DOKUZ EYLÜL UNIVERSITY
GRADUATE SCHOOL OF NATURAL AND APPLIED SCIENCES

RECEIVER DESIGN FOR INDOOR WLAN BY USING MICROSTRIP STRUCTURES

by
Deniz KORKMAZ

August, 2005
İZMİR

RECEIVER DESIGN FOR INDOOR WLAN BY USING MICROSTRIP STRUCTURES

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M.Sc THESIS EXAMINATION RESULT FORM

We have read the thesis entitled **"RECEIVER DESIGN FOR INDOOR WLAN BY USING MICROSTRIP STRUCTURES"** completed by **DENİZ KORKMAZ** under supervision of **Asst. Prof**. **Dr. YEŞİM YÜKSEL** and we certify that in our opinion it is fully adequate, in scope and in quality, as a thesis for the degree of Master of Science.

Asst. Prof. Dr. YEŞİM YÜKSEL

Supervisor

(Jury Member) (Jury Member)

Prof.Dr. Cahit HELVACI
Director
Graduate School of Natural and Applied Sciences

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Deniz KORKMAZ

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ABSTRACT

Wireless applications are becoming more and more predominant today. There are various Wireless Local Area Networks (WLAN), which have been implemented to meet some demands. When designing such a network a number of factors have to be considered. These are range, maximum bit rate, portability, power consumption and cost. This thesis covers the specifications, design and simulation of a system that addresses these factors. The objective of this thesis is to build a Receiver for Frequency Shift Keyed (FSK) wireless local area network (WLAN) that operates in the microwave frequency range by using microstrip structures. The system consists in antenna, amplifier, power divider, bandpass filter, diode detector and voltage comparator circuits. The circuits are built using AWR Microwave Office software since they are microstrip circuits that are planar, small, easy and fast to make, and cheap except voltage comparator.

Keywords: Receiver, microstrip circuits, WLAN

BİNAİÇİ KABLOSUZ YEREL ALAN AĞLARI İÇİN MİKROŞERİT YAPILARLA ALICI TASARISI

ÖZ

Günümüzde kablosuz uygulamalar gitgide üstün olmaktadır. İhtiyaçları karşılamak için gerçekleştirilen çeşitli Kablosuz Yerel Alan Ağları mevcuttur. Böyle bir ağ tasarlanırken bazı faktörler göz önünde bulundurulmalıdır. Bunlar; menzil, maksimum bit hızı, taşınabilirlik, güç tüketimi ve maliyettir. Bu tez; bu faktörleri karşılayan bir sistemin teknik özelliklerini, tasarısını ve simülasyonunu kapsar. Bu tezin amacı, mikro-şerit yapılar kullanarak mikrodalga frekans bölgesinde çalışan Frekans Kaydırma Anahtarlamalı Kablosuz Yerel Ağı için bir alıcı oluşturmaktır. Sistem, anten, yükselteç, güç bölücü, bant-geçiren filtre, diyot detektör, gerilim karşılaştırıcı devrelerinden oluşur. Gerilim karşılaştırıcısı hariç devreler, düzlemsel, küçük, ucuz, yapımı kolay ve hızlı olan mikroşerit devreler olmasından dolayı, AWR Microwave Office yazılımı kullanılarak oluşturuldu.

Anahtar Sözcükler: Alıcı, mikroşerit devreler, Kablosuz Yerel Alan Ağları

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CHAPTER ONE
INTRODUCTION

1.1 Wireless Local Area Networks

Over the past decade, computer technology has improved by leaps and bounds. This can be illustrated by Moore's law. Gordon Moore, a co-founder of the computing giant, Intel, predicted in the 1970s that computing power would double every 18 months. So far, this prediction has held true. Users of this technology are finding it hard to keep up with the latest models because of the pace at which new technology is developed. The computer – once a high performance machine that only large companies could afford - has come down tremendously, both in size, and in price. Personal computers are amongst many electronic devices which have been made more compact and light weight. The reduction in size and weight of such powerful computers has given modern day workers the ability to carry their work with them. We have been progressively moving towards an increasingly more portable society, lending momentum to the development of wireless technology. (Web_23, 2005)

Conventional wired Local Area Network (LAN) systems offer reliable and a high rate of data transmission, between computer terminals, using cables such as coaxial or optical fibre. However, once the cables are in place, future expansion or reconfiguration of the LAN must be met with either replacing or removing existing cables. This may cost as much as a new installation which can sometimes be as high as 40% of total system cost. Typically, LANs are found in work places where the office environment may require constant change as user progressively outpace installed system capabilities.

Wireless Local Area Network (WLAN) development is still in its infancy, however it can be considered as a viable alternative to the high cost of maintenance and reconfiguration of wired network systems.

There are two competing technologies that can be used to implement a WLAN – unguided optical signals and radio-frequency waves. Current developments suggest a stronger future for radio-frequency waves as the favoured technology for the implementation of a WLAN.

A Local Area Network, which implements wireless interconnectivity between peer-topeer (e.g.: PC-to-PC, PC-to-hub, or printer-to-hub) and point-to-point (LAN-to-LAN) connectivity within a building or campus, may be termed as a wireless LAN or WLAN. In place of copper or fiber physical layer, typically employed in a wired network, WLANs employ radio or electromagnetic waves and are therefore also known as wireless radio LANs. Normally, WLANs provide a mechanism to connect to the wired network through access points, which act as bridges to a backbone distribution system.

The primary advantages of a WLAN are:

1. The lack of wiring, leading to lower installation and maintenance costs
2. Mobility; with a wireless connection, users can be free to roam a facility while having access to network data, which can be useful for office or hospital use.
3. Seamless network connection allowing ad hoc networks to be setup.

Building-to-building wireless bridges are also emerging as an application area for WLANs, offering an attractive alternative to traditional means of networking when

1. Phone company lines are too expensive or slow for large data traffic
2. Distance/terrain makes campus wiring impossible or difficult
3. Companies want to manage their own interoffice communications

Due to the reasons discussed above, wireless LANs have gained strong popularity in a number of vertical markets, including the health-care, retail, manufacturing, warehousing, and academia. Today wireless LANs are becoming more widely recognized as a general purpose connectivity alternative for a broad range of business customers. Business Research Group, a market research firm, predicts a sixfold expansion of the worldwide wireless LAN market by the year 2000, reaching more than $2 billion in revenues. Table 1.1, shows various configurations of wireless networks, including WLAN. (Web_23, 2005)

Table 1.1 WLANs and other wireless technologies (Web_23, 2005)

	WLAN	LAN-LAN Bridge	WWAN (wide area network)	WMAN (metro area network)	WPAN (Personal area network)
Coverage Area	In building or campus	Building to Building	National	Metro area	A few feet
Function	Extension to wired LAN	Alternative to wired network	Extension of wired network	Extension of wired network	Alternative to cable
User Fee	No	No	Yes	Yes	No
Typical Throughput	1-11 Mbps	2-100 Mbps	1-32 Kbps	10-100 Kbps	0.1 to 4 Mbps

A WLAN implements a flexible data communication system frequently augmenting rather than replacing a wired LAN within a building or campus. WLANs use radio frequency to transmit and receive data over the air, minimizing the need for wired connections. Point-to-point is a method of communication between one receiver and one location. Point-to-point has a higher bandwidth than point-to-multipoint for reasons including that it has less overhead to manage the data paths and there is only one receiver per transmitter.

1.2 Microstrips Circuits

In addition to traditional coaxial and waveguide components, many other types of transmission line and waveguide can be used at microwave frequencies. One of the most often used structures today is the microstrip, shown in Fig. 1.1. From a few hundreds of

MHz to 10 GHz, microstrips are widely used because of its simplicity and its low fabrication cost. The wave is guided between the bottom ground plane and the top metal strip. Some of the fields spill over from the dielectric into air. The presence of the dielectric complicates things, but people have been able to use quasi-static analysis to obtain formulas for the impedances and propagation constants in microstrip lines. In formulas, the so called efective dielectric constant is used. This is just some kind of average between the permittivity of air and the dielectric that gives us a rough picture of the portion of field that remains in the dielectric.

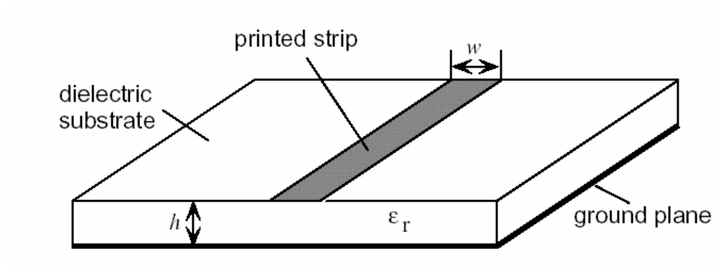


Figure 1.1 A microstrip transmission line printed on a grounded dielectric substrate (WEB_18, 2003)

- The higher the dielectric constant, the thinner the line, keeping the thickness of the dielectric and the impedance of the line constant.
- The thinner the dielectric, the thinner the line is, keeping the dielectric constant and impedance of the line constant.
- The higher the dielectric constant, the smaller the circuit is. The wider the line, the lower the impedance.

Microstrip circuits became popular because they are planar, small, easy and fast to make, and cheap. However, they cannot handle very high power levels and they are more lossy than coax or waveguide. (Web_18, 2003)

Advantages of microwave circuits:

- More bandwidth. A 1% bandwidth provides more frequency range at microwave frequencies thet at HF, for example.
- Microwave signals travel predominately by line of sight. Plus, they don't reflect off the ionosphere like RF signals (<30 MHz pr so). Consequently, communication links between (and among) satellites and terrestrial stations are possible.
- At microwave frequencies, the electromagnetic proparties of many materials are changing with frequency. This is due to molecular, atomic and nuclear resonances. This behavior is useful for remote sensing and other applications.There is much less background noise at microwave frequencies than at RF.

Microwave circuits have also some difficulties. As was mentioned, microwave circuits are much more difficult to analyze than low frequency ones. Voltage is not well defined if the distance between the two points is not electrically small. At microwave frequencies, "electrically large" distances may be just a few millimeters! Moving the leads around will also likely affect voltage measurements. One must carefully choose lumpet elements (L, C, R, diyotes, transistors, etc.) for use in the microwave region. Typical low frequency components do not behave as expected. To "transport" electrical signals from one position to another, one must use special wires. It is more common to speak of "guiding" signals at these frequencies. (Web_25, 2003)

1.3 Fundamentals Of This Project

WLANs are commonly used to connect devices such as computers and printers or other peripherals. In recent years more and more users expect to have the ability to operate their portable computer globally while remaining connected to the communications network and service providers. So far there are various WLAN, which have been implemented to meet these demands. When designing such a network a number of factors have to be considered. These are range, maximum bit rate, portability, power consumption and cost. This thesis covers the specifications, design and implementation of a system that addresses these factors. In this project, it was built a receiver for a Frequency Shift Keyed (FSK) WLAN by using microstrip structures. In all microstrip structures, the length and width of the microstrip lines is found by using TX-Line in the AWR-Microwave Office software package. Width and length values of each microstrip lines are calculated according to impedance and frequency values. Properties of substrate are important to calculate.

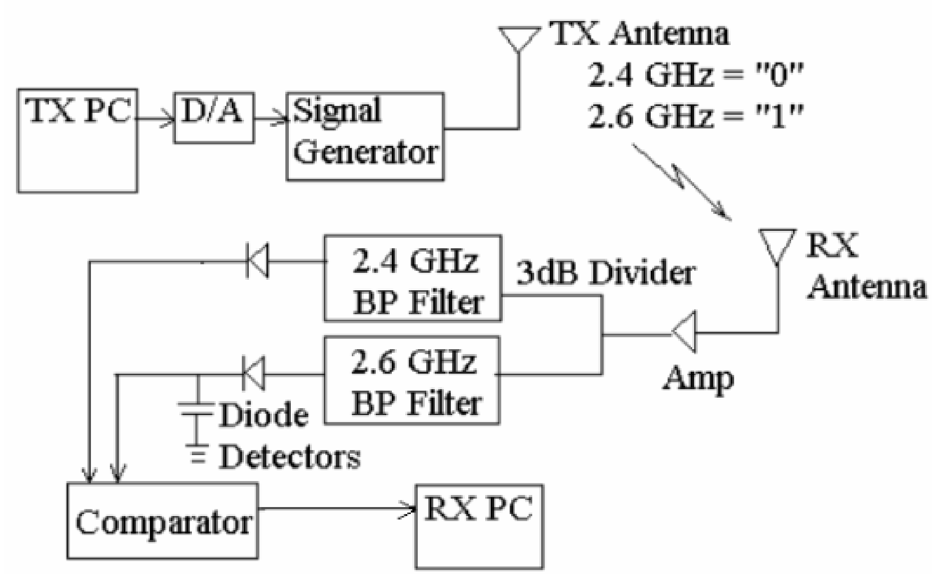


Figure 1.2 Wireless local area network block diagram (Furse, Woodward, & Jensen, 2004)

Figure 1.2 shows a block diagram of the FSK transmitter/receiver system. The concept of this FSK system operation is that a digital “0” is transmitted at 2.4 GHz, and a digital “1” is transmitted at 2.6 GHz. Thus, either a 2.4 GHz or 2.6 GHz signal is transmitted from the first computer to receiver board. The job of the receiver is to determine which frequency has been received, and convert it back into a digital “0” or “1”. The receiver antenna must be well-matched to both the 2.4 and 2.6 GHz signals. The signal from the antenna is then amplified and split into two identical copies using a 3 dB power divider. The narrow-band filters at 2.4 and 2.6 GHz pass or reject depending on if the signal was a “0” or “1”. Only one filter should pass at a time. The signal envelopes are extracted using diode detectors fabricated from zero-bias schottky diodes, and the resulting waveforms are run into a voltage comparator whose output voltages are set to those of TTL logic. This output signal can then be received by the a personal computer and interpreted using acquisition software.

Each block of a such receiver system was done as a separate chapter. Chapter II, III, IV, V, VI, VII cover antenna, amplifier, power divider, band-pass filter, diode detector and voltage comparator circuits respectively.

CHAPTER TWO
ANTENNAS

2.1 Microstrip Dipole Antenna

In wireless communication systems, such as WLAN, research and development efforts are aiming at smaller size and better performance. In addition to the use of signal processing techniques to improve communication channel capacity, the radiation characteristics of the portable antenna system is also very important for communication performance. An antenna can also be viewed as a transitional structure (transducer) between free-space and a transmission line (such as a coaxial line). An important property of an antenna is the ability to focus and shape the radiated power in space e.g.: it enhances the power in some wanted directions and suppresses the power in other directions.

Microstrip antennas are frequently used in today's wireless communication systems. Thanks to their low profile, they can be mounted to the walls of buildings, to the fuselages of airplanes or to the reverse sides of mobile phones. Moreover, microstrip antennas are fabricated using the same technology as producing printed circuit boards. Therefore, the fabrication is relatively simple and well reproducible. Finally, microstrip antennas can be simply integrated directly to microwave circuits, which are based on microstrip technology, and therefore, no special transmission lines, symmetrization circuits or connectors are needed on the contrary to classical antennas. Nevertheless, there are several disadvantages of microstrip antennas. Narrow operation band is the main disadvantage. Microstrip antennas find use in low-power and high frequency applications.

Simplistically, a single-layer microstrip antenna consists of a dielectric material (substrate) sandwich between two metal layers. The top metal layer only partially covers the substrate and forms the microstrip circuit. The bottom metal layer acts as a ground plane. Microstrip antennas will find their use as planar devices that can be mounted, in close proximity, to practically any structure. The ground plane acts as a shield to prevent radiation from the antenna being reflected from these structures.(Web_9, 2005)

Ideally, the antennas used for this project should be directional with good matching networks that provide a low standing wave ratio (SWR). But a simple dipole antenna can have a good SWR at both of the frequencies that are used in the circuit. So they may be used. The only problem with it is that it is not directional.

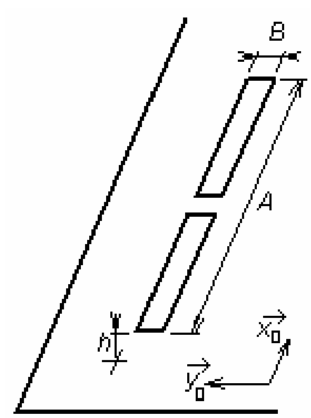


Figure 2.1 Microstrip dipole plus reflector, global view

One antenna is attached to the transmitter, and the other is connected on the front end of the receiver. At the present, the available power is a problem. More amplification would be desired either before transmission occurs or after reception of the signal. The antennas and connectors are the weak link in the circuit. Some possibilities for better antennas are a dipole with a corner reflector, a patch Yagi Uda, or a patch log periodic antenna.

The antennas were not built in this project. So they were not explained in detail.

CHAPTER THREE
AMPLIFIER CIRCUIT

3.1 Introduction

RF and microwave power amplifiers (PAs) and transmitters are used in a wide variety of applications including wireless communication, jamming, imaging, radar, and RF heating. The generation of significant power at RF and microwave frequencies is required not only in wireless communications, but also in applications such as jamming, imaging, RF heating and miniature DC/DC converters. Each application has its own unique requirements for frequency, bandwidth, load, power, efficiency, linearity, and cost. RF power can be generated by a wide variety of techniques using a wide variety of devices. Power amplifiers can be combined into transmitters in a similarly wide variety of architectures, including linear, Kahn, envelope tracking, outphasing, and Doherty. A power amplifier is a circuit for converting DC input power into a significant amount of RF/microwave output power. (Raab, 2002)

Amplification is necessary in a wireless network. The reduction in the signal due to losses during transmission, reception, and power dissipation in circuit components must be compensated by using a device to provide sufficient gain for the receiver circuit. Some real world issues when choosing an amplifier are its cost, size, and gain. The gain of the amplifier element correlates with the range that can be obtained by the link.

3.1.1 Linearity

The need for linearity is one of the principal drivers in the design of modern power amplifiers. Linear amplification is required when the signal contains both amplitude and phase modulation. It can be accomplished either by a chain of linear PAs or a combination of nonlinear PAs. Nonlinearities distort the signal being amplified, resulting in splatter into adjacent channels and errors in detection. Signals such as CW, FM, classical FSK, and GMSK (used in GSM) have constant envelopes (amplitudes) and therefore do not require linear amplification. (Raab, 2002)

3.1.2 Efficiency

Efficiency, like linearity, is a critical factor in PA design. Three definitions of efficiency are commonly used. Drain efficiency is defined as the ratio of RF output power to DC input power:

$$\eta = Pout/Pin \qquad 3.1$$

Power-added efficiency (PAE) incorporates the RF drive power by subtracting it from the output power; i.e. $(Pout - PDR)/Pin$. PAE gives a reasonable indication of PA performance when gain is high; however, it can become negative for low gains. An overall efficiency such as $Pout/(Pin + PDR)$ is useable in all situations. This definition can be varied to include driver DC input power, the power consumed by supporting circuits, and anything else of interest. (Raab, 2002)

3.1.3 Oscillation

Most amplifiers work without problems, but in some cases oscillations may occur with high frequencies/high gains and/or poor layout. Oscillations around the 3dB bandwidth of the amplifier are usually due to input/output feedback. They may cause waveform distortion and be affected by touching the amplifier on power and signal cables. This type of oscillation is usually in the hundreds of MHz or GHz range, and is frequently caused by the input or output device parasitically oscillating. To solve oscillation problems, the source of the oscillation must be located. This may be the amplifier itself or components and leads outside the amplifier.

Capacitances of pF and tens of pF from the input to ground can have a significant effect on the stability and response of the amplifier. The initial effect of adding input capacitance (sensor, leads, and connectors) is to cause a frequency peaking in the mid-band, a high frequency noise boost, and increased sensitivity to bias or ground noise. The frequency boost can be corrected by a compensating capacitance across the input stage feedback resistor, but the net effect is one of reduced bandwidth and increased noise. With larger input capacitances, this boost may result in an oscillation. The boost results in ringing after pulse edges. Clearly, input capacitance is undesirable from both performance and stability viewpoints

Parasitic oscillations are usually at high frequencies and are possibly caused by input or output devices oscillating. This can occasionally occur if the devices happen to be higher gain than normal and a high frequency resonant circuit is caused by the input/output components. Small changes to circuit lead lengths and position can stop this resonance. Occasionally, a small capacitor of a few pF can also move the oscillation. (Web_16, 2005)

3.1.4 Stability

In a two-port network, oscillations are possible if the magnitude of either the input or output reflection coefficient is greater than unity, which is equivalent to presenting a negative resistance at the port. This instability is characterized by $|\Gamma_{in}| > 1$ or $|\Gamma_{out}| > 1$, which for a unilateral device implies $|S_{11}| > 1$ or $|S_{22}| > 1$. Thus the requirements for stability are

$$|\Gamma_{in}| = \left| S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \right| < 1 \qquad 3.2$$

and

$$\left|\Gamma_{out}\right| = \left|S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_S}{1 - S_{11} \cdot \Gamma_S}\right| \langle\, 1 \qquad 3.3$$

These are defined by circles, called stability circles, that delimit $\left|\Gamma_{in}\right| = 1$ and $\left|\Gamma_L\right| = 1$ on the Smith chart. The radius and center of the output and input stability circles are derived from the S parameters (Pozar, 1998, pg. 614). The concept of instability with varying input or output matching conditions is significant, as we would desire an amplifier to be unconditionally stable under all expected conditions of source and load impedances. The example of input stability circles is shown in Figure 3.1.

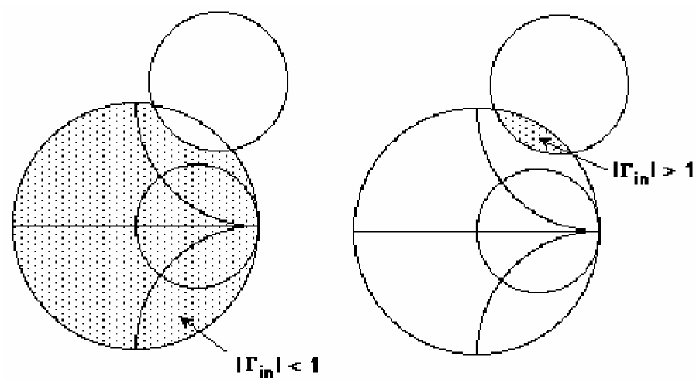


Figure 3.1 Stability circles (Raab, & the others, 2002)

This same derivation can be accomplished analytically. The conditions for stability are

$$\mathrm{K} = \frac{1 - \left|S_{11}\right|^2 - \left|S_{22}\right|^2 + \left|\Delta\right|^2}{2 - \left|S_{12} \cdot S_{21}\right|} \rangle\, 1 \qquad 3.4$$

and $\left|\Delta\right| < 1$, where $\Delta = S_{11}S_{22} - S_{12}S_{21}$, the determinant of the scattering matrix.

Calculating the Rollett Stability Factor K from the four S Parameters at each frequency will give insight into the device's stability. At frequencies where K is greater than or equal to 1, the device is unconditionally stable regardless of input or output terminations. When K < 1, the device is conditionally stable indicating that certain input or output terminations may cause the device to be unstable.

If an amplifier is conditionally stable, it can be rendered unconditionally stable by adding resistance to the input and/or output of the amplifier so that the total loop resistance at the input and output is positive. The use of resistive loading or feedback can compromise the noise performance of an amplifier unless accomplished in connection with an analysis of the amplifier noise figure. (Web_10, 2005)

3.1.5 Noise in Amplifiers

The lower limit of amplifier signal capability is set by noise. Three sources of noise in transistor amplifiers are:

1. **Thermal noise** due to random motion of charge carriers due to thermal agitation: Available noise power P_{av}=kTB, where B is the bandwidth.
2. Shot noise due to random flow of carriers across a junction, which produces a noise current of $I_{in}^2 = 2qI_{dc}B$, where q is the charge of the electron and I_{dc} is DC current.
3. Partition noise due to recombination in the junction, which produces a noise current of $I_p^2 = \frac{2 \cdot k \cdot T}{r_e'} \cdot \alpha_o \cdot (1-\alpha_o) \cdot B$

In these expressions, kT=-174 dBm in 1 Hz bandwidth, and the other parameters are elements of the transistor equivalent circuit.

The noise figure F of an amplifier is defined as the ratio of the total available noise power at the output of the amplifier to the available noise power at the output that would result only from the thermal noise in the source resistance. Thus F is a measure of the excess noise added by the amplifier. Amplifier noise can also be characterized by an equivalent noise temperature of the source resistance that would provide the same available noise power output. This equivalent noise temperature is given by

$$T_e = (F - 1)T_o \qquad 3.5$$

It is unusual for noise figure and gain circles to be concentric, as maximum gain conditions are not the same as minimum noise figure conditions. (Web_10, 2005)

3.2 Design of Microwave Power Amplifiers Using S Parameters

Microwave amplifiers combine active elements with passive transmission line circuits to provide functions critical to microwave systems and instruments. The history of microwave amplifiers begins with electron devices using resonant or slow-wave structures to match wave velocity to electron beam velocity.

The design techniques used for BJT and FET amplifiers employ the full range of concepts we have developed in the study of microwave transmission lines, two-port networks and Smith chart presentation.

The development of S-parameter matrix concepts grew from the need to characterize active devices and amplifiers in a form that recognized the need for matched termination rather than short or open circuit termination.

The magnitude and phase angle of each of the S parameters typically vary with frequency, and characterization over the complete range of interest is necessary. (Web_10, 2005)

3.2.1 Power Amplifier Design

Amplifiers offer the RF designer multi-stage performance in a package which looks like a discrete transistor. Amplifiers advantages of wide bandwidth, impedance match, and a choice of gain and output power levels result from their being monolithic circuits which contain InGaP HBTs (indium gallium phosphide heterojunction bipolar transistors).

The internal circuit configuration is a Darlington pair, embedded in a resistor network as shown in the schematic diagrams, Figure 3.2. Like a discrete bipolar transistor, this type of circuit is current- rather than voltage-controlled. This means that for a range of current around a recommended value, the device voltage varies much less than in proportion to current. A constant-current DC source would be ideal for providing a stable operating point. By contrast, if a constant-voltage DC source were used the current would vary widely with small changes in supply voltage, temperature change, and device-to-device variations; such operation is not recommended. Practical applications typically use a series resistor between the Amplifier and a source of DC voltage to approximate constant current. (Web_6, 2005)

Amplifier is a 2-port device: RF input, and combined RF output and bias input. It has 4 leads including 2 ground leads; connecting both of them to external ground will minimize common path impedance for best RF performance. Internal resistors in Figure 3.2 determine the DC operating point of the transistors and provide feedback to set RF gain, bandwidth, and input and output impedances to optimum values.

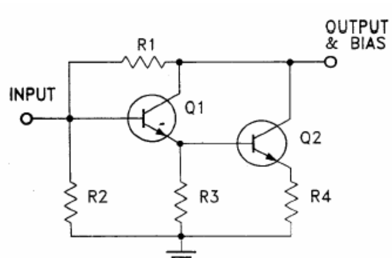


Figure 3.2 Schematic diagram (Web_6, 2005)

The design of power amplifiers involves less emphasis on noise parameters, and more emphasis on linearity and intermodulation, as well as efficiency and thermal considerations. To design a power amplifier, one must use large-signal S-parameters and be aware of nonlinear effects.

Where careful design of the input matching network is required to realize the full capabilities of low-noise amplifiers, in power amplifiers more emphasis tends to be on optimizing the output matching network. There are, however, special problems associated with the very low input impedance that can be found in bipolar power devices, which require special treatment in the input matching network if wideband operation is to be achieved. (Web_10, 2005)

3.2.2 Device Grounding

The entire stability picture changes when the device is mounted on a microstrip circuit board. The device is elevated above electrical ground by placing the device on the top side of the microstrip board and using plated through holes to attach each emitter lead to ground. High frequency PHEMT devices generally require very good grounds because of their very small geometry and very high frequency gain. Most bipolar devices are more tolerant of greater inductance in series with the emitter grounding.

3.2.3 Bias Circuits and Bias Circuit Instabilities

Once the microwave amplifier is designed, it remains to provide the dc bias voltages and currents required for the active device. This is no simple problem, as the arrangements to introduce the biases can disturb the microwave circuit. Generally, high impedance microstrip traces can be used as decoupling inductors, but caution must be exercised not to create a low frequency oscillator circuit in the bias network.

A common cause of trouble is the use of an inductor with a large bypass capacitor, which can create a resonator in the MHz region that can support oscillation of the active element, which will have very high gain at lower frequencies.

Bias-circuit instabilities are a common source of problems in amplifiers and other active circuits. These generally result from the use of inductors and capacitors in the bias circuit without regard to resonances or situations where 180° phase shift can occur.

Two examples of circuit configurations that can promote bias oscillations at low frequencies are:

• The use of an inductor with bypass capacitors on both sides to filter dc supply to earlier stages of an amplifier; this can have 180° phase shift at a video frequency where the active element has substantially more gain than at microwave frequencies.

• The use of an inductor with bypass capacitor to isolate dc supply input to the base or gate of the active element; this can form a resonator for video frequency oscillations. (Web_10, 2005)

A practical biasing configuration is shown in Figure 3.3. Bias current is delivered from a voltage supply Vcc through the resistor Rbias and the RF choke (inductor), shown as RFC in the figure. The resistor reduces the effect of device voltage (V_c) variation on the bias current by approximating a current source.

Blocking capacitors are needed at the input and output ports. They should be of a type having low ESR (effective series resistance), and should have reactance low enough not to affect insertion loss or VSWR adversely at low frequency. The blocking capacitors must be free of parasitic resonance up to the highest operating frequency. Use of a bypass capacitor at the connection to the DC supply is advised to prevent stray coupling to other signal processing components.

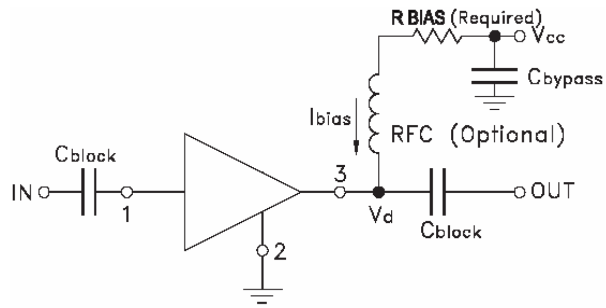


Figure 3.3 Amplifier schematic from MiniCircuits (Web_6, 2005)

In this circuit, DC blocking capacitors are added at the input port (pinnumber 1 on the packaged amplifier) and at the output port (pin 3). Bias current is given by the equation:

$$I_{bias} = (V_{cc} - V_d) / R_{bias} \qquad 3.6$$

Table 3.1 lists the values of the bias resistor needed with several values of supply voltage for each of the Amplifier models. These values take into account the variation of device voltage, both lot-to-lot and with temperature (-45 to 85 ^{o}C ambient). Also, they are chosen from the readily available “1%” resistor values. The table lists the power dissipated by the bias resistor in the 12 Volt case, as an example. The greater the difference between the supply and device voltage, the easier it is to maintain constant operating conditions; this is discussed in detail later.

The bias current values in Table 3.1 are the recommended values. Exceeding the maximum current values shown there, however, could cause excessive junction temperature and premature failure. Substantially lower currents, while not degrading device reliability, could cause unpredictable RF performance because of non-optimum internal operating points. (Web_6, 2005)

Table 3.1 Bias resistor values for various supply voltages (Web_6, 2005)

Improved ERA Model	ERA-1	ERA-2	ERA-21	ERA-3	ERA-33	ERA-4	ERA-50	ERA-51	ERA-6
Bias Current	40 mA	40 mA	40 mA	35 mA	40 mA	65 mA	60 mA	65 mA	70 mA
Device Voltage (nom.)	3.4V	3.5V	3.5V	3.2V	4.3V	4.5V	4.4V	4.5V	4.9V
Supply Voltage	Bias Resistor (ohms) at Supply Voltage								
7	90.9	88.7	88.7	107	69.8	38.3	47.5	40.2	30.1
8	113	113	113	133	93.1	52.3	63.4	53.6	43.2
9	137	137	137	162	115	66.5	78.7	68.1	56.2
10	162	162	162	191	140	80.6	95.3	82.5	69.8
11	187	187	187	221	165	95.3	113	97.6	84.5
12	215	215	210	249	191	110	127	113	97.6
13	237	237	237	280	215	127	143	127	113
14	261	261	261	309	243	143	162	143	127
15	287	287	287	340	267	158	178	158	140
16	309	316	316	365	287	174	196	174	154
17	332	340	340	392	316	187	210	191	169
18	357	365	365	422	340	205	226	205	182
19	383	392	392	453	365	221	243	221	196
20	412	412	412	475	392	237	261	237	210
Nominal dissip. in Resistor for 12V	0.34 W	0.34 W	0.34 W	0.31 W	0.31 W	0.51 W	0.45 W	0.50 W	0.52 W

The list below summarizes a few main points about microwave/RF bias networks

1. One has to look at the performance of the circuit at all frequencies. It seems to be a common mistake to ignore the circuit's performance at other frequencies than the one that will be used.

2. At some point in the biasing circuit one should terminate all of the frequencies good enough that they don't have to worry about what the impedance of the power supply circuitry is. This means at some point there should be isolation by using a high series impedance or a low shunt impedance, ect .

3. One should include their biasing circuits as part of their RF circuit's simulations.

4. One should use caution when attempting to create broad band short using different size capacitors in shunt.

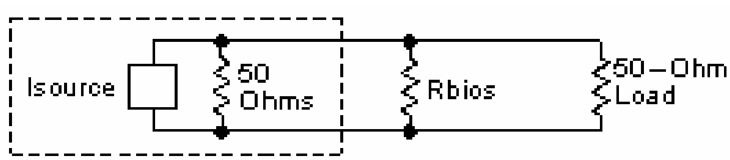
3.2.4 Advantage of an RF Choke

The advisability of using an RF choke in series with Rbias is evident from the following analysis. Figure 3.4 shows an equivalent circuit of the output of an Amplifier as a current source in parallel with an internal 50-ohm source resistance, loaded by both R_{bias} and an external 50-ohm load. The current in the 50-ohm load is:

$$I_L = I_{source} \times R_{bias} / (2\,R_{bias} + 50) \qquad 3.7$$

and the loss in power gain relative to not having the output loaded by R_{bias} is:

$$L = 20 \log [(2\,R_{bias} + 50) / 2\,R_{bias}]\ \mathrm{dB} \qquad 3.8$$



Equivalent ERA output circuit

Figure 3.4 Effect of bias resistor on the output, without an RF choke (Web_6, 2005)

Suppose, for example, that the Amplifier ERA-3 is used with a 12V supply without a choke. From the above expression, the effect of the 249 ohms bias resistor (from Table 3.1) is found to be a 0.8 dB reduction in the gain of the amplifier.

An RF choke should be chosen such that its reactance is at least 500 ohms (10 times the load impedance) at the lowest operating frequency. It must also be free of parasitic (series) resonance up to the highest operating frequency.

The circuit designer might consider using a commercially available inductor as the RF choke in the bias circuit of Figure 3.3. The low end of the useful frequency range is controlled by the value of the inductance; the higher the value, the lower the frequency. The high end of the frequency range is determined by the series resonant frequency of the inductor; it tends to decrease as the value of the inductance increases. (Web_6, 2005)

3.2.5 Higher Bias Resistor Values Reduce Variation

Increasing the supply voltage allows a higher value of R_{bias} to be used as shown in Table 3.1, and that reduces the variation in bias current. The benefit is that RF performance, especially the 1-dB compression point, is made more constant. The following quantifies this effect.

The device voltage V_d is a function of both Ibias and temperature T. Device voltage increases with bias current, and the variation can be expressed as a rate of change ΔV_d / ΔI_{bias}, in mV per mA. Device voltage decreases with increasing temperature, as ΔV_d / ΔT millivolts per degree C. (Web_6, 2005)

3.2.6 Minimizing Power Dissipation

In addition to bias current stability, stability of power dissipation of the Amplifier device is favored by using a high V_{cc} value. This is because of the negative temperature coefficient of device voltage V_d. In particular, if V_{cc} is at least two times V_d, then P_D, the amplifier power dissipation decreases with increasing temperature, as shown by the following analysis:

$$P_D = V_d \cdot I_{bias} = V_d \, (V_{cc} - V_d) / R_{bias} \qquad 3.9$$

Taking the derivative of P_D with respect to V_d and setting it equal to zero, we find that the maximum value of P_D occurs when $V_d = V_{cc} / 2$.

Gain increases only slightly with increasing current. Less change occurs in the other characteristics – flatness, VSWR, and noise figure and in most cases it is insignificant for practical application. (Web_6, 2005)

3.3 The ERA-3SM Amplifier

For the wireless LAN circuit, the ERA-3SM amplifier from MiniCircuits is used. It provides about 15 dBm gain at the two operating frequencies.

To provide for correct functionality, the choice of circuit components is critical. A schematic illustrating the necessary components is shown in Figure 3.3. A proper bias current is achieved by placing a resistor on the DC bias branch of the circuit. The amplifier also requires DC blocking capacitors at its input and output to allow only the high frequency signals to enter the passive circuitry. Surface mount components are used to reduce the inductive effects produced by long leads in high frequency applications.

DC blocking capacitors are necessary on the input and output of the amplifier to block DC voltages and allow the RF signal to pass. The data sheets specify that the capacitors should have a low effective series resistance (ESR) and should be free of parasitic (parallel) resonance up to the highest operating frequency. The value of the capacitance is not specified in this case, but 100 pF seems to be a very common value for this type of application. With some searching, capacitors can be found that have a low ESR and a parallel resonance that is above 2.6 GHz. (Web_4, 2005)

pin connections

PORT	cb
RF IN	1
RF OUT	3
DC	3
CASE GND	2,4
NOT USED	—
DEMO BOARD	ERA-TB

Figure 3.5 ERA-SM and its pin connections (Web_13, 2005)

Using a 12V DC supply, a bias current of 35 mA is desired for proper operation. For this current, the resistor used in this cricuit has a value of 249 ohms and power dissipation of 311 mW.

To improve the gain of the amplifier, an inductor can be used in the bias branch to reduce the amount of RF power that is lost in the branch. The value of the inductor is chosen to be larger than ten times the load impedance (in this case, the load impedance is the characteristic impedance, Zo = 50 ohms). The inductor must also be "free from parasitic (series) resonance up to the highest operating frequency." An inductor with a value of 47 nH is used in the circuit. The use of a bypass capacitor is also desirable in the bias branch to approximate a steady input current. The value of the capacitor is 100 pF. The layout of the amplifier circuit with component values as labels is shown in Figure 3.6.

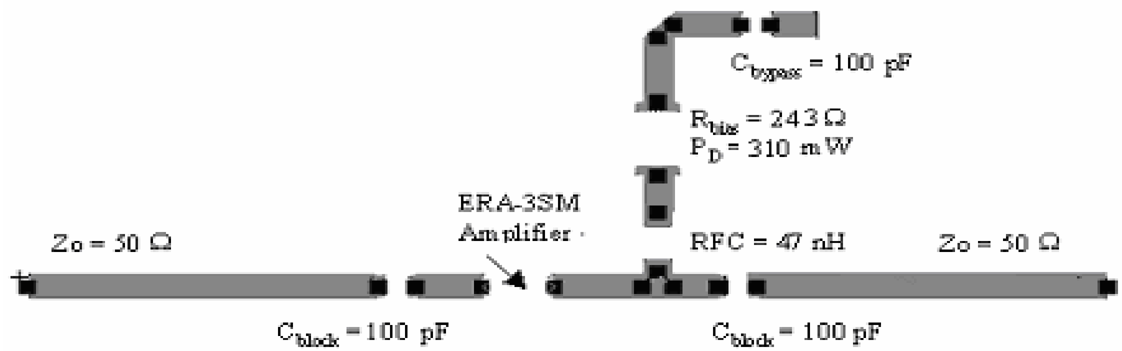


Figure 3.6 Amplifier layout

An important issue when using RF amplifiers is the issue of grounding. It is very important to ensure that ground is placed near to the amplifier to reduce the high-frequency effects.

When preparing the layout for the circuit, spacing for the amplifier and other components must be allotted to allow room for soldering the components. This is done in the AWR simulation software. The artwork for a transistor can be used to prepare space for the amplifier since there is no amplifier layout in the libraries. Space for other components is allotted by using lumped elements with artwork .

The case sizes for each of the components can be found on their respective data sheets. Some unit conversions may be necessary depending on the values that are specified. When the circuit design is complete in the Schematic window, the circuit is synchronized and can be viewed in the Layout window where the sizes can be verified, and the overall circuit size is noted.

The circuit can be tested in two ways. The first method is to use the Network Analyzer to view the overall performance of the amplifier. The amplifier gain can be seen over the frequency range of interest after calibrating the Network Analyzer. Another method is to test the circuit using the Signal Generator on the input of the amplifier and the Spectrum Analyzer on the output to see the gain at discrete frequencies. The same circuit holder that is used for TRL measurements on the network analyzer is used to connect the microstrip circuit to the coaxial lines for the input and the output of the circuit. For amplifiers with a gain higher than 15 dBm, the network analyzer does not display the correct information for the gain, and the signal generator and spectrum analyzer must be used. This is especially true when multiple amplifiers are cascaded to produce higher gain circuits.

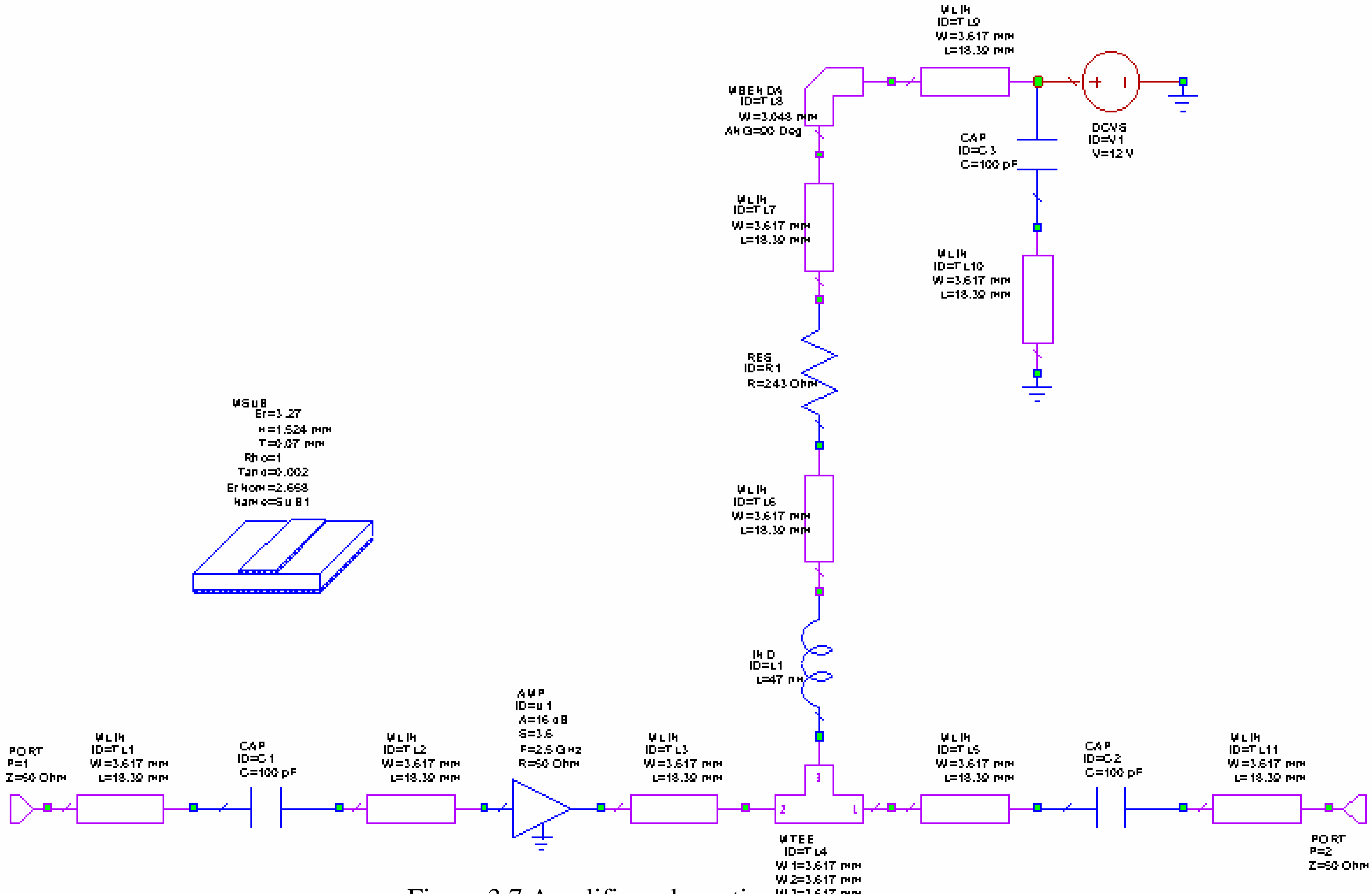


Figure 3.7 Amplifier schematic

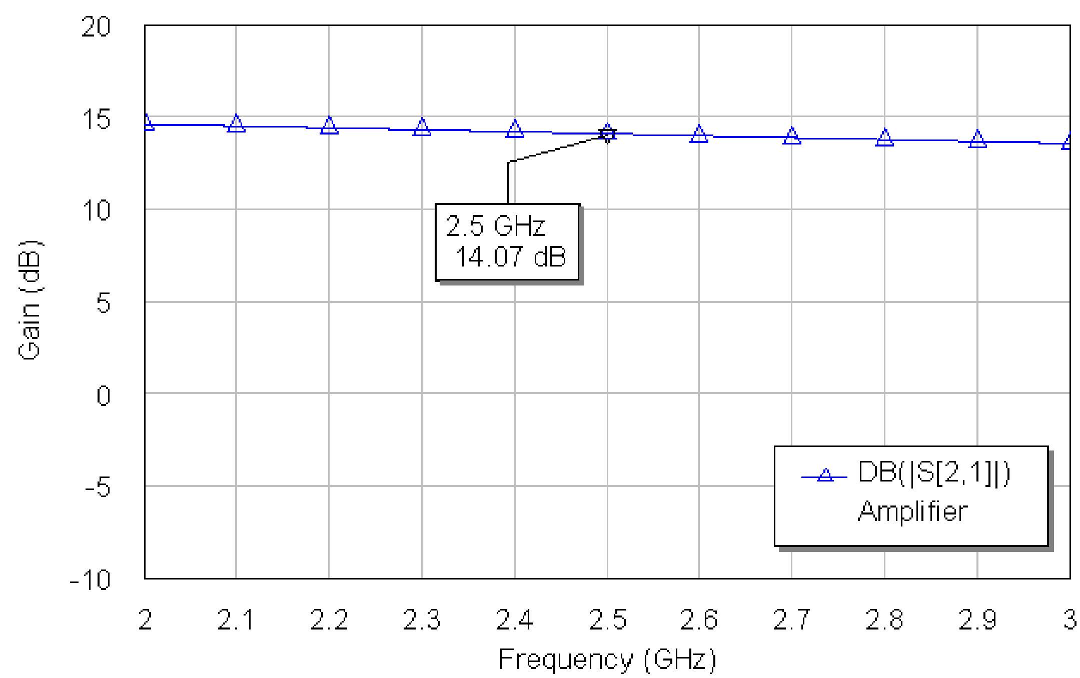


Figure 3.8 Gain versus frequency for amplifier

Consequently, amplification is necessary in a wireless network. The reduction in the signal due to losses during transmission, reception, and power dissipation in circuit components must be compensated by using an amplifier to provide sufficient gain for the receiver circuit. Some real world issues when choosing an amplifier are its cost, size, and gain. The gain of the amplifier element correlates with the range that can be obtained by the link. So the ERA-3SM amplifier from MiniCircuits that has gain of 14 dBm was selected. As it is shown in Figure 3.8, AWR simulation result of S_{21} parameter of amplifier circuit proves that we obtained gain of 14 dBm at 2.5 GHz.

CHAPTER FOUR
POWER DIVIDER CIRCUIT

4.1 Introduction

Power dividers and combiners are used in microwave frequencies because of the different behavior of circuits. In microwave frequencies, the modeling of the circuits is different from the expected lumped model. Therefore, power dividers in microwave frequencies are not just resistive power dividers as in the low frequency or DC model. The important issues while using power dividers are the specifications or properties of the dividers. In microwave frequency applications, for example power can be reflected from the output ports. Then to avoid the reflected power expected power divider should satisfy the maximum power transfer theory, which says the impedance of the load should be the complex conjugate of the impedance seen from the load end when load is not connected. This is called "matching" of the ports. This leads to transferring maximum power to the load and because of no reflection, avoid burning out the power supply. Also, again for maximally transfer the power to the output ports, the network should not dissipate power and this property is called "losslessness". (Web_11, 2003)

4.2 Definitions

Coupling and Insertion Loss: The total transmission loss within a power divider is the sum of the coupling loss and the insertion loss. These are defined as follows:

- *Coupling Loss* - Expressed in dB, coupling loss is the power level that would be expected at each of n-ports when the input power level is equally split n-ways exclusive of other losses. See Table 4.1 for coupling loss versus quantity of output ports.

- *Insertion Loss* - The net unrecoverable power dissipated within the circuit.

Isolation: Expressed in dB, isolation is the magnitude of power leakage from one output port to another expressed as the signal amplitude difference between a given port and a reference port under matched source and load impedance conditions.

The maximum isolation achievable with a power divider strongly depends on the impedance match presented to the input port by the external signal source. (Web_5, 2003)

4.3 Reciprocal and Lossless Networks

One important property is reciprocity. A network is reciprocal if the power transfer and the phase do not change when the input and output are interchanged. This means that for reciprocal networks, the scattering matrices are symmetrical. In order for a network to be reciprocal, it has to be linear, time invariant, made of reciprocal materials, and there cannot be any dependent voltage or current sources in the network..

Reciprocal circuits have a symmetrical scattering matrix, which means that $S_{ij} = S_{ji}$. For example, the scattering matrix of a reciprocal two-port looks like

$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{21} \\ S_{21} & S_{22} \end{bmatrix} \qquad 4.1$$

Another simplification that can be made in a scattering matrix is when the network is lossless, which means it absorbs no power. This means that the scattered power is equal to the incident power, or mathematically (where † is transpose sign)

$$b^{\dagger}\, b = a^{\dagger}\, a \qquad 4.2$$

It turns out that this is equivalent to saying that the scattering matrix is unitary, which is written as

$$S^{\dagger}\, S = I \qquad 4.3$$

where I is the identity matrix.

Using S-parameters offers some advantages. S-parameters are important in microwave design because they are easier to measure and work with at high frequencies than other kinds of parameters. They are conceptually simple, analytically convenient, and capable of providing a great insight into a measurement or design problem. (Web_18, 2003)

4.4 Three-Port Networks

There are many applications when one might want to use a network with more ports for example, if there is a need to split the power in one transmission line or waveguide into several others, or combine the power from several lines into one. One application was the waveguide T used in power measurements. Now it will be looked at some important properties and examples of microwave networks with 3 and 4 ports. If we wish to make a network that will have three ports and will be used as a two-way power splitter, we would like this network to be lossless and matched at all ports. Such a circuit is also reciprocal if it is passive and contains no material anisotropy. From the reciprocity condition, we know that the scattering matrix has to be symmetrical, and from the matched condition we know that all three reflection coeficients at the three ports are zero.

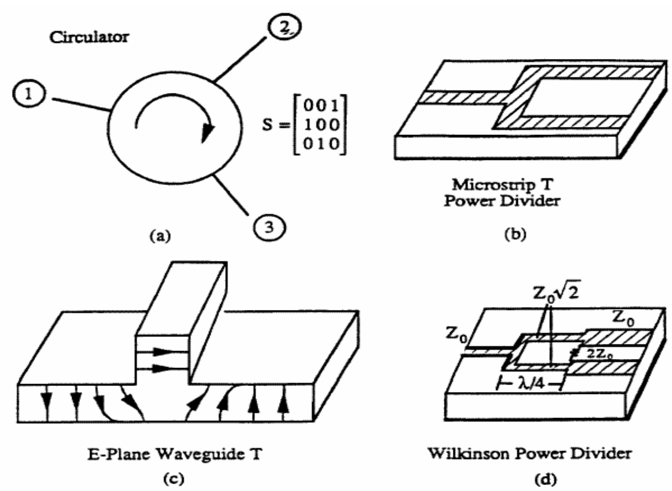


Figure 4.1 The symbol and scattering matrix for a circulator (a), and examples of lossless (b), (c) and resistive (d), power dividers. (Web_18, 2003)

One device that has such scattering parameters is called a circulator, and its symbol is shown in Fig. 4.1(a). It has the property that power coming into port 1 will go out of port 2, and none will go out of port 3, power going into port 2 will only go out port 3, and power going into port 3 will only go out port 1. It is now obvious where the name comes from.

If a three port is reciprocal and lossless, but not necessarily matched at all ports, we get a power divider, such as the ones shown in Fig. 4.1(b),(c). Finally, if we relax the lossless condition, we get a resistive power divider, an example of which is shown in Fig. 4.1(d). Such resistive power dividers can be made such that $S_{23} = S_{32} = 0$ so that the two output ports are isolated. They can also be designed to have more than 2 outputs. (Web_18, 2003)

4.5 Power Dividers

Power dividers are necessary to split or combine electrical signals. While the specific division can be arbitrary, equal splitting is the most common type of divider implemented. In high frequency applications, the designs are based on planar transmission lines that must provide the same functionality as low frequency resistive network designs.

Ideally, a power divider would be matched at all ports, lossless, and reciprocal. Is this possible?

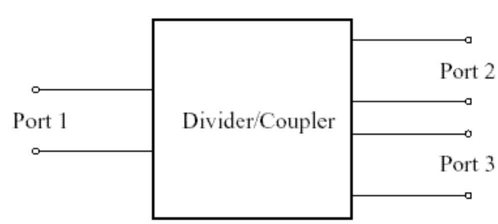


Figure 4.2 Three port microwave network for used as a power divider (or combiner). (Web_17, 2003)

Consider a 3-port device with S-matrix

$$S = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{12} & 0 & S_{23} \\ S_{13} & S_{23} & 0 \end{bmatrix} \qquad 4.4$$

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix} \qquad 4.5$$

Suppose all 3 ports are matched so that $S_{ii} = 0$. Also assume the network is reciprocal so $S_{ij} = S_{ji}$. Equation 4.4. If the network is lossless, then

$$\begin{aligned} |S_{12}|^2 + |S_{13}|^2 &= 1 \qquad & S_{13}S_{23}^* &= 0 \\ |S_{12}|^2 + |S_{23}|^2 &= 1 \qquad & S_{12}S_{13}^* &= 0 \\ |S_{13}|^2 + |S_{23}|^2 &= 1 \qquad & S_{12}S_{23}^* &= 0 \end{aligned} \qquad 4.6$$

The second column shows that 2 of the 3 unique S-parameters must be zero. But, if both of them are zero, one of the equations in the first column will be violated. *So: It is impossible to have a lossless, matched, reciprocal 3 port device.* Relaxing any constraint makes it possible for the other two constraints to be satisfied, however. The possibilities are a network that is lossless and reciprocal, but not matched; reciprocal and matched but lossy; and matched/lossless using a nonreciprocal material. (Web_17, 2003)

There are three types of dividers:

1. Wilkinson
2. T-Junctıon
3. Split-Tee

4.5.1 The T-Junction Power Divider (Three-Port Network)

A junction between two dissimilar width sections also introduces a large discontinuity. A standard T-junction power divider is shown in Figure 4.3.

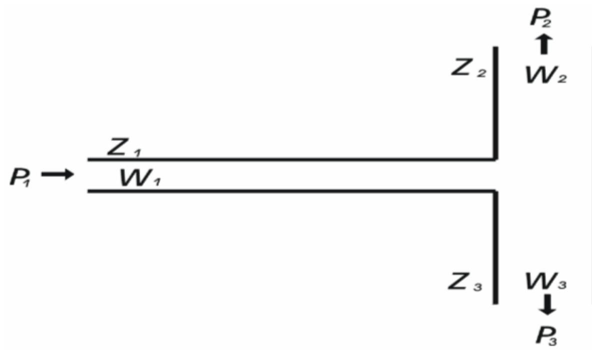


Figure 4.3 Diagram of T-junction power divider. (Web_12, 2003)

In this diagram, the input power P_1 is delivered to the intersection on a microstrip of width W_1 and impedance Z_1. The line then branches into two arms with power, width and impedance given by P_2, W_2, Z_2 and P_3, W_3, Z_3 respectively. The design equations for this divider are

$$Z_1 = \frac{Z_2 \times Z_3}{Z_2 + Z_3}, \quad P_2 = \left(\frac{Z_1}{Z_2}\right) P_1, \quad P_3 = \left(\frac{Z_1}{Z_3}\right) P_1 \qquad 4.7$$

This simplest type of matched T-junction is the lossless 3dB power divider. It can be seen from the equations above that if $Z_2 = Z_3 = 2Z_1$ the power will split evenly into the arms of the T with each arm having half the original power. It is interesting to note that the impedances of the two arms act just like resistors wired in parallel. To match the impedances of the arms of the T to the impedance of the base, the arms must have twice the impedance of the base. (Web_12, 2003)

4.5.2 Impedance Mismatches Quarter-Wave Transformer

A general mismatch in impedance between two points on a transmission line can be compensated with a quarter-wave transformer.

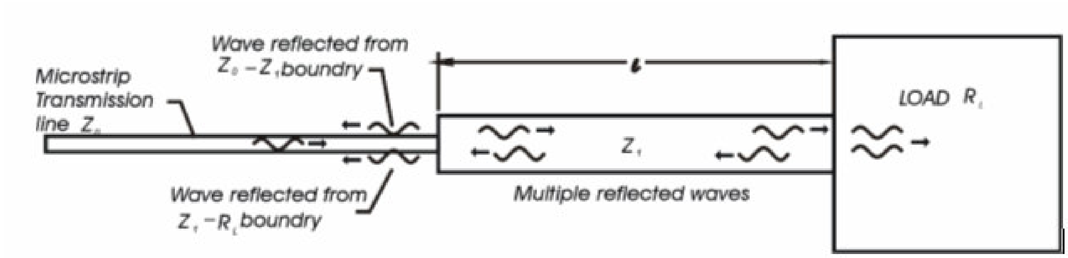


Figure 4.4 Diagram of quarter-wave impedance transformer showing multiple reflections. (Web_12, 2003)

In a quarter-wave transformer, we want to match a load resistance R_L to the characteristic feedline impedance Z_0 through a short length of transmission line of unknown length and impedance Z_1. The input impedance looking into the matching section of line is given by;

$$Z_{in} = Z_1 \frac{R_L + jZ_1 \tan \beta l}{Z_1 + jR_L \tan \beta l} \qquad 4.8$$

If we choose the length of the line = λ / 4 then β.l = (2π / λ).(λ / 4) = π / 2, divide through by tanβ and take the limit as $\beta l \longrightarrow \pi / 2$ to achieve

$$Z_{in} = \frac{Z_1^2}{R_L} \qquad 4.9$$

For a perfect transition with no reflections at the interface between microstrip and load, Γ =0 so $Z_{in} = Z_0$ and this gives us a characteristic impedance

$$Z_1 = \sqrt{Z_0 R_L} \qquad 4.10$$

which is the geometric mean of the load and source impedances. With this geometry, there will be no standing waves on the feedline although there will be standing waves on the λ/4 matching section. Why was the value of l = λ/4 chosen? In fact, any odd multiple (2n + 1) of l = λ/4 will also work.

Power dividers/combiners fulfill two complementary roles. As dividers, these devices take a single input signal and split its power equally between two or more outputs. The phase difference between all outputs is extremely small. As a combiner, these devices take two or more inputs and produce the vector sum at the output. Table 4.1 shows the coupling losses of reactive hybrid power dividers assuming no power loss. (Web_12, 2003)

Table 4.1 Ideal coupling loss (Web_12, 2003)

Quantity of Outputs	Coupling (dB)
2	– 3
3	– 4.78
4	– 6
5	– 7
6	– 7.78
8	– 9
16	– 12
N	-10logN

4.5.3 Designing a Power Splitter

The power splitter (3-dB coupler) is the next sequential component in the circuit. The input of the circuit needs to be matched to the characteristic impedance of 50 ohms. Equal power is desired on both branches of the output circuit. As an optional constraint, each branch of the circuit can be tuned to the frequency that will eventually be detected on that branch, 2.4 and 2.6 GHz, by using quarter wave transformers.

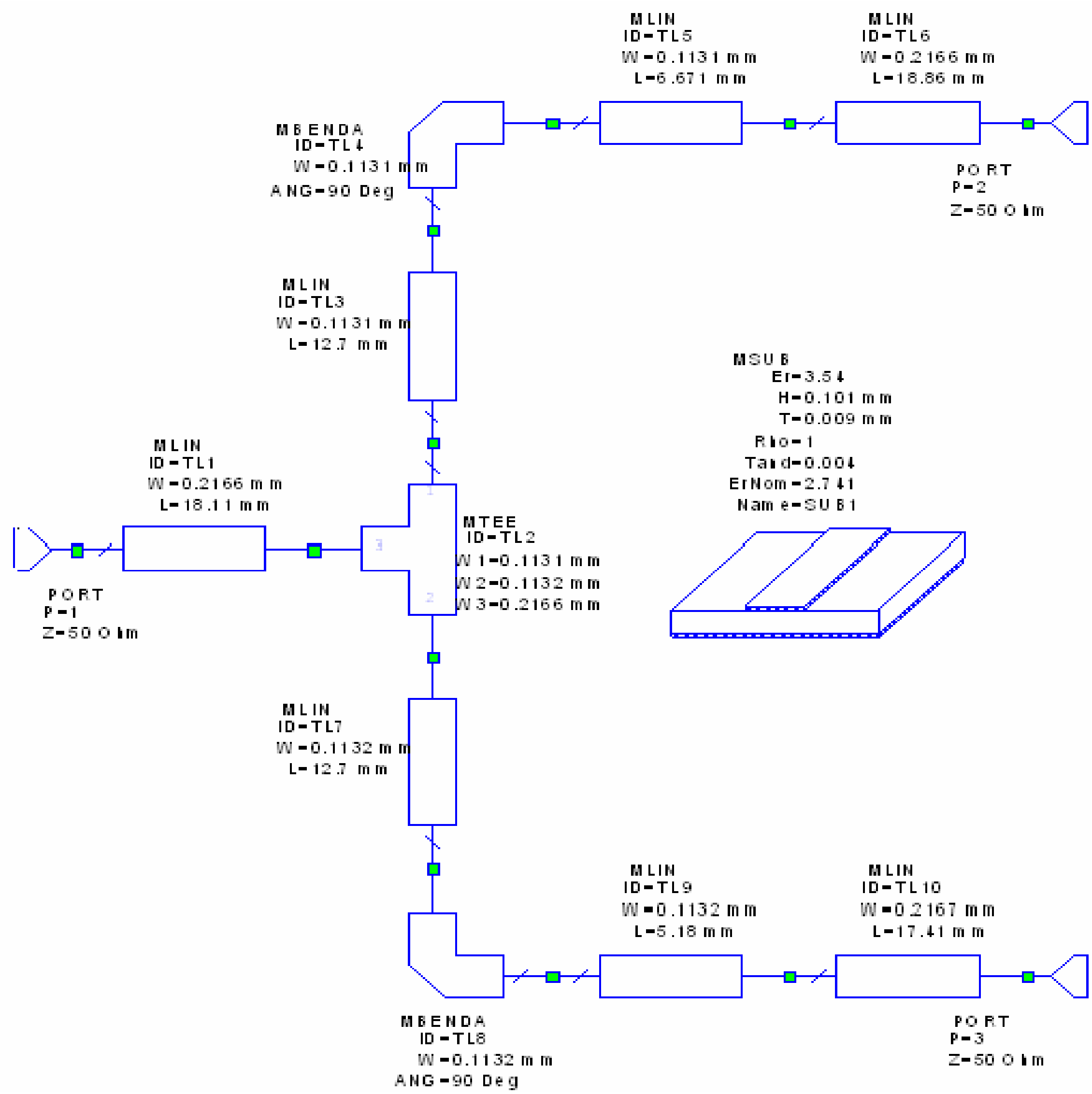


Figure 4.5 Schematic diagram of the power divider

A schematic of the power divider is shown in Figure 4.5. The input line of the power divider is 50 ohms. This value is obtained by a line with the proper width as dictated by using TX-Line in the AWR-Microwaveoffice software package. For effective matching, the equivalent impedance of the two parallel lines of the output should also be 50 ohms. The parallel combination of two 100 ohm lines will produce the proper matching and power division. Since the eventual output of the power splitter will also need to be matched to 50 ohms, a logical choice is the use of quarter wave transformers. The length of the 100 ohm line can be zero since a quarter wave transformers can be placed on each branch immediately after the input line is split. The impedance for the quarter wave transformer is found by taking the square root of the 100 ohm impedance times the 50 ohm impedance producing 70.71 ohms. The length and width of the microstrip lines is found by using TX-Line in the AWR-Microwaveoffice with an electrical length of 90 degrees (one quarter wavelength) at each respective frequency and a characteristic impedance of 70.71 ohms. (Pozar, 1998, p. 359-362).

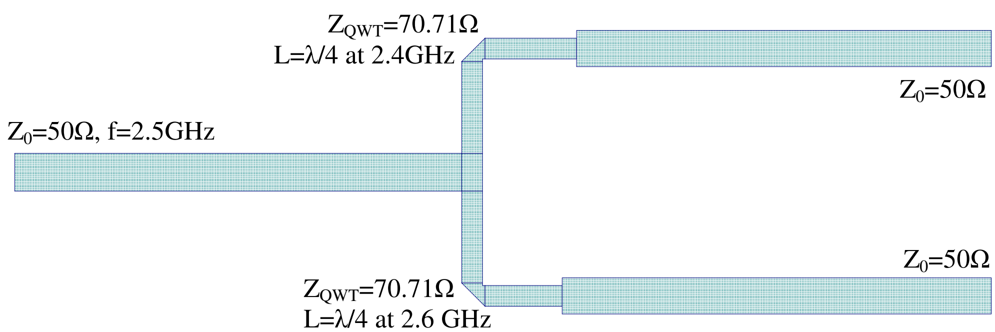


Figure 4.6 Dual frequency T-Junction power divider layout

Properties of substrate selected:

ε_r = 3.54, H = 0.10099mm, T = 0.009mm, tand = 0.004

where, ε_r is the dielectric constanat, H is the height of substrate metarial, T is the thickness of conductor, tand is the loss tangent

Width and length values of each microstrip lines are calculated according to impedance and frequency values. Properties of substrate are important to calculate. The length and width of the microstrip lines is found by using TX-Line in the AWR-Microwaveoffice at each respective frequency and its characteristic impedance. In this way;

For Z_0 = 50Ω, f = 2.5GHZ; W = 0.21662mm, L = 18.107mm
For Z_{QWT1} = 70.71 Ω, f = 2.4GHz; W = 0.11313mm, L = 19.371mm
For Z_{QWT2} = 70.71 Ω, f = 2.6GHz; W = 0.11316mm, L = 17.88mm
For Z_0 = 50 Ω, f = 2.4GHz; W = 0.2166mm, L = 18.862mm
For Z_0 = 50 Ω, f = 2.6GHz; W = 0.21664mm, L = 17.409mm

where; Z is the impedance, W is the width of microstrip line, L is the length of microstrip line.

The circuit can be simulated using the test AWR-Microwaveoffice. A frequency sweep is desired from about 2-3 GHz. The parameters of interest are S_{11}, S_{21}, and S_{31}. The simulation for the reflection coefficient, S_{11}, is shown in Figure 4.7. This parameter is a measure of how well the input of the power splitter is matched. Plots of the two transmission coefficients, S_{21} and S_{31}, are shown in Figure 4.8. The responses can be tuned by shortening or lengthening the quarter wave sections until a low (less than or equal to –20 dB) reflection coefficient is attained (it's –36.9 dB as shown in Figure 4.7) and transmission coefficients with values near –3 dB are attained (it's about –3 dB as shown in Figure 4.8). The layout for the power divider is shown in Figure 4.6.

Testing is performed on the circuit using the Network Analyzer. The parameters for S_{11} and S_{21} can be measured. When this measurement is performed, the third port should have a matching load placed on it so that there will not be any reflections due to an unmatched termination. The circuit can then be rotated 180 degrees, and the value for S_{31} can be attained by placing the matched load on the port that is not connected to the network analyzer.

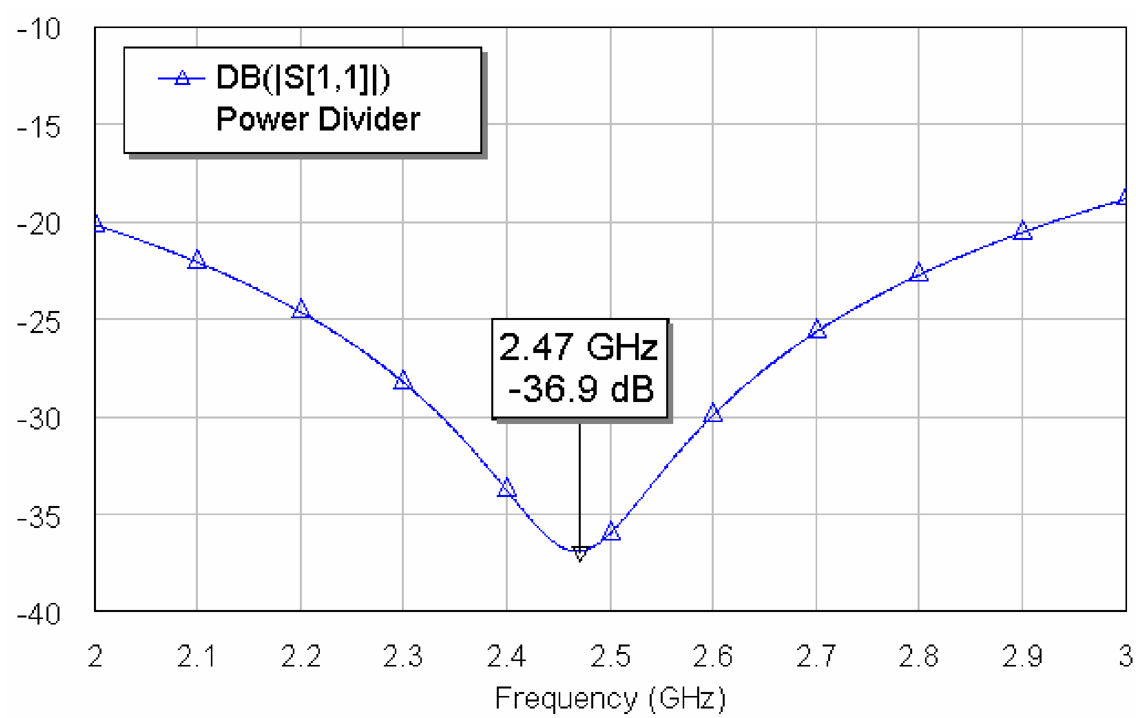


Figure 4.7 Simulation result of S_{11} parameter for power divider

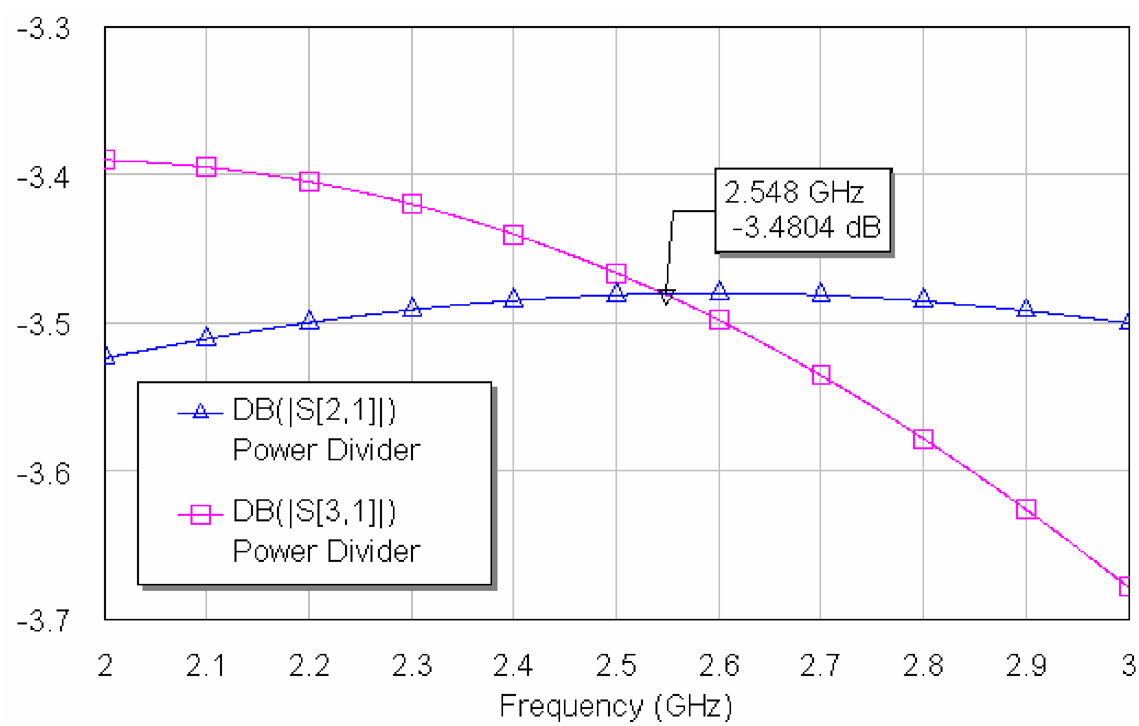


Figure 4.8 Simulation result of S_{31} and S_{21} for power divider

CHAPTER FIVE
COUPLED LINE FILTER CIRCUIT

5.1 Introduction

The rapid growth of wireless communications has placed an increasing demand for the design techniques for a planar bandpass filters to meet the various needs such as size reduction, performance, cost requirement, and so on. For this purpose, there has recently been increasing interest in a coupled-line bandpass filter with various configurations. (Yun & the others, 2000)

Filters of some sort are essential to the operation of most electronic circuits. It is therefore in the interest of anyone involved in electronic circuit design to have the ability to develop filter circuits capable of meeting a given set of specifications. Unfortunately, many in the electronics field are uncomfortable with the subject, whether due to a lack of familiarity with it, or a reluctance to grapple with the mathematics involved in a complex filter design. (Web_1, 2004)

In many applications, especially including satellite and mobile communications, keeping RF filter structures to a minimum size, weight, low insertion loss, high selectivity together with linear phase of flat group delay in pass band is very important. Recent advance in high-temperature superconduting (HTS) circuits, low-temperature cofired ceramic (LTCC) circuits and microwave monolithic intergrated circuits (MMIC) has additionally stimulated the development of various planar filters, especially narrowband bandpass filters which play an important role in modern communications system. (Hong & Lancaster, 2001)

This project describes the design and fabrication of microwave bandpass filter by using microstrip layout. A microwave filter is a two port network used to control the frequency response within a system by allowing the transmission of certain frequencies in the passband while attenuating frequencies in the stopband. The development of the microstrip filters are simulated by using (MWO) Microwave Office simulator software.

5.2 Fundamentals of Filter

In circuit theory, a filter is an electrical network that alters the amplitude and/or phase characteristics of a signal with respect to frequency. Ideally, a filter will not add new frequencies to the input signal, nor will it change the component frequencies of that signal, but it will change the relative amplitudes of the various frequency components and/or their phase relationships. Filters are often used in electronic systems to emphasize signals in certain frequency ranges and reject signals in other frequency ranges. Such a fitler has a gain which is dependent on signal frequency. A band-pass filter passes signals falling within a relatively narrow band of frequencies and attenuates signals outside of that band.

There are five basic filter types (notch, low-pass, high-pass, all-pass, and bandpass).

Notch or Band-Reject: A filter with effectively the opposite function of the bandpass is the band-reject or notch filter.

Low-Pass**:** A low-pass filter passes low frequency signals, and rejects signals at frequencies above the filter's cutoff frequency.

High-Pass:The opposite of the low-pass is the high-pass filter, which rejects signals below its cutoff frequency.

All-Pass or Phase-Shift: Filter response type has no effect on the amplitude of the signal at different frequencies. Instead, its function is to change the phase of the signal without affecting its amplitude. This type of filter is called an all-pass or phase-shift filter.

Bandpass: The number of possible bandpass response characteristics is infinite, but they all share the same basic form. Bandpass amplitude response curves are shown in Figure 5.1. The curve in Figure 5.1(a) is what might be called an “ideal” bandpass response, with absolutely constant gain within the passband, zero gain outside the passband, and an abrupt boundary between the two. This response characteristic is impossible to realize in practice, but it can be approximated to varying degrees of accuracy by real filters. Curves in Figure 5.1(b) through (f) are examples of a few bandpass amplitude response curves that approximate the ideal curves with varying degrees of accuracy. Note that while some bandpass responses are very smooth, other have ripple (gain variations in their passbands. Other have ripple in their stopbands as well. The stopband is the range of frequencies over which unwanted signals are attenuated. Bandpass filters have two stopbands, one above and one below the passband.

Just as it is difficult to determine by observation exactly where the passband ends, the boundary of the stopband is also seldom obvious. Consequently, the frequency at which a stopband begins is usually defined by the requirements of a given system–for example, a system specification might require that the signal must be attenuated at least 35 dB at 1.5 kHz. This would define the beginning of a stopband at 1.5 kHz. The rate of change of attenuation between the passband and the stopband also differs from one filter to the next. The slope of the curve in this region depends strongly on the order of the filter, with higher-order filters having steeper cutoff slopes. The attenuation slope is usually expressed in dB/octave (an octave is a factor of 2 in frequency) or dB/ decade (a decade is a factor of 10 in frequency). Bandpass filters are used in electronic systems to separate a signal at one frequency or within a band of frequencies from signals at other frequencies. A filter could also reject unwanted signals at other frequencies outside of the passband, so it could be useful in situations where the signal of interest has been contaminated by signals at a number of different frequencies. (Web_1, 2004)

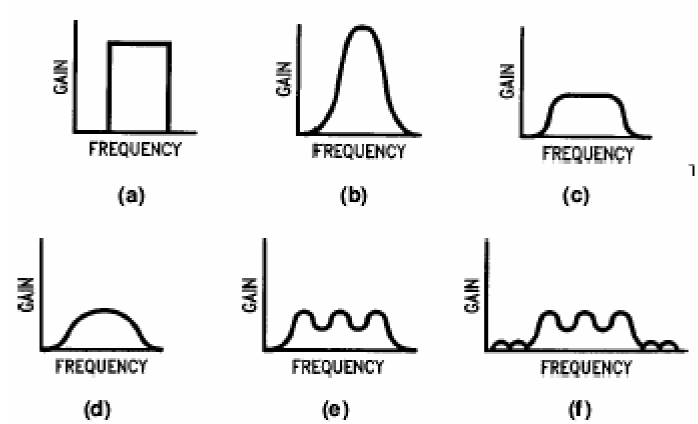


Figure 5.1 Examples of bandpass filter amplitude response (Web_1, 2004)

5.3 Filter Approximations

We saw several examples of amplitude response curves for various filter types. These always included an “ideal” curve with a rectangular shape, indicating that the boundary between the passband and the stopband was abrupt and that the roll off slope was infinitely steep. This type of response would be ideal because it would allow us to completely separate signals at different frequencies from one another. Unfortunately , such an amplitude response curve is not physically realizable. We will have to settle for the best approximation that will still meet our requirements for a given application. Deciding on the best approximation involves making a compromise between various properties of the filter's transfer function. The important properties are listed below:

Filter Order: The order of a filter is directly related to the number of components in the filter, and therefore to its cost, its physical size, andthe complexity of the design task. Therefore, higher-order filters are more expensive, take up more space, and are more difficult to design. The primary advantage of a higher order filter is that it will have a steeper rolloff slope than a similar lower-order filter.

Ultimate Rolloff Rate: Usually expressed as the amount of attenuation in dB for a given ratio of frequencies. The most common units are ``dB/octave'' and ``dB/decade''.

While the ultimate rolloff rate will be 20 dB/decade for every fitler pole in the case of a low-pass or high-pass filter and 20 dB/decade for every pair of poles for a bandpass filter, some filters will have steeper attenuation slopes near the cutoff frequency than others of the same order.

Attenuation Rate Near the Cutoff Frequency: If a filter is intended to reject a signal very close in frequency to a signal signal that must be passed, a sharp cutoff haracteristic is desirable between those two frequencies. Note that this steep slope may not continue to frequency extremes.

Transient Response: Curves of amplitude response show how a filter reacts to steady-state sinusoidal input signals. Since a real filter will have far more complex signals applied to its input terminals, it is often of interest to know how it will behave under transient conditions. An input signal consisting of a step function provides a good indication of this. Figure 5.2 shows the responses of two low-pass filters to a step input. Curve in Figure 5.2(b) has a smooth reaction to the input step, while curve (a) exhibits some ringing. As a rule of thumb, filters will sharper cutoff characteristics or higher Q will have more pronounced ringing.

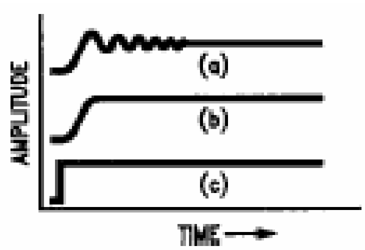


Figure 5.2 Step response of two different filters. Curve (a) shows significant ``ringing'', while curve (b) shows none. The input signal is shown in curve (c) (Web_1, 2004)

Monotonicity: A filter has a monotonic amplitude response if its gain slope never changes sign–in other words, if the gain always increases with increasing frequency or always decreases with increasing frequency. Obviously, this can happen only in the case of a low-pass or high-pass filter. A bandpass or notch filter can be monotonic on either side of the center frequency, however.

Passband Ripple: If a filter is not monotonic within its passband, the transfer function within the passband will exhibit one or more “bumps”. These bumps are known as “ripple”. Some systems don't necessarily require monotonicity, but do require that the passband ripple be limited to some maximum value (usually 1 dB or less). Examples of passband ripple can be found in Figures 5.1(e) and (f). Although bandpass filters do not have monotonic transfer functions, they can be free of ripple within their passbands.

Stopband Ripple: Some filter responses also have ripple in the stopbands. Examples are shown in Figure 5.1(f). We are normally unconcerned about the amount of ripple in the stopband, as long as the signal to be rejected is sufficiently attenuated. Given that the “ideal” filter amplitude response curves are not physically realizable, we must choose an acceptable approximation to the ideal response. The word “acceptable” may have different meanings in different situations. The acceptability of a filter design will depend on many interrelated factors, including the amplitude response characteristics, transient response, the physical size of the circuit and the cost of implementing the design. If we are willing to accept some deviations from this ideal in order to build a practical filter, we might end up with a curve like the one in which allows ripple in the pass- band, a finite attenuation rate, and stopband gain greater than zero. (Web_1, 2004)

5.3.1 Chebyshev Approximation

Approximation to the ideal filter is the Chebyshev or equal ripple response. As the latter name implies, this sort of filter will have ripple in the passband amplitude response. The amount of passband ripple is one of the parameters used in specifying a Chebyshev filter. The Chebyschev characteristic has a steeper rolloff near the cutoff frequency, but at the expense of monotonicity in the passband and poorer transient response. A few different Chebyshev filter responses are shown in Figure 5.3. The filter responses in the figure have 0.1 dB and 0.5 dB ripple in the passband, which is small compared to the amplitude scale in Figure 5.3(a) and (b), so it is shown expanded in Figure 5.3(c). It provides the greatest stopband attenuation but also the greatest overshoot. It has the worst for group delay flatness

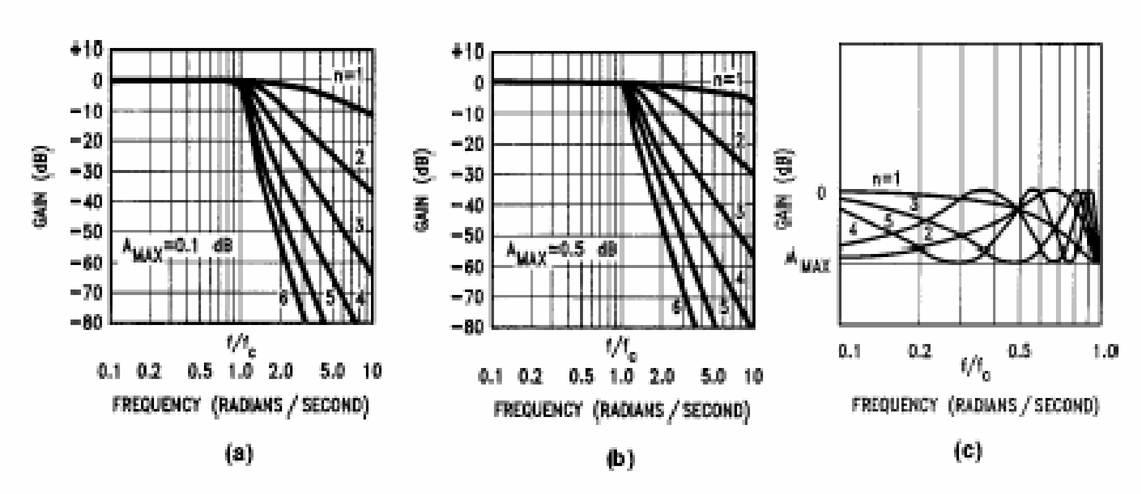


Figure 5.3 Examples of Chebyshev amplitude responses. (a) 0.1 dB ripple (b) 0.5 dB ripple. (c) Expanded view of passband region showing form of response below cutoff frequency. (Web_1, 2004)

Note that a Chebyshev filter of order n will have n-1 peaks or dips in its passband response. Note also that the nominal gain of the filter (unity in the case of the responses in Figure 5.3) is equal to he filter's maximum passband gain. An odd order Chebyshev will have a dc gain (in the low-pass case) equal to the nominal gain, with ``dips'' in the amplitude response curve equal to the ripple value. An even-order Chebyshev low-pass will have its dc gain equal to he nominal filter gain minus the ripple value; the nominal gain for an even-order Chebyshev occurs at the peaks of the passband ripple. Therefore, if it’s designed a fourth-order Chebyshev low-pass filter with 0.5 dB ripple and it’s wanted it to have unity gain at dc, it will be had to design for a nominal gain of 0.5 dB. The Chebyshev's cutoff frequency is normally the frequency at which the ripple (or Amax) specification is exceeded. The addition of passband ripple as a parameter makes the specification process for a Chebyshev filter a bit more complicated ,but also increases flexibility. Figure 5.4 shows the step response of 0.1 dB and 0.5 dB ripple Chebyshev filters of various orders. (Web_1, 2004)

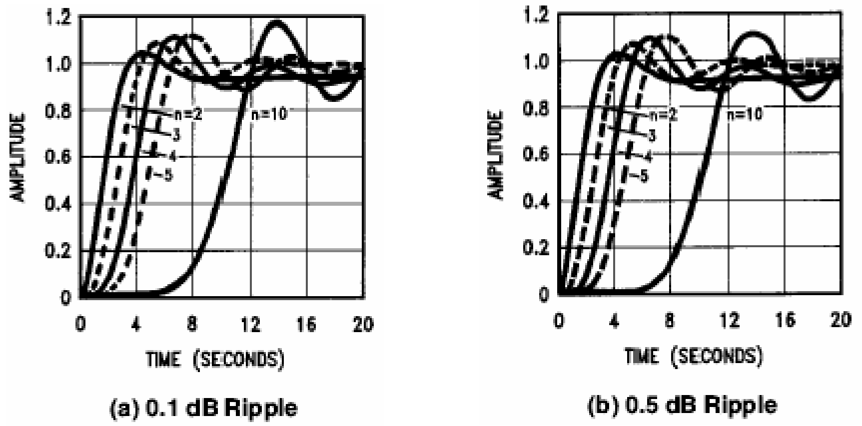


Figure 5.4 Step responses for Chebyshev low-pass filters. In each case, $\omega_0 = 1$, and the step amplitude is 1.0. (Web_1, 2004)

5.4 Coupled Microstrip Lines

The couple microstrip lines are exhibiting a field pattern with the quasi-TEM. A signal propagating on a coupled line can be described by the superposition of an even mode and odd mode that exhibit different propagation factors and characteristic impedance.

5.5 Coupled Line Filters

Parallel coupled microstrip filters are easily designed and implemented. Both approximate and exact realizations are available. These filters work well for bandwidths between 5% and 50%. There are two problems with these filters when implemented in microstrip: asymmetric response, and a parasitic passband at twice the center frequency. There are two basic methods of reducing the parasitic passband: equlizing the phase velocities, and providing different lengths for the even and odd modes. Equalizing the phase velocities may be done by suspending the substrate, opening a slot in the ground plane, or providing an overlay. However, all of these add cost to the original filter. (Riddle, 1988)

With the added tool of the impedance or admittance inverter, it can be analyzed and designed a number of transmission line filters. As it has been seen in connection with directional couplers, coupled transmission lines have frequency sensitive coupling, and can be analyzed by the even-odd mode method. The result of this analysis is tabulated in Table 8.8 of Pozar p. 478, and we can see that there are among the less useful permutations several that have bandpass characteristics. In particular, the configuration that represents coupled λ/2 open lines is the easiest to construct in microstrip and stripline.

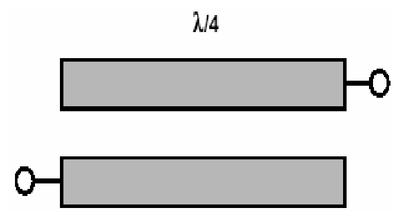


Figure 5.5 Bandpass coupled line circuit fitler (Riddle, 1988)

The equivalent circuit of two coupled λ/4 open lines can be shown to be as depicted:

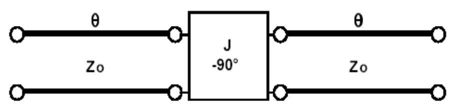


Figure 5.6 The equivalent circuit of two coupled λ/4 open lines (Riddle, 1988)

So we can see that a structure of a number of coupled lines will admit to an quivalent circuit of alternating series and parallel resonant circuits, and the design parameters of the prototype filter can be imposed onto the structure of parallel coupled lines.

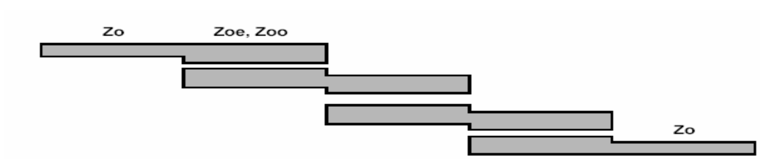


Figure 5.7 Layout of coupled line bandpass fitler (Riddle, 1988)

In microstrip or stripline, the transmission line conductors of the coupled line filter take the form shown here, with the offsets between connected λ/4 sections added to permit seeing the individual coupled line pairs.

5.5.1 Coupled Line Filter Design

The goal is to design and fabricate a 2.4 and 2.6 GHz microstrip bandpass filters. The bandpass filter is used to pass a particular band of frequencies while providing attenuation in the stopband. It must pass the 2400±50 MHz and 2600±50 MHz signals that will be radiated by the antenna. To distinguish which signal is present on a line of the circuit, bandpass filters are used. A pass-band of 100 MHz is desirable with an insertion loss that is as low as possible. The falloff in the stop bands should be steep with values at the opposite frequency of at least –30 dB. A ripple of 0.5 dB is sufficient for the design. The filter should also be matched to the characteristic impedance of 50 ohms. (Web_14, 2004)

The circuit model for symmetrically coupled lines can be defined in terms of two normal modes. The parameters needed in the design of coupled line structures are the even and odd mode characteristic impedance (Z_{oe} and Z_{oo}). Analysing individual modes of propagation along two coupled lines by network analysis techniques yields expressions for the design parameters. The expressions assume quasi-TEM propagation, and symmetry of structure. Both these assumptions hold in the case of a pair of embedded symmetrically coupled transmission lines at low microwave frequencies. From the impedance inverter values, the even- and odd-mode impedance (Z_{oe}, Z_{oo}) of a coupled line can be calculated for a given characteristic impedance. The line width and spacing of the coupled line are then determined from the given even- and odd-mode impedances. (Ton, Shih & Bui, 1987)

When designing the filters, normalized frequencies and impedances are used. In Table 5.1, the parameters for fifth order coupled line filters that meet the specifications for this project are shown. These values are used in the AWR MWO software package to find the lengths, widths, and separations for each of the respective coupled line sections. This process is performed at both 2.4 and 2.6 GHz to produce the microstrip implementations of the circuits as shown in Figure 5.8 and 5.10. Each filter section is one quarter wavelength long (Pozar, 1998, p. 475-485). According to Δ=5% (fractional bandwidth of the passband), ripple of 0.5 dB, Z_0=50 Ω and N=5 it is obtained the even and odd mode coupled line characteristic impedances Z_{0e} and Z_{0o}.

It is calculated the admittance inverter constants (Equations 5.1, 5.2, 5.3)

$$Z_o \cdot J_1 = Z_o \cdot J_4 = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_1}} = \sqrt{\frac{\pi \cdot \Delta}{2 \cdot g_5 \cdot g_6}} = 0{,}214521 \qquad 5.1$$

$$Z_o \cdot J_2 = Z_o \cdot j_5 = \frac{(\pi \cdot \Delta)}{2\sqrt{g_1 \cdot g_2}} = \frac{(\pi \cdot \Delta)}{2 \cdot \sqrt{g_4 \cdot g_5}} = 0{,}054205 \qquad 5.2$$

$$Z_o \cdot J_3 = \frac{(\pi \cdot \Delta)}{2\sqrt{g_2 \cdot g_3}} = 0{,}044413 \qquad 5.3$$

Find the even-mode characteristic impedances (Equations 5.4, 5.5, 5.6)

$$Z_{oe1} = Z_{oe4} = Z_o \cdot \left[1 + Z_o \cdot j_1 + (Z_o \cdot j_1)^2\right] = 63{,}0270 \qquad 5.4$$

$$Z_{oe2} = Z_{oe5} = Z_o \cdot \left[1 + Z_o \cdot j_2 + (Z_o \cdot j_2)^2\right] = 52{,}8571 \qquad 5.5$$

$$Z_{oe3} = Z_o \cdot \left[1 + Z_o \cdot j_3 + (Z_o \cdot j_3)^2\right] = 52{,}3192 \qquad 5.6$$

Find the odd-mode characteristic impedances (Equations 5.7, 5.8 , 5.9)

$$Z_{oo1} = Z_{oo4} = Z_o \cdot \left[1 - Z_o \cdot j_1 + (Z_o \cdot j_1)^2\right] = 41{,}5749 \qquad 5.7$$

$$Z_{oo2} = Z_{oo5} = Z_o \cdot \left[1 - Z_o \cdot j_2 + (Z_o \cdot j_2)^2\right] = 47{,}4366 \qquad 5.8$$

$$Z_{oo3} = Z_o \cdot \left[1 - Z_o \cdot j_3 + (Z_o \cdot j_3)^2\right] = 47{,}8779 \qquad 5.9$$

Table 5.1 Coupled line filter parameters for N=5

Delta = 0.05	g_n	$Z_0 J_n$	Z_{0e}	Z_{00}
S1	1,7058	0.214521	63,0270	41,5749
S2	1,2296	0.054205	52,8571	47,4366
S3	2,5408	0.044413	52,3192	47,8779
S4	1,2296	0.214521	63,0270	41,5749
S5	1,7058	0.054205	52,8571	47,4366

A plot of the simulations for both the 2.6 and 2.4 GHz filters is shown in Figures 5.9 and 5.11. The frequency sweep is from 2-3 GHz and the parameter of interest is the transmission coefficient, S_{21}, from the input port to the output port for each filter. As can be seen in the plot, the filters have smooth responses and steep falloffs outside the pass-band. When the plots appear as they should pass bands for 2.4 and 2.6 GHz respectively. The schametic of a coupled line filter is shown in Figures 5.8 and 5.10.

It does not matter which side is used for input and which is used for output.

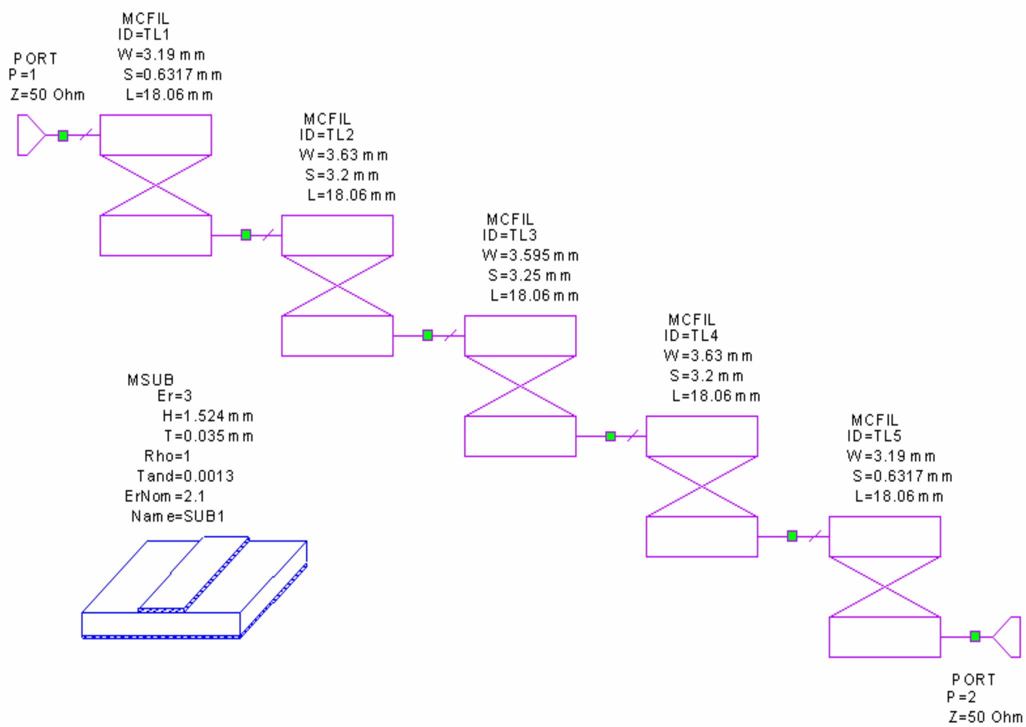


Figure 5.8 Coupled line bandpass filter schematic for 2.6 GHz

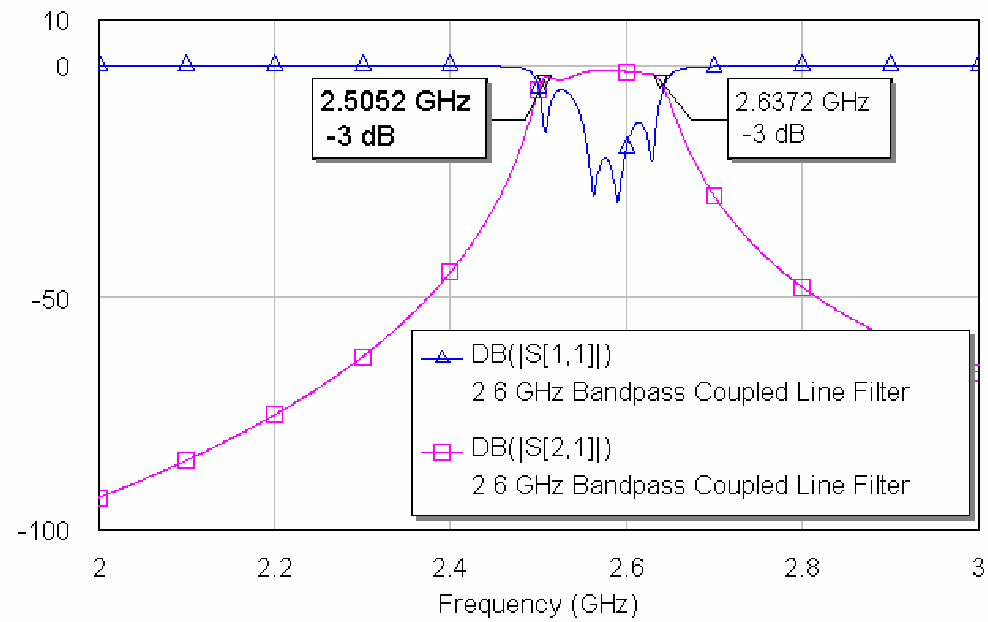


Figure 5.9 Simulation result of transmission (S21) and reflection (S11) coefficients in dB (fc=2.6 GHz, Δ=5%)

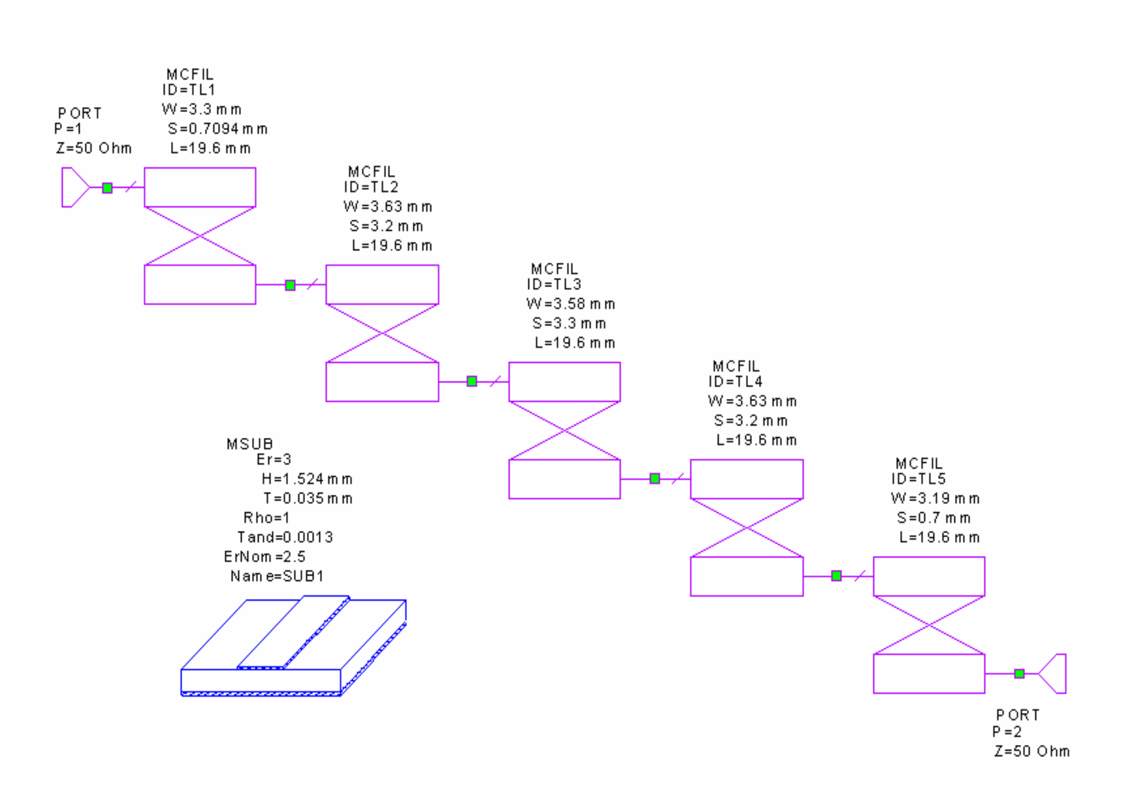


Figure 5.10 Coupled line bandpass filter schematic for 2.4 GHz

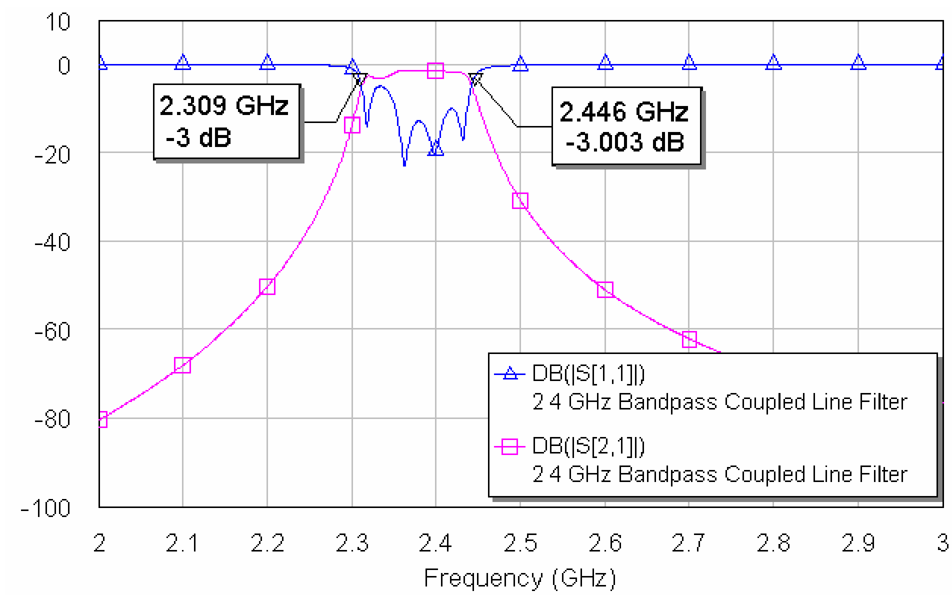


Figure 5.11 Simulation result of transmission (S21) and reflection (S11) coefficients in dB (fc=2.4 GHz, Δ=5%)

Table 5.2 Behavioural variation guideline for strip based coupled line geometrical parameters, and Z_{oo} and Z_{oe} (Yun, & the others, 2000)

Geometry parameter	Zoe	Zoo
↑W	↓ Zoe	↓ Zoo
↓ W	↑ Zoe	↑ Zoo
↑ S	↓ Zoe	↑ Zoo
↓ S	↑ Zoe	↓Zoo
↑ H	↑ Zoe	↓ Zoo
↓ H	↓ Zoe	↑ Zoo
Key:- Increase (↑); Decrease (↓); Width (W); Separation (S); Substrate Height (H).		

Design a coupled- line filter with the following specifications:

- Center frequency: 2.4 and 2.6 GHz
- Bandwidth Target: 5%
- 0.5 dB equal ripple Chevyshev response
- Order of the multi-section bandpass coupled-line filter, N=5

TX-Line Program in AWR was used to calculate the dimensions of the copper strips on substrates. Circuit simulations were performed for two filters. This substrate offered the minimum power loss in the passband. The size of the filter is approximately 98 x 34.92 mm^2. This type of filter will reduce the size. Note that the skirt is much steeper for the higher-order filter.

The signal generator and spectrum analyzer can also be used to test the response of the filters. The signal generator should be connected to one side of the circuit via the circuit holder, and the spectrum analyzer should be connected to the other side of the circuit. In the case of the filter, the circuit is symmetric. It does not matter which side is used for input and which is used for output. The spectrum analyzer will show what power is output when a constant power is input from the signal generator. The frequency range can be swept, and the response should correspond to the one obtained from the network analyzer.

CHAPTER SIX
DETECTOR CIRCUIT

6.1 Introduction

The goal of the detector circuit is to convert a high frequency signal into a DC voltage. Cost, size, weight and power consumption are inherent limiting factors in this design.

Schottky diode detectors are commonly used as amplitude demodulators and level detectors in wireless and other RF and microwave signal processors. Detector designs are simple to realize using low cost, plastic packaged, silicon Schottky diodes. The module needs to have minimum component count and extremely low power new developments in low-barrier Schottky diodes suggested the viability of a detector with zero DC bias. (Web_2, 2005)

Input matching requirements were realised by using a microstrip line with a short circuit stub located the diode. This inductive stub doubled as the DC return path in the detector circuit, hence minimizing the component count. (Crumle & Evans, 1996)

6.2 Microwave Diode Detectors

The goal of the detector circuit is to convert a high frequency signal into a DC voltage. In this case, a diode rectifier and integrating capacitor are used to meet this goal.

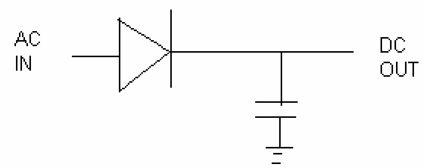


Figure 6.1 Microwave diode detector

An effective matching circuit is needed to prevent reflections and maximize the power input to the diode detector circuit. Detector circuits with matching networks are shown in Figure 6.2.

A passive network is connected to the diode cathode, and then this combination is modeled with an RLC load. Using the explicit formulas of filter approximations such as Chebyshev, a broadband matching network is synthesized between the RLC load and a resistive element which is not equal to the characteristic impedance. Then another matching network (filter) is synthesized between the resistive element and the characteristic impedance. This method produces diode detectors over its useful broad bandwidth without using diode biasing (self bias) and without using resistance in the diode matching network.

Figure 6.2 shows a block diagram of the microwave square law diode detector. The detector circuit is composed of an input matching network, a diode (usually Schottky barrier diode) and a passive network.

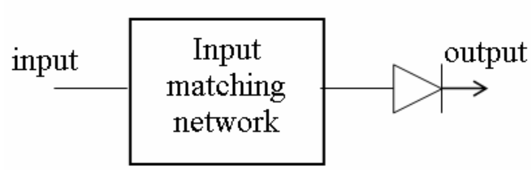


Figure 6.2 Microwave detector block diagram

To design a microwave broadband square law diode detector, the input matching network must transform the diode impedance to the input characteristic impedance (usually 50 Ω) over the entire bandwidth. When external dc bias is used for the diode, the diode voltage sensitivity decreases and consequently the detector voltage sensitivity also decreases. In order to obtain the best voltage sensitivity, no external dc bias is used for the diode and the diode is biased with the dc rectified (or detected) current resulting from the rectification of the incident microwave signal (self bias). (Al Anwar, El Hennawy, Hashish, & El Said Mostafa, 2001)

The steps of the proposed Computer Aided Design (CAD) method for the design of broadband microwave planar square law diode detector are given as follows:

1 - Select a suitable diode that has low package parasitic in the desired bandwidth and a zero- volt cut off frequency considerably larger than the upper frequency end of the desired bandwidth. The diode zero-volt cut off frequency is given by:

$$f_{CO} = 1 \big/ \left(2\pi R_S C_{JO}\right) \tag{6.1}$$

where R_S, is the series resistance of the diode and C_{jo}, is the zero-volt junction capacitance of the diode. The diode must not have dc external bias in order to achieve the highest possible diode voltage sensitivity.

2 - Add a passive network e.g. open circuit transmission line, (at the diode cathode) as shown in Figure 6.3

Figure 6.3 Open circuit stub connected to the diode cathode

3 - The parameters of the RLC model shown in Figure 6.4 for the diode and passive network combination shown in Figure 6.5 by making the model and the circuit have equal impedances (or equal input reflection coefficients) over the desired bandwidth by optimizing for ($Z_{model} = Z_{circuit}$) over the desired bandwidth.

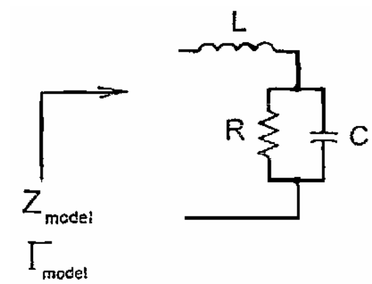


Figure 6.4 The RLC model (Al Anwar, El Hennawy, Hashish, & El Said Mostafa, 2001)

Such that the variables of optimization are R, L, C and the passive network dimensions. $\Gamma_{circuit}$ is the input reflection coefficient of the diode and the passive network combination (see Figure 6.4) and Γ_{model} is the input reflection coefficient of the RLC model (see Figures 6.4 and 6.5).

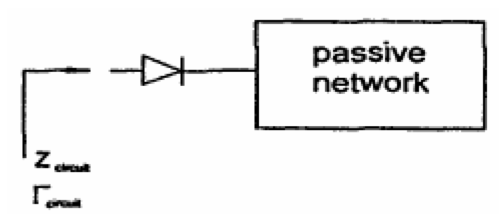


Figure 6.5 Passive network connected to the diode cathode. (Al Anwar, & the others, 2001)

Changing the dimensions of the passive network connected to the diode during optimization will change the circuit impedance ($Z_{circuit}$) over the desired bandwidth and will facilitate and fasten the optimization convergence. This is the reason of using the passive network, noting that it may consist of a single open circuit transmission line as mentioned above (see Figure 6.3). Usually fast convergence is achieved.

4 - Use the RLC model parameters to synthesize Chebyshev low pass broadband matching network ended by a resistive element (R_T) as shown in Figure 6.6

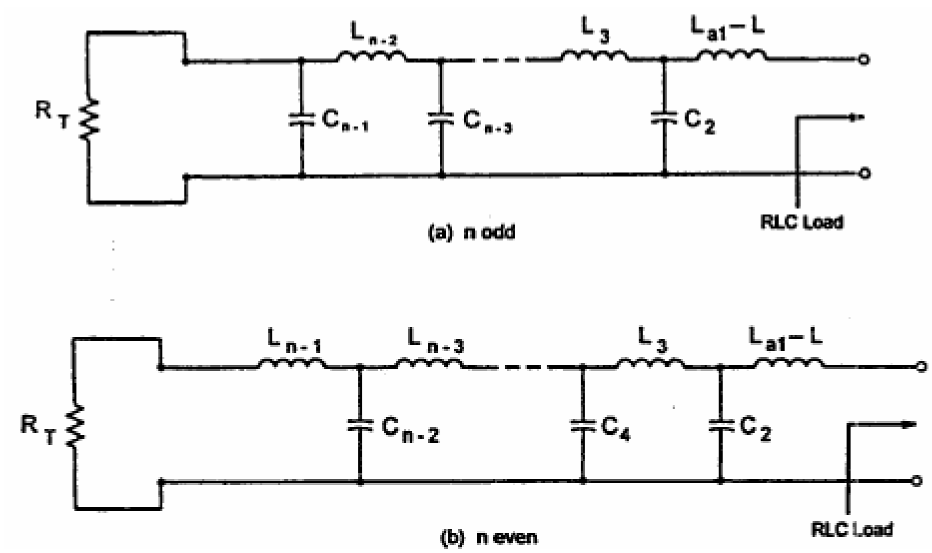


Figure 6.6 L,C ladder matching network for the RLC load (Al Anwar, & the others, 2001)

5 - Design another Chebyshev low pass filter between R_T and the characteristic impedance (Zo = 50 Ω) with a cut off frequency equal to that of the matching network designed in step 4.

6 - The combination of the two LC ladder networks designed in steps 4 and 5 represents the matching network between the load (the diode and passive network) and the characteristic impedance Zo = 50 Ω (see Figure 6.7). The transmission lines parameters from the lumped to distributed transformation formulas using the calculated LC ladder components of steps 4 and 5. (Al Anwar & the others, 2001)

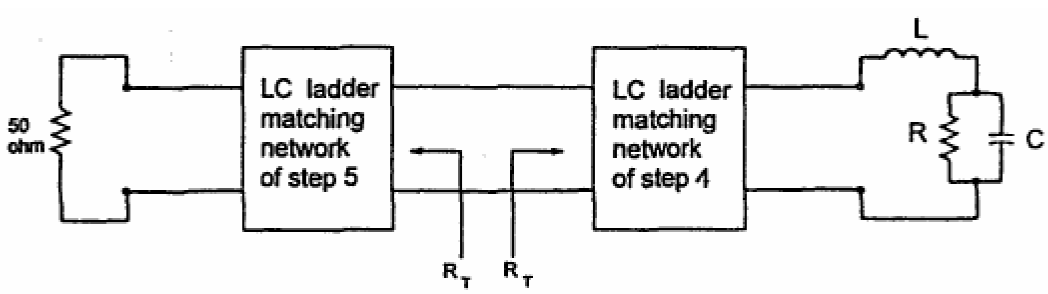


Figure 6.7 The total matching network between the complex load and 50 Ω(Al Anwar, & the others, 2001)

6.3 Schottky Detector Fundamentals

6.3.1 Schottky Equation

Schottky diode detector operation is based on the equation that characterizes the current-voltage relationship in a diode junction, as shown:

$$I = I_{SAT}\left(e^{\frac{q(V-IR_S)}{nKt}} - 1\right) \qquad 6.2$$

where;

n: ideality factor (typically 1.0)

K: Boltsmann's constant, 1.38044×10^{-23} (joule/Kelvin)

q: electronic charge, 1.60206×10^{-19} (coulombs)

t: temperature (Kelvin)

R_S: series resistance (Ω)

I_{SAT}: saturation current

This equation may be simplified for a Schottky diode operated as a zero biased detector at 300 K, where R_S may be neglected and n = 1.0 to:

$$I = I_{SAT}\,(e^{38.6\,V} - 1) \qquad 6.3$$

This equation is appropriate for all junction diodes (PN, PIN and Schottky). However, only for the Schottky diode is this equation valid at high microwave frequencies.

The dynamic resistance of a Schottky diode (the slope of the I-V characteristic) is a function of forward current and may be expressed as:

$$R_V = \frac{q}{nKt} \times \frac{1}{I} \qquad 6.4$$

This expression also gives the video resistance of the diode at zero bias as:

$$R_V = \frac{0.026}{I_{SAT}} \qquad 6.5$$

The video resistance value may then be calculated directly from saturation current. As a detector, the Schottky diode's performance has been analyzed by simulating the application of a sinusoidal signal at a DC operating point and deriving the generated non-sinusoidal response. Because of the non-linear nature of the diode equation, this response is not sinusoidal but is rich in harmonics. Among the second order terms is a DC component, which is a measure of the magnitude of the applied signal and becomes the generated detector signal. (Web_2, 2005)

6.3.2 Square Law and Linear Response

The response of a Schottky detector is generally presented as a curve of detected output voltage vs. applied input power in a circuit where the detector diode terminates a transmission line. At small signals the output voltage is closely proportional to the input power or the square of the input voltage. This is called the square law region. In this region, a 10 dB increase in input power results in a 10 times increases in output voltage. At large input signals the detector voltage is more directly proportional to the input RF voltage. This is the linear region and it is where a 10 dB increase in input power results in a 5 times increases in output voltage. Figure 6.8 demonstrates this effect.

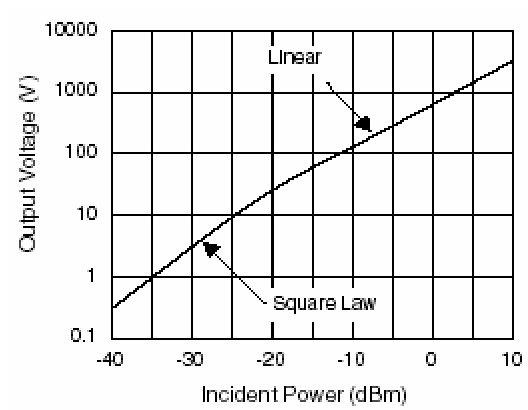


Figure 6.8 Square law and linear regions (Web_20, 2005)

All Schottky diodes operating at zero external bias generate virtually the same open circuit detector voltage at the same input power. In this respect, all Schottky diodes may be considered zero bias detectors. Improved detector sensitivity may be achieved by maximizing the RF voltage on the diode junction. This is accomplished by increasing the source impedance. (Web_2, 2005)

6.3.3 Characteristics

The Schottky diode is what is called a majority carrier device. This gives it tremendous advantages in terms of speed because it does not rely on holes or electrons recombining when they enter the opposite type of region as in the case of a conventional diode. By making the devices small the normal RC type time constants can be reduced, making these diodes an order of magnitude faster than the conventional PN diodes. This factor is the prime reason why they are so popular in radio frequency applications. (Web_21, 2005)

6.3.4 Applications

The Schottky diode is widely used in the electronics industry finding many uses as a general purpose rectifier. However it has come into its own for radio frequency applications because of its high switching speed and high frequency capability. In view of this it is used in many high performance diode ring mixers. In addition to this their low turn on voltage and high frequency capability and low capacitance make them ideal as RF detectors. (Web_21, 2005)

6.3.5 Schottky Barrier Diode

The Schottky Barrier Diode is actually a variation of the point-contact diode in which the metal semiconductor junction is a surface rather than a point contact. The large contact area, or barrier, between the metal and the semiconductor in the Schottky barrier diode provides some advantages over the point-contact diode. Lower forward resistance and lower noise generation are the most important advantages of the Schottky barrier diode. The applications of the Schottky barrier diode are the same as those of the point-contact diode. The low noise level generated by Schottky diodes makes them especially suitable as microwave receiver detectors and mixers. The Schottky barrier diode is sometimes called the Hot-Electron or Hot-Carrier Diode because the electrons flowing from the semiconductor to the metal have a higher energy level than the electrons in the metal. The effect is the same as it would be if the metal were heated to a higher temperature than normal. Figure 6.9 is an illustration of the construction of a Schottky barrier diode. (Al Anwar & the others, 2001)

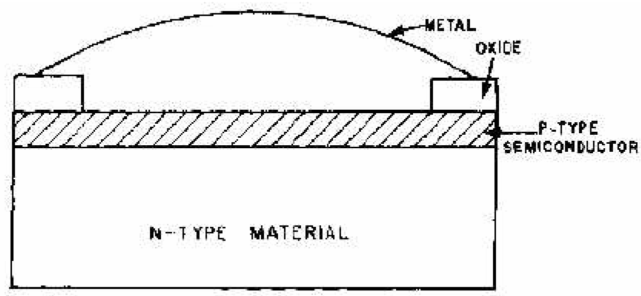


Figure 6.9 Schottky-barrier diode (Web_19, 2005)

6.4 Design

The mirostrip substrate used in this circuit is RT Duroid 5880™, which have the following parameters: {dielectric constant ε_r = 2.2, dielectric thickness h = 1.575 mm, conductor (copper) thickness for each side t = 0.036 mm}. Two chosen diodes for the detectors are beam lead Schottky barrier diodes with low parasitic from 1 GHz to 26 *GHz.* Self bias is chosen for all the diodes in order to achieve the highest possible voltage sensitivity.

6.4.1 - 2.4 GHz Diode Detector

A microstrip square law diode detector 2,4 GHz is designed using the proposed method. The design steps are as follows:

1 -The beam lead Schottky barrier diode **HSCH-5338™** is chosen for the detector circuit.

2-An open circuit transmission line is chosen to be connected to the diode cathode (see Figure 6.3).The initial chosen length for the open circuit stub is 12,31 mm and the initial width is 12,76 mm

3 - Butterworth synthesis can be used to synthesize LC ladder matching network of order (n=3) between the RLC load and a resistive load R_T 2.4 GHz .Then Butterworth equations (Pozar, 1998) are used to synthesize LC ladder filter of order (n=3) between R_T and 50 ohm at 2,4 GHz .The total LC matching network is composed of 5 elements as shown in Figure 6.10. Part (a) of the matching network shown in Figure 6.10 is the LC matching network between the complex load and R_T. Part (b) of the matching network is the LC matching network (filter) between 50 ohm and R_T.

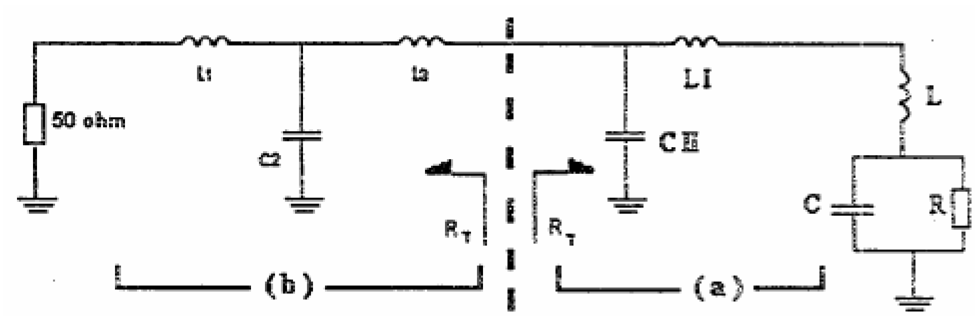


Figure 6.10 The total LC matching network between the complex load and 50 Ω (Al Anwar, 2001)

4 - The distributed matching network parameters are calculated using low pass to band pass transformation formulas given by reference (Pozar, 1998, p. 486-496). The microstrip transmission lines dimensions are calculated using the microstrip synthesis

formulas given by reference (Pozar, 1998, p. 160-177) then final circuit optimization is performed. (Al Anwar & the others, 2001)

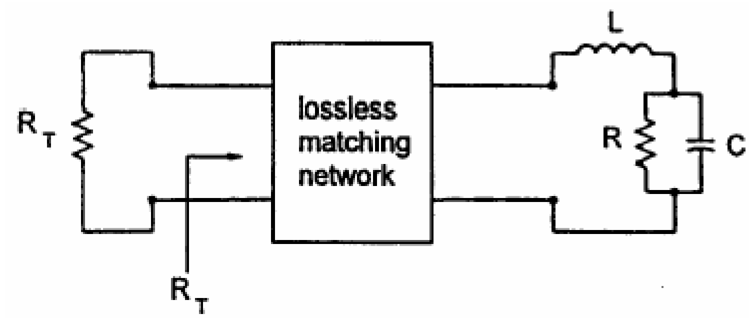


Figure 6.11 Matching network for the R,L,C Load (Al Anwar & the others, 2001)

Schottky diode detectors are used to detect small signals close to the noise level and to monitor large signals well above the noise.

6.4.2 - 2.6 GHz Diode Detector

A single stub matching network can be calculated for circuit at 2.6 GHz. 2.6 GHz diode detector is designed such as 2.4 GHz diode detector. The 2,6 GHