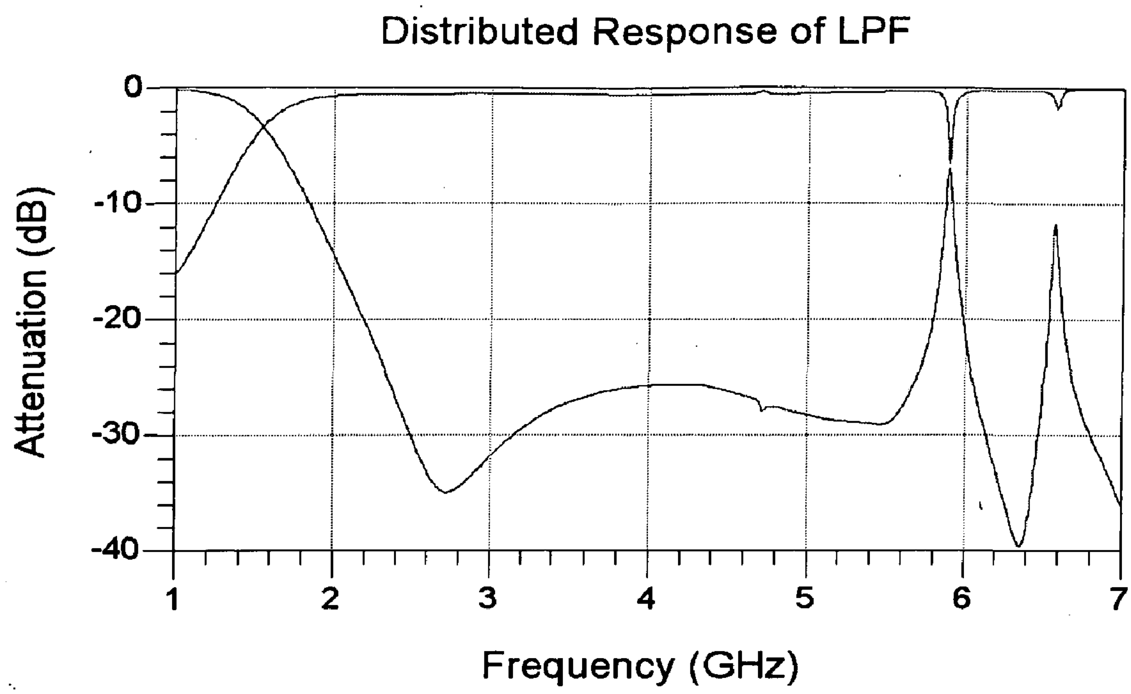


Fig. 3.31 –Result from the Data Display window of ADS, illustrating the S-parameters of the step-impedance low pass filter.

It is essential to fabricate and test this designed Stepped impedance Low pass filter, to know the attenuation in pass band as our mixer IF signal lies in this pass band. The stubs dimension obtained from the mathematical formulae is implemented. The microstrip layout of the LPF is then tested on HP-VNA and the measured response (S_{21}) is shown in Fig. 3.22.

In this chapter we designed the individual blocks of the mixer, simulated in the ADS tool. We have implemented Wilkinson power combiner and LPF too and tested on VNA. These tested results and simulated results are satisfactory and the final mixer was designed. We simulated the amplifier to observe the gain characteristics of the circuit. We have optimized the individual parts of the mixer before performing simulation in EDA tool and it is essential to do this. Then the final mixer was designed and to be simulate in ADS shown in Fig. 3.33.

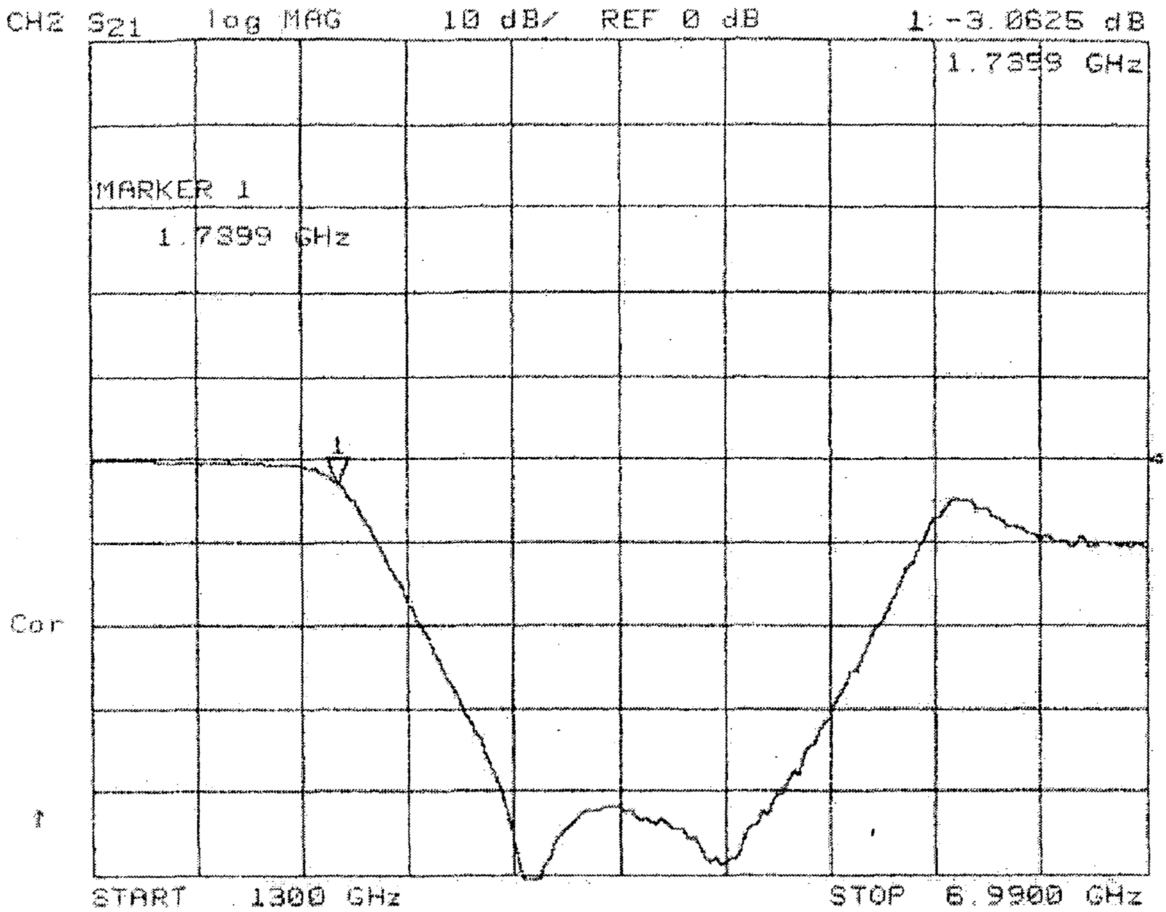


Fig. 3.32. Measured Response on HP-VNA

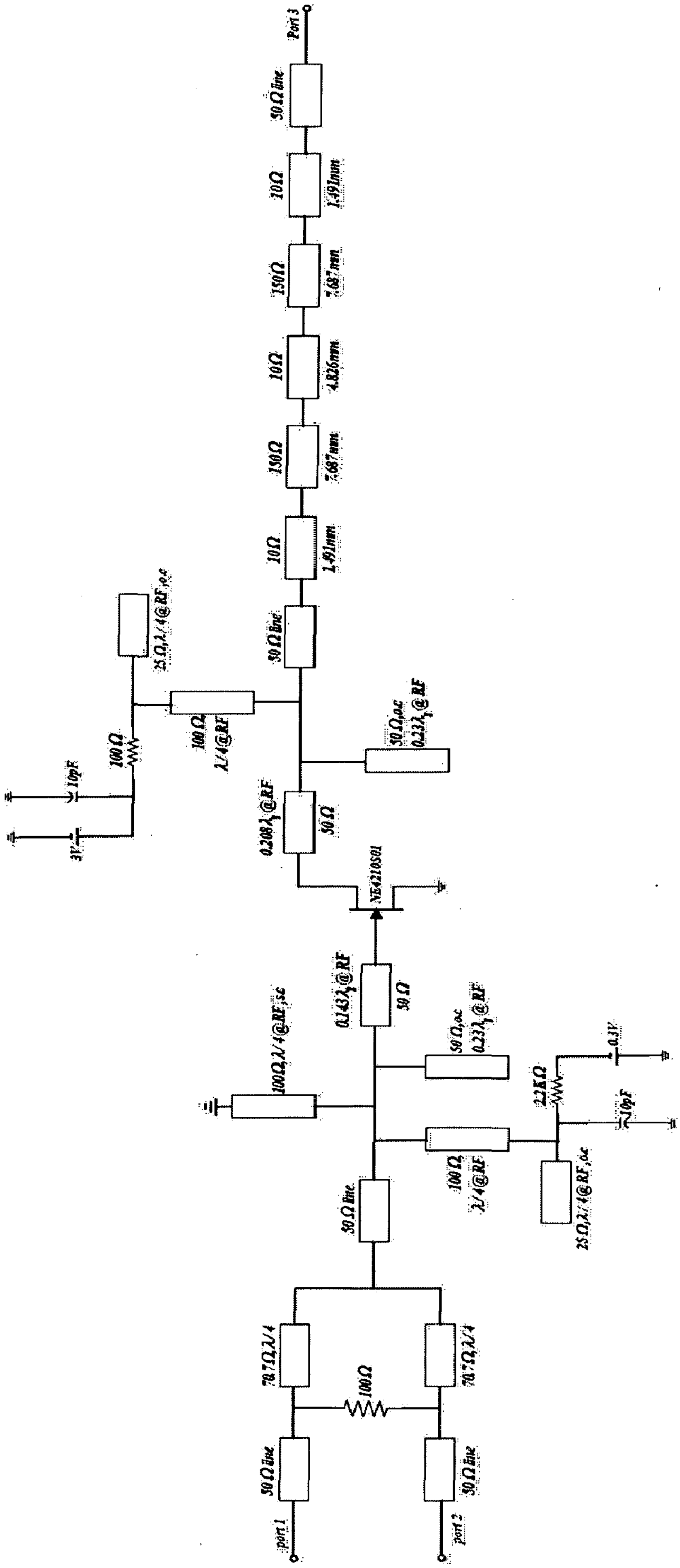


Fig. 3.33 Designed Mixer (Model Simulated in ADS)

4. ANALYSIS AND DEVELOPMENT

The history of circuit simulation for RF design [26] has meandered back and forth between improvements to device models and improvements to the circuit simulators themselves. In the early 1970s, SPICE was pioneered and prompted a range of device modeling efforts. It was not until the late 1980s and the commercialization of harmonic balance simulators for PCs once again accelerated, particularly for microwave monolithic integrated circuits and GaAs technology. Now, in the early 2000s, PC simulation technique matured to the point where nonlinear characterization of circuits can be performed with relative ease.

This chapter we discuss the harmonic-Balance analysis technique and mixer implementation. Harmonic Balance analysis is most useful for strongly and weakly nonlinear circuits that have single or multitone excitation. Harmonic Balance calculates the circuit's steady-state response.

4.1 Mixer Analysis:

In principle, mixers can be analyzed by any nonlinear method that can manage two tones as input signals. Time domain direct integration and harmonic/spectral balance methods are suitable algorithms, since they can handle very strong nonlinearities and two-tone analysis. Active mixers do not demand highly accurate device models; any model that is reasonably representative of the device is adequate

Analysis sometimes called simulation is used to find the response of a known circuit for a specified excitation. The designed mixer is simulated with the help of harmonic balance simulation.

4.2 Harmonic Balance Simulation:

Harmonic balance is a frequency-domain analysis technique for simulating distortion in nonlinear circuits. It is usually the method of choice for simulating analog RF and microwave problems, since these are most naturally handled in the frequency domain. Within the context of high-frequency circuit and system simulation, harmonic balance offers several benefits over conventional time-domain transient analysis. Harmonic balance simulation obtains frequency-domain voltages and currents, directly calculating steady-state spectral content of voltages or currents in the circuit.

In the context of high-frequency circuit and system simulation, harmonic balance has a number of advantages over conventional time-domain transient analysis: speed, accuracy, stability.

Harmonic balance Simulation Process:

The harmonic balance method [27] is iterative. It is based on the assumption that for a given sinusoidal excitation there exists a steady-state solution that can be approximated to satisfactory accuracy by means of a finite Fourier series. Consequently, the circuit node voltages take on a set of amplitudes and phases for all frequency components. The circuit is initially split into linear part and nonlinear part. Linear part of the circuit contains any matching networks, the bias network, device parasitics, sources, and so on. The nonlinear circuit contains only those elements within the device model whose value is a function of voltage or current. The currents flowing from nodes into linear elements, including all distributed elements, are calculated by means of a straightforward frequency-domain linear analysis. Currents from nodes into nonlinear elements are calculated in the time-domain. Generalized Fourier analysis is used to transform from the time-domain to the frequency-domain.

Nonlinear devices (transistors, diodes, etc.) in Harmonic Balance are evaluated (sampled) in the time-domain and converted to frequency-domain via the FFT. A frequency-domain representation of all currents flowing away from all nodes is available. According to Kirchhoff's Current Law (KCL) these currents should sum to zero at all nodes. The probability of obtaining this result on the first iteration is extremely small.

Therefore, an error function is formulated by calculating the sum of currents at all nodes. This error function is a measure of the amount which KCL is violated and is used to adjust the voltage amplitudes and phases. If the method converges (that is, if the error function is driven to a given small value), then the resulting voltage amplitudes and phases approximate the steady-state solution. The method is called harmonic balance because the harmonics in the linear and nonlinear "sides" must balance each other out. The harmonic balance procedure is thus an iterative procedure.

As can be seen close inspection of Fig. 4.1, the algorithm (harmonic balance) can be applied generally to any circuit whose driving function, and thus response, is periodic, and whose nonlinearity may be modeled as a time-domain expression of the chosen state variables. Most RF amplifiers, mixers, attenuators, and filters fall into this class of circuit, as well as many systems. In our case of mixers, where both an RF and an LO signal provide two (usually) nonharmonically related input fundamental frequencies, a two dimensional Fourier transform is required to support all possible linear combinations of these frequencies that are created within the circuit, including the IF [28].

The simulation procedure in ADS using harmonic balance is mentioned in the Appendix. After simulation of was done in ADS and satisfactory results obtained we proceed to the mixer implementation.

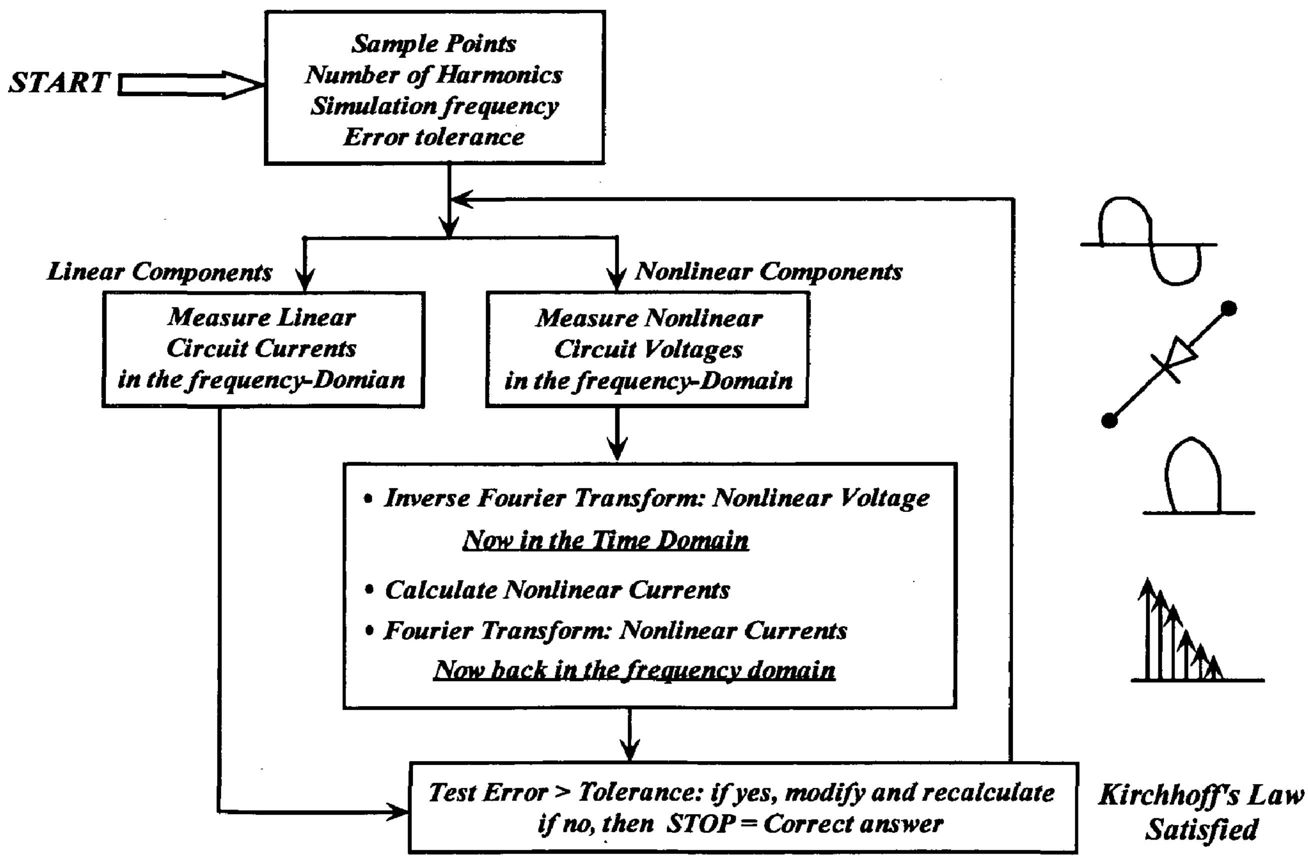


Fig. 4.1 Harmonic Balance Simulation Flow Chart

4.3 Mixer Implementation:

This section describes the conversion of circuit schematic to an actual transmission line layout. We used AUTOCAD for creating layout though the layout of the circuit is automatically generated using design synchronization in Agilent ADS tool, we prefer AUTOCAD for creating layout. Once layout is obtained we physically realize the circuit on the microstrip since microstrip is the one of the really practical ways to realize such circuits. By using photolithography procedure we make microstrip layout of the circuit. The photolithography procedure is described in the Appendix 4.

The layout of the circuit schematic of the designed mixer is shown in Fig. 4.2. The lumped elements and transistor are properly mounted on the microstrip layout at appropriate place as shown in Fig. 4.2.with the help of soldering. Precautions were taken while making physical circuit and while soldering the lumped components. Etching was done properly and we use wet etching in the photolithography procedure. The whole process of layout and fabrication was done in the RFIC section at the Institute. Voltage is given to the circuit through a copper wire in which one end of the wire is soldered on the PCB and other end is given to supply.

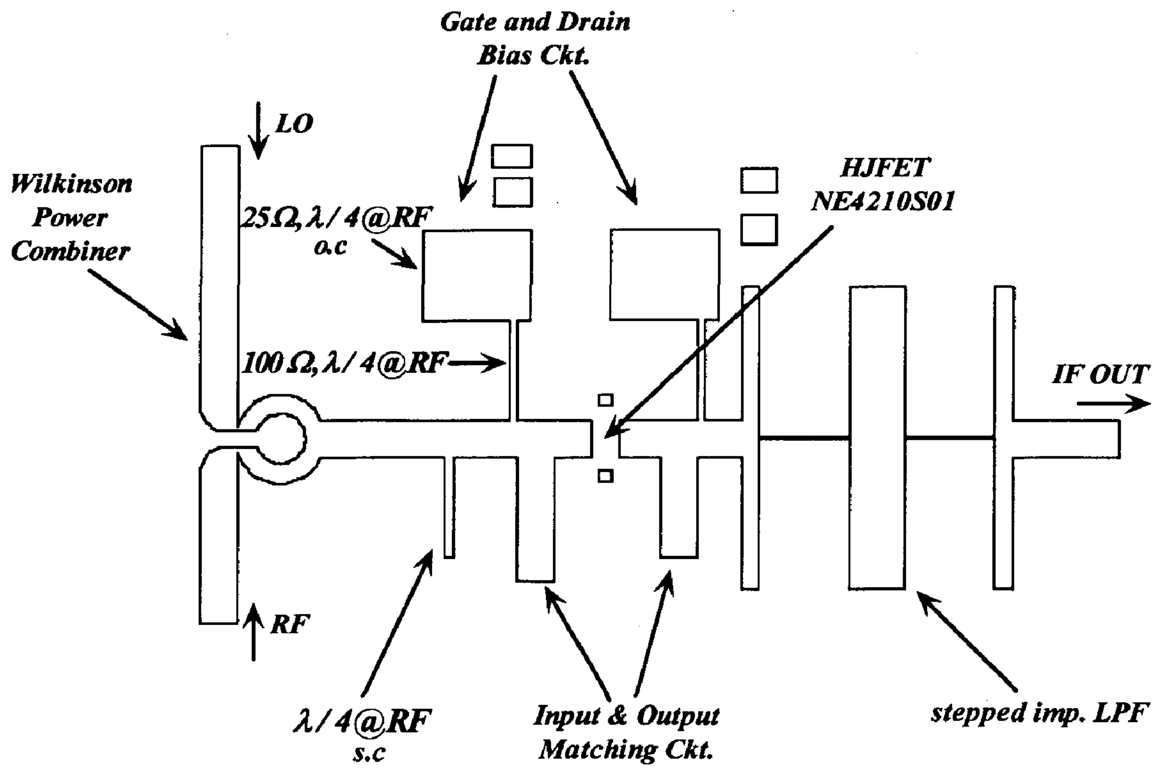


Fig. 4.2 Layout of the Designed Mixer

While implementing Wilkinson power combiner some precautions are to be taken. One such precaution is connecting input ports with respect to the chip resistor. As shown in Fig. 4.2, the input ports should be connected to the from the chip resistor terminals. The arc length should be 8.45mm. The gap left for chip resistor is 2mm. The total mean circumference is 18.9078mm. The outer circle radius is 3.8578mm while radius of inner circle is 2.1638mm.

The source terminal of the transistor was grounded by making hole on the PCB layout and a contact was made to the ground of the circuit via a thin copper wire. The PCB was mounted on a 0.5" aluminum metal and tightened with screws. The connectors for input and output ports are placed as shown in the Fig.4.5. The fabricated mixer dimensions are 8.5cm x 5cm.

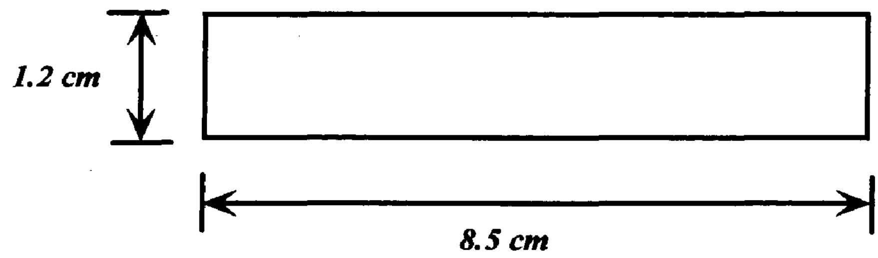


Fig. 4.3 Side View of base metal

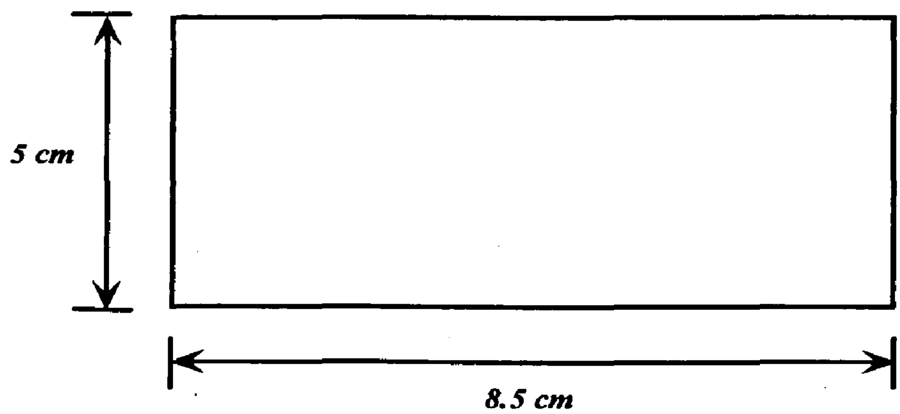


Fig. 4.4 Top view of base metal

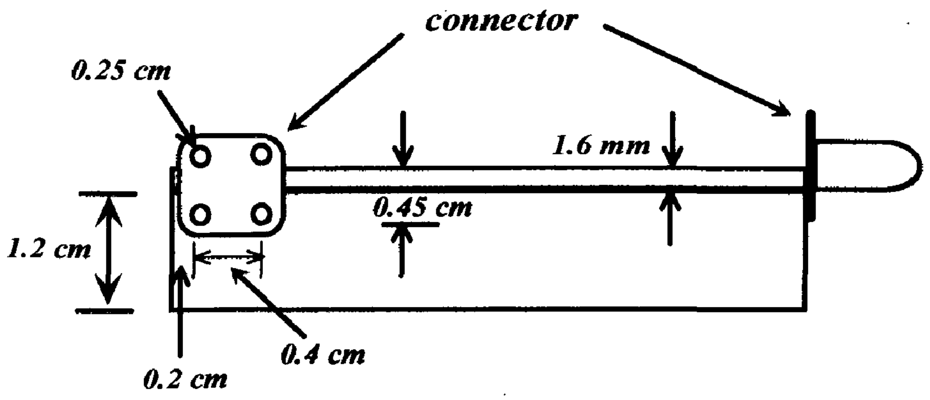


Fig. 4.5 Different dimensions of metal and connector

Final mixer was fabricated for measurement test which was shown in Fig. 4.6.

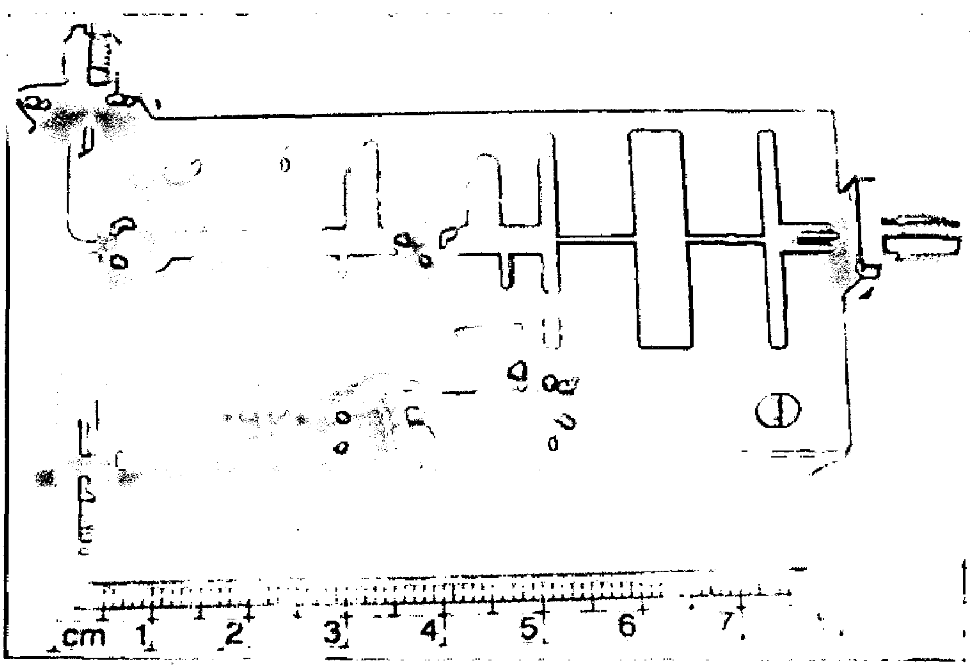


Fig. 4.6 Fabricated mixer

6. RESULTS

The mixer is simulated in ADS, the steps to be followed to simulate in ADS software is described in Appendix 1,2,3. By specifying the number of harmonics to be considered for the LO and RF input frequencies and the maximum order (highest order of sums and differences) to be retained, we get the frequency domain result of the mixer at all relevant frequencies using harmonic balance analysis technique which is incorporated in ADS tool.

The mixer test setup is placed in the Advance Microwave Lab. Mixer test equipments includes Rohde & Schwarz FSP- Spectrum analyzer with range of 9khz - 30GHz and Rohde & Schwarz signal generators SMR 27 with range 10 MHz – 27 GHz. A spectrum analyzer gives a frequency-domain representation of an input signal, displaying the average power density versus frequency. Thus, its function is dual to that of the oscilloscope, which displays a time-domain representation of an input signal. A spectrum analyzer is basically a sensitive receiver that tunes over a specified frequency band and gives a video output that is proportional to the signal power in a narrow bandwidth. Spectrum analyzers are invaluable for measuring modulation products, harmonic and intermodulation distortion, noise and interference effects.

Fig. 5.1 shows the block diagram of measurement setup and the actual setup is shown in Fig. 5.2. The mixer is tested under proper bias conditions with two power supplies. The mixer is operated in temperature between 18-22^{o}c as two R & S signal generators generates two input signals at the desired frequencies i.e at 5.1 GHz (RF) and 5.3 GHz (LO) respectively. Proper care has to be taken with the supply sources as they can damage the transistor with initial spikes in current or current from power supplies. Standard 50 ohm cables are used to connect between device and signal generators and the analyzer.

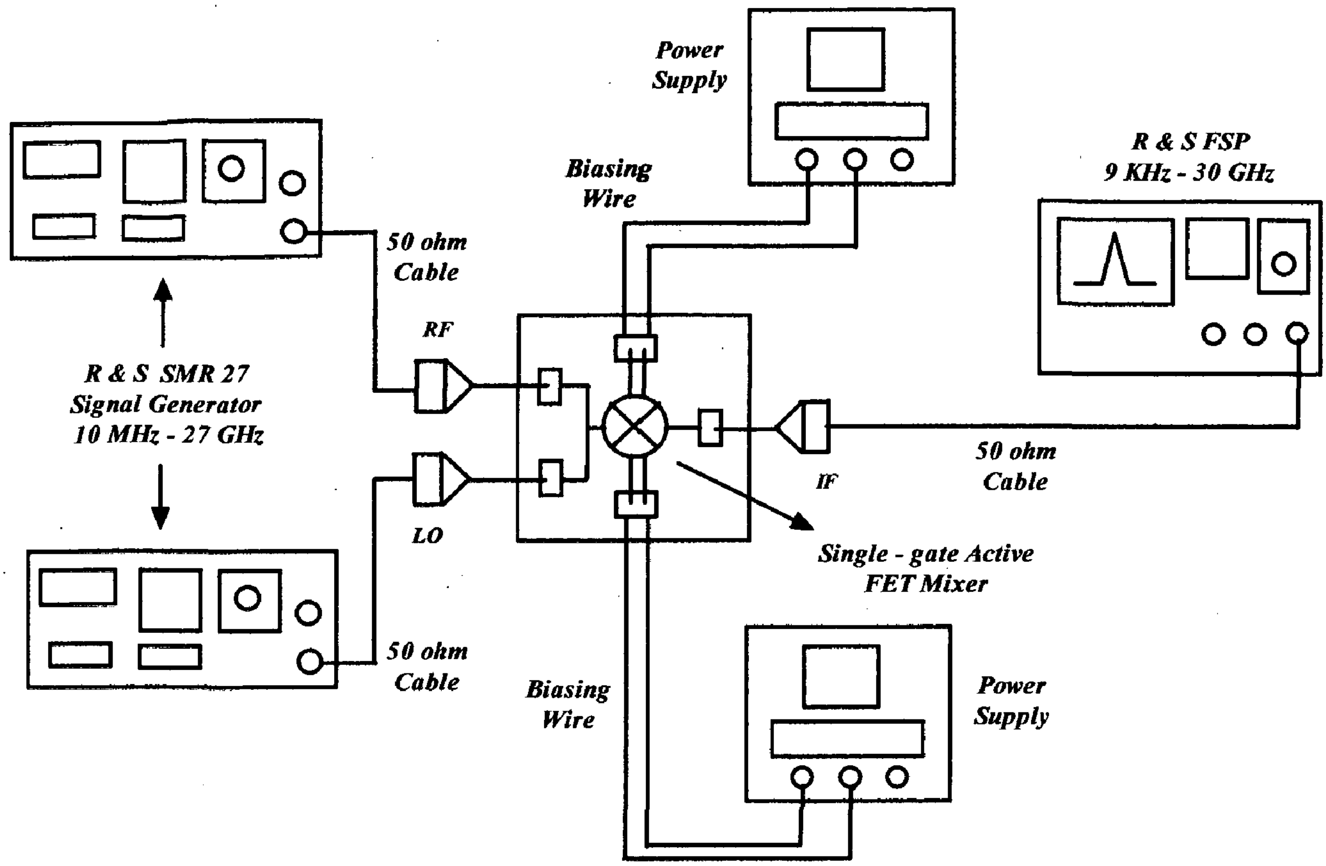


Fig. 5.1 Mixer Measurement Setup

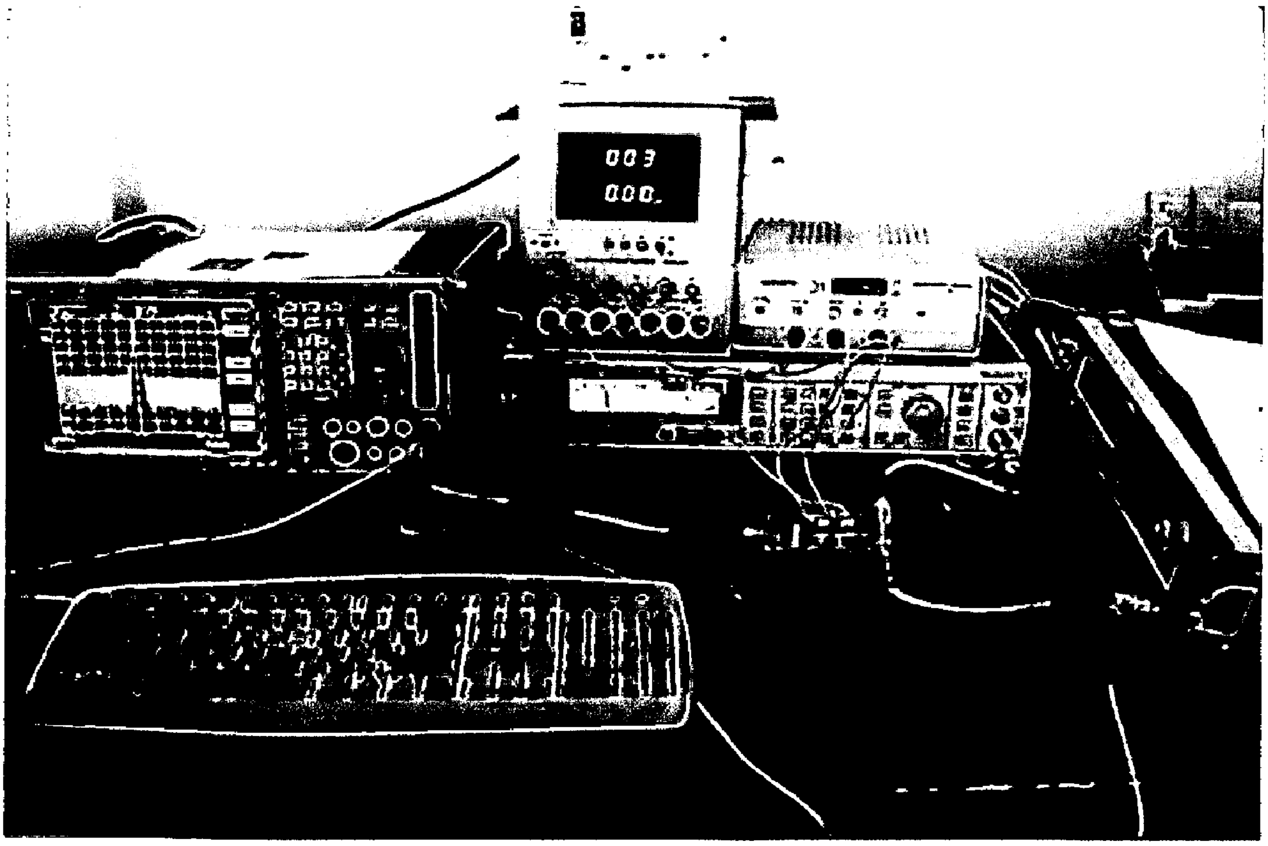


Fig. 5.2 Measurement setup at Advance Microwave Lab

The following results are measured during the test.

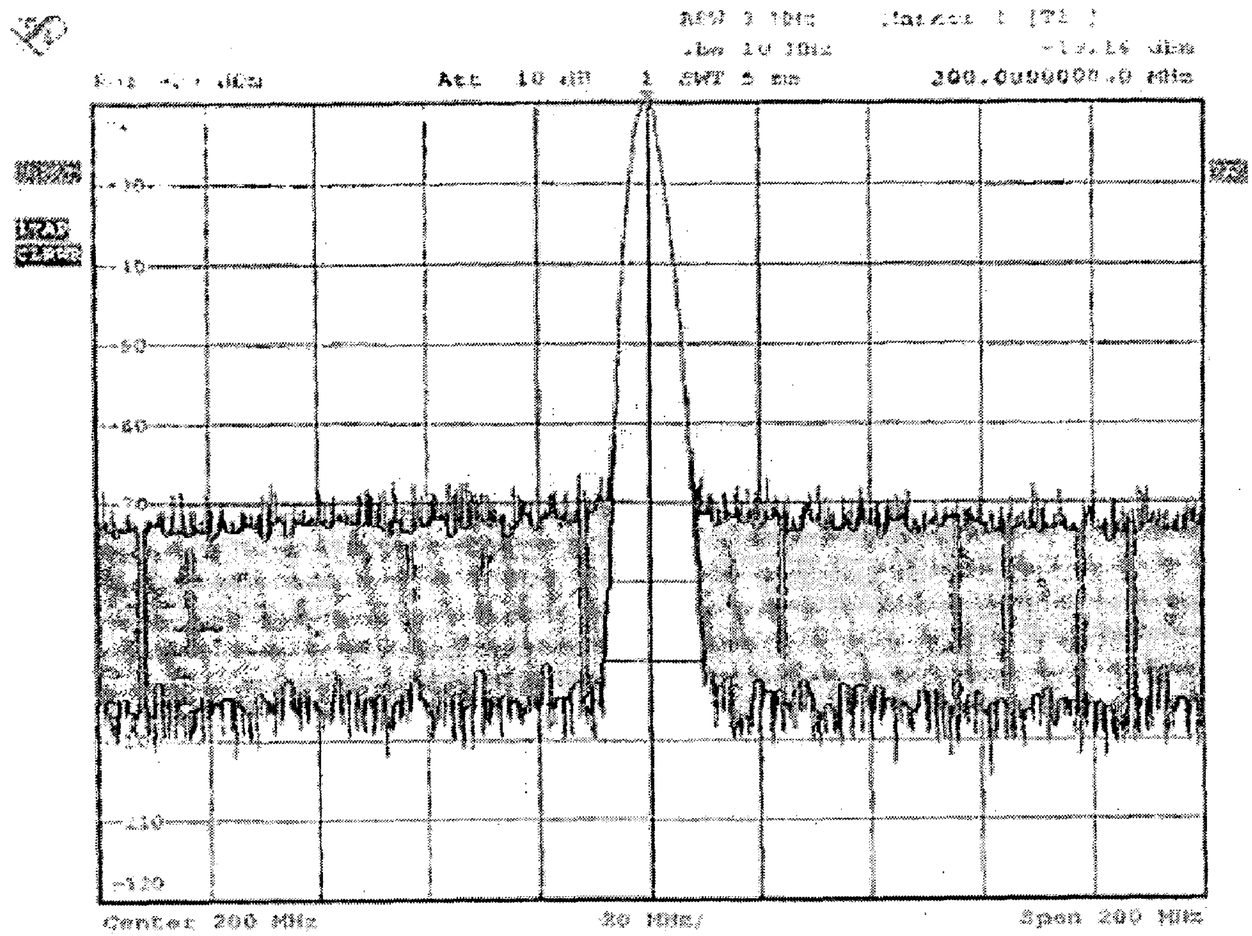


Fig. 5.3 IF generation on FSP

Mixer output tones:

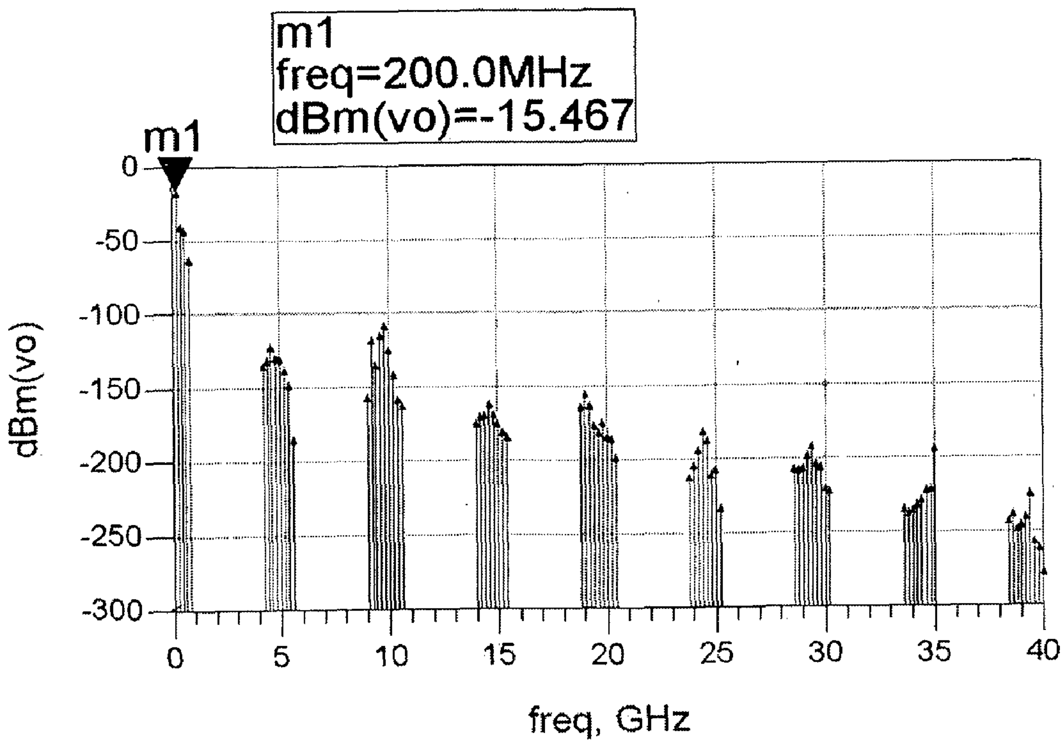


Fig. 5.4 Simulated Mixer Output Tones.

Table 5.1 Measured and Simulated IF Power

RF POWER, 5.1 GHz	-20 dBm
LO POWER, 5.3 GHz	2 dBm
SIMULATED IF POWER	-15 dBm
MEASURED IF POWER	-18 dBm

Gain Compression P1dB:

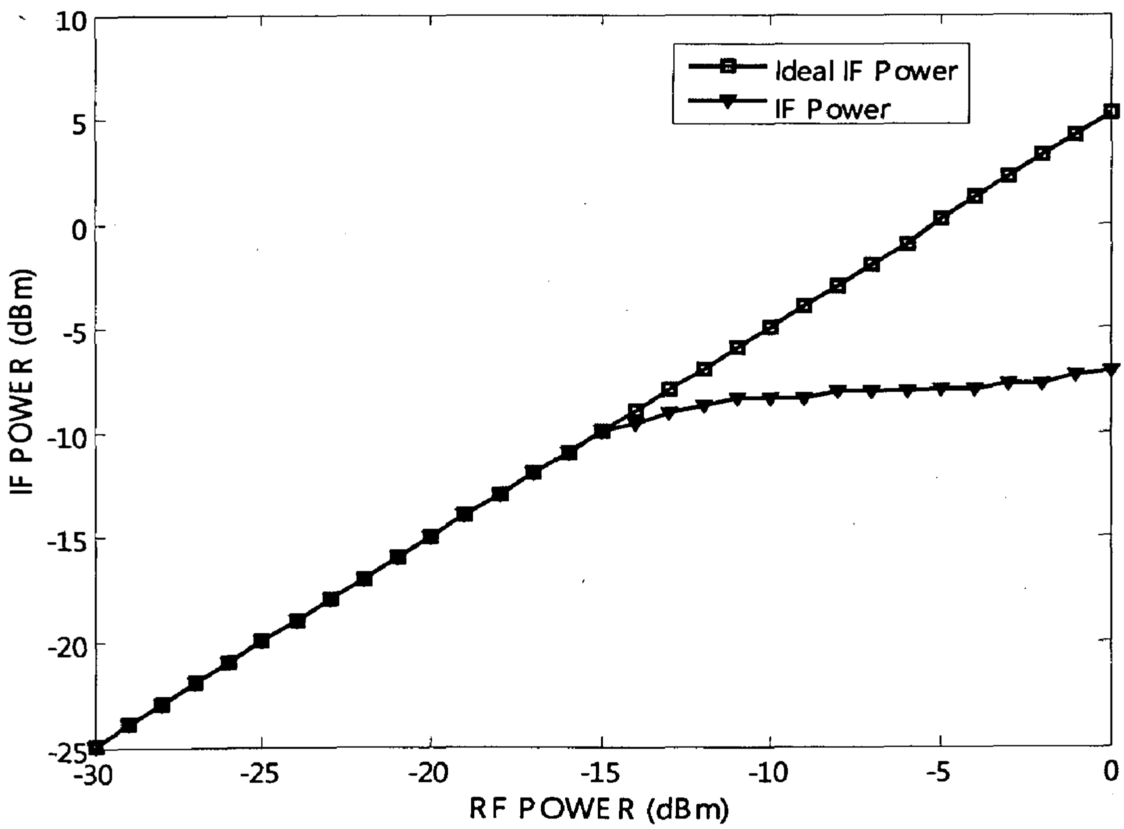


Fig. 5.5 Gain compression plot (Simulated result).

From the above plot we can observe that gain is compressed by 1dB at -14dBm of RF power. After increasing RF power, gain will not increase and there is no conversion gain of the mixer. This shows the conversion gain degrades at large input signals levels due to nonlinearity in the signal path. Model graph is as shown in Fig.2.10.

Conversion gain vs LO power:

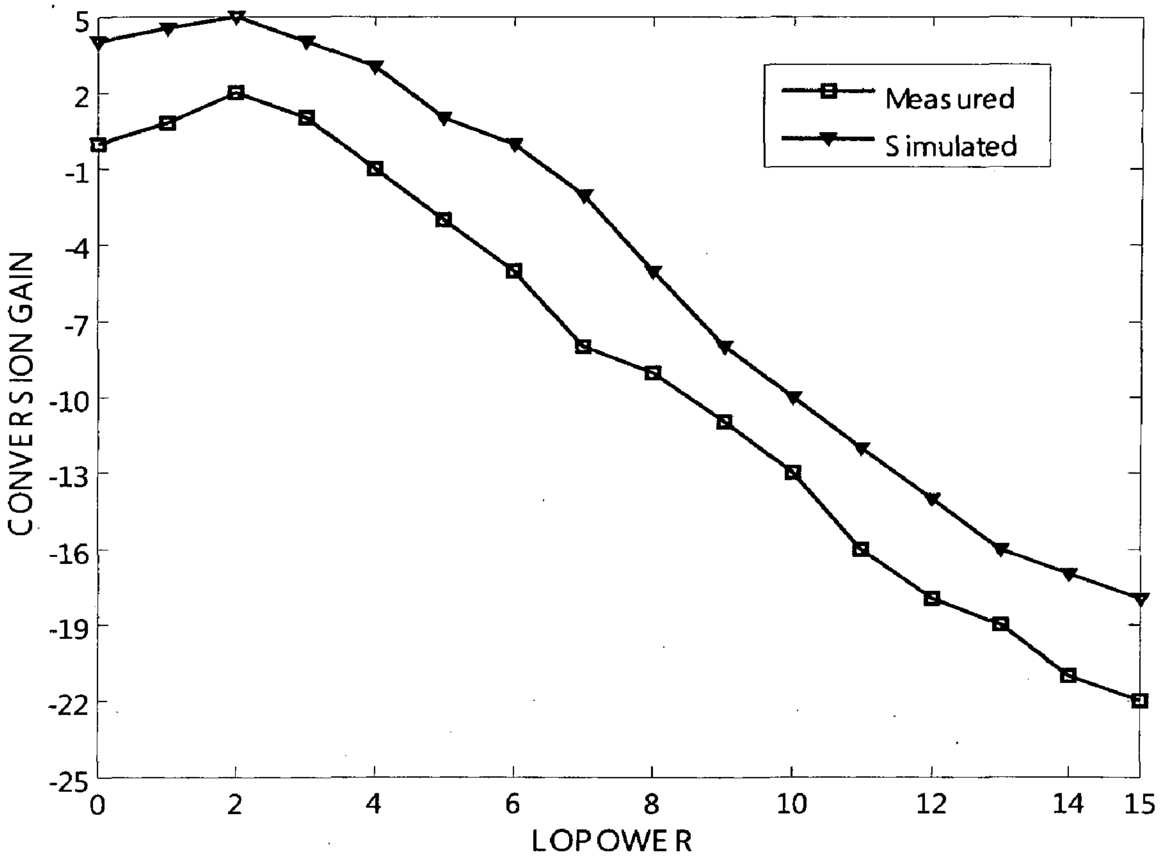


Fig. 5.6 Conversion gain vs LO Power

RF frequency = 5.1 GHz.

LO frequency = 5.3 GHz.

RF Power = -20 dBm.

LO Power is varied from 0 dBm to 15 dBm.

Conversion gain vs RF Power:

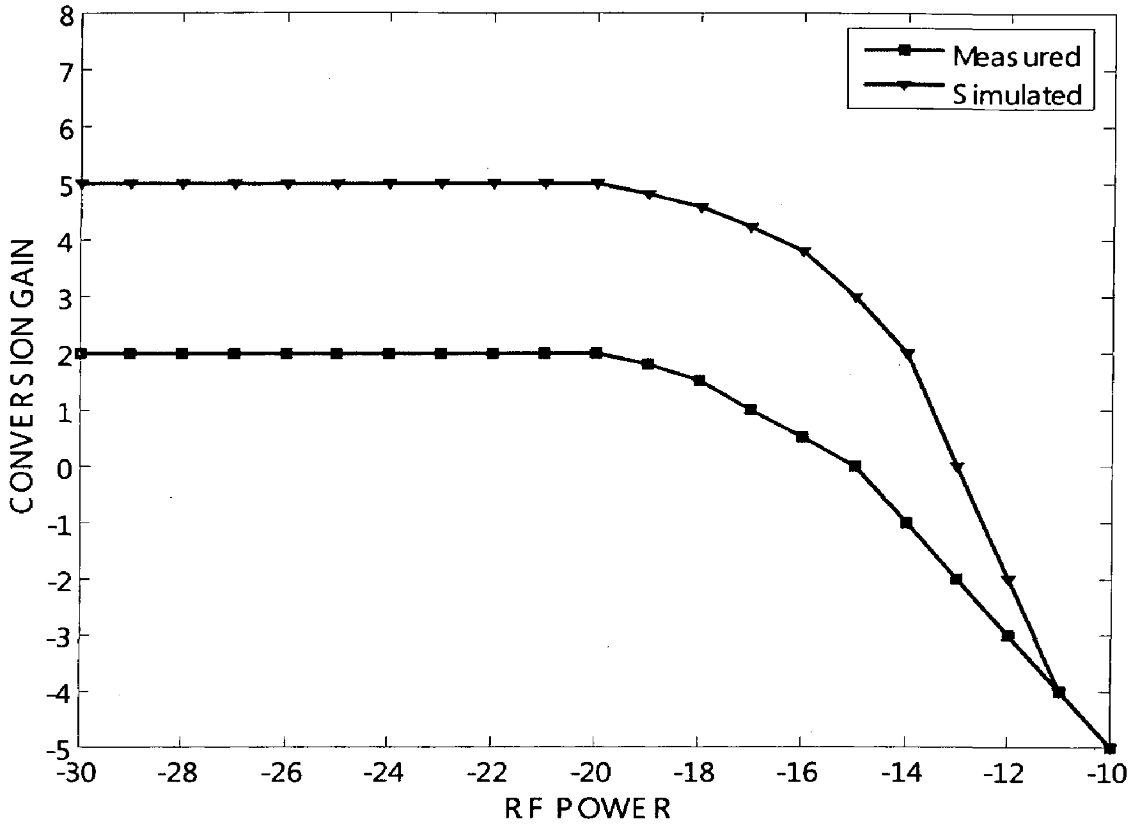


Fig. 5.7 Conversion gain vs RF Power

RF frequency = 5.1 GHz.

LO frequency = 5.3 GHz.

LO Power = 2 dBm.

RF Power varied from -30 dBm to -10dBm.

IF Frequency vs Conversion Gain: Measured Performance

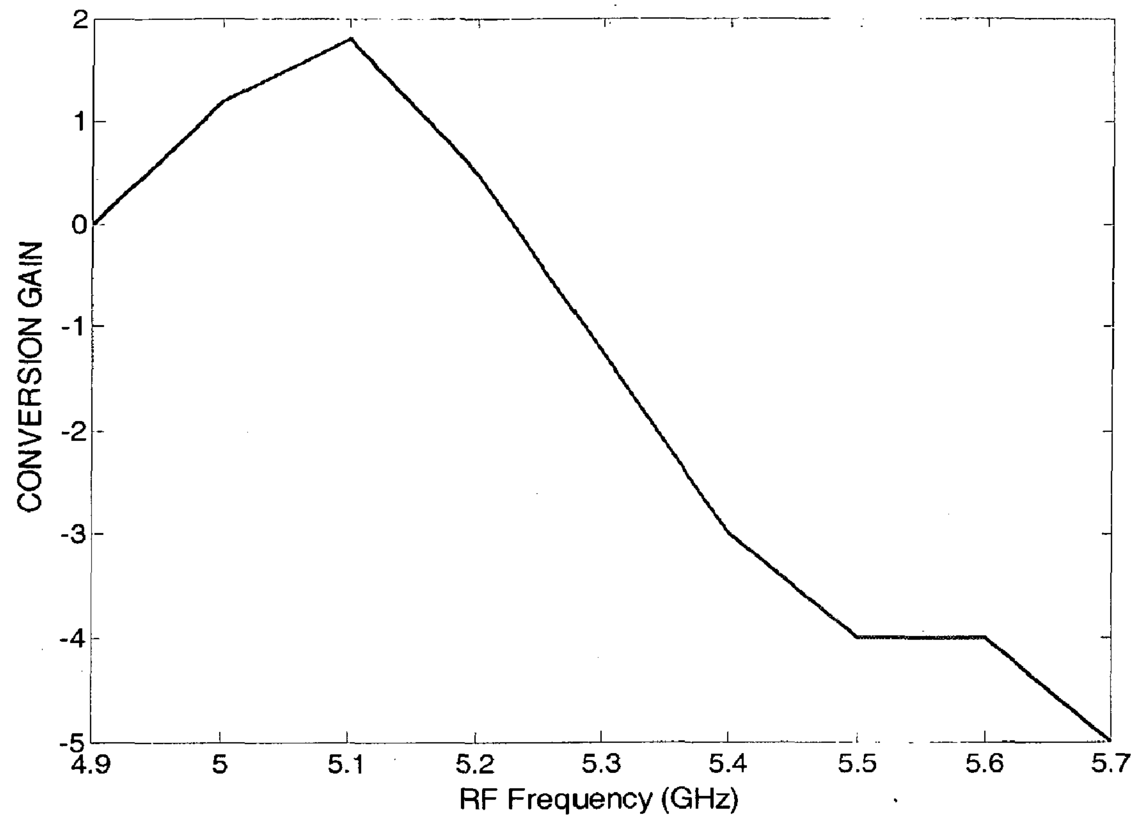


Fig. 5.8 IF frequency vs Conversion Gain (Measured Plot)

LO Power = 2 dBm

RF Power = -20 dBm

LO frequency = 5.3 GHz

RF Frequency varied from 4.9 GHz to 5.7 GHz

In this chapter, from the above plots, we have shown the performance of the mixer both in simulation and in measurement. The measurement results are slightly deviated from the simulated results, due to the 1 dB loss in 50 ohm cables, the operating conditions and substrate losses. The isolation between RF-LO is 15 dB is shown in Fig. 3.16 (c) and between LO-IF is 20 dB. Thus we have a good agreement in the simulation and measurement results.

6. CONCLUSION AND FUTURE SCOPE

The main task involved in this thesis is to come up with the physically built mixer that works in 5GHz UNII band. Important figure of merits of device such as gain, noise figure, compression point and isolation are introduced and described. Basic theory and operation of mixers have been presented. Mixer and its individual components are designed, analyzed on ADS, implemented. Also the specifications of unknown substrate such as dielectric constant and thickness for implementation of mixer are calculated and experimentally measured. The present mixer is low cost, works with low LO power and also meets the specifications stated in Table 1.1 which are suitable for 5 GHz WLAN applications.

Here presented the future work for present mixer design:

Future Work:

- With an antenna, amplifier at the receiver frontend and with A/D converter at the receiver back end, total receiver can be designed or even implemented.
- Mixer bandwidth can be improved and can be operated in dual band by the involvement of latest power combining techniques.

As already stated that other type of mixers offer better performance for some of these (Conversion Gain, Noise figure, Isolation) characteristics, but none offers better performance for all of them. Every designer chooses a particular type, topology and comes up with optimum performance characteristics for one of the mixer performance specifications. So choice of mixer depends upon the application and requirements of the designer. From past to the present in the designing of mixer there is no research gap, this shows the importance of the mixer in communication systems. After all

"No Mixers, No Communication Transceivers".

APPENDIX

1. PERFORMING BASIC MIXER SIMULATION:

For a successful analysis:

- By convention, the mixer input port is considered to be port 1, the IF output port is port 2, and the LO input port is port 3. Set up sources and port numbers so that they match the mixer convention. Do this by editing the *NUM* field for these components.
- Ensure that frequencies are established for all of the frequencies of interest—RF, LO, and IF—in the design.
- Add the harmonic Balance simulation component to the schematic and double-click to edit it. Fill in the fields under the Freq tab:
  - Enter the RF, LO, and IF as fundamental frequencies and set the order.
  - Assign the LO frequency to Freq [1]; it is easier to achieve convergence if the frequency of the signal with the largest amplitude is assigned to freq [1].
  - Since the number of tones in the simulation can affect simulation time, set the Maximum order to limit the number of tones that are considered in the simulation.

2. FINDING MIXER OUTPUT TONES:

The description is a two-tone harmonic balance analysis. One tone is used for the RF, one for LO. At the IF we expect to find two fundamental tones at f_{RF}-f_{LO} and $f_{RF+}f_{LO}$. We also expect to find spurious tones.

To find the amplitudes and relationships of mixer output tones:

1. From the **Sources-Freq Domain** palette, select **P_1Tone** and place one instance of it at the input (PORT1) and another at the LO input (PORT2).
2. Edit the RF input source:
   - Num = 1. This is port 1.
   - Z = 50 ohms (default value)
   - P = dbmtow(P_RF). This converts the 0-dBm input to watts (the power unit used by the system).
   - Freq = Rffreq.
3. Edit the LO input source:
   - Num = 2. This is port 2.
   - Z = 50 ohms (default value)
   - P = dbmtow(P_LO)
   - Freq = LOfreq.
   - Noise = No. Unless we are doing a noise simulation, this parameter is ignored.
4. From the **Simulation-HB** palette, select the **HB** simulation component and place it on the Schematic. Edit to select the **Freq** tab and edit the following
   - Maximum order = **8**
   - Frequency = **LOfreq.** This is Freq [1]. Set its order to **8.**
   - Click **Add** to enter the second fundamental, Freq [2]. Set its frequency to **Rffreq,** and leave its order set to **8.**
5. **Simulate.** When the simulation is finished, plot V_{if} in dBm, then place markers on the two fundamental IF tones, RFfreq – LOfreq as shown in fig.4.1

3. GAIN COMPRESSION SIMULATION:

1. Follow the same steps as in mixer tones simulation up to step 4. Note we have to sweep RF power, so sweep parameter option will be in HB simulation component.
2. VIF is the output voltage at the IF output frequency and must be selected from many frequencies in the output data set. This frequency is selected by using the *mix* function. In this simulation, LO freq = 5.3 GHz and RF freq = 5.1GHz. If we are interested in the downconverted IF frequency, 200MHz, we can select it from:

$$V_{IF} = \text{mix}\ (V_o, \{1,-1\}).$$

   V_o is the output voltage at different frequencies.

   The indices in the curly brackets are ordered according to the HB fundamental analysis frequencies. Thus, {1,-1} represents LO freq-RF freq.

3. Other equations are added in the display panel which calculate the conversion gain

$$\text{ConvGain} = \text{dBm}(V_{IF}) - \text{P\_RF}.$$

Here we can identify the 1 dB gain compression power to be about -14dBm of RF power in fig. 5.5.

4. PHOTOLITHOGRPHY PROCEDURE:

- Note down the dimensions of the designed circuit after finalizing the circuit. Draw the layout of the circuit in AUTOCAD software. Take printout using 1:1 ratio on transparency. This is the mask of the circuit.
- Clean the copper laminate used in MIC using powdered soap and rinse it under running water. Dry the substrate and other glassware using heat convertor. Substrate can be further cleaned using Acetone.
- Small volume of Negative Photo Resist (Negative PR) is taken and it is placed on the substrate. Coat the PR on the substrate using spin coating system and dry it in oven for 10 minutes at 100^0 c.
- The mask of the circuit is aligned on the coated side of the substrate and fixed it with the substrate using tape.
- The circuit is exposed under UV light for 2 to 2.5 minutes.
- Remove the mask from the substrate and develop the exposed circuit in developer (TCE) for 1-1.5 minutes and then dip it in fixer (Acetone) for around 1 minute.
- Use Blue dye so that layout is clearly visible. Heat the circuit in oven at 100^0C for 10 minutes.
- Use brown tape to protect ground plane of Microstrip and etch it using $FeCl_3$ solution. Clean the circuit under running water and use acetone for further cleaning.

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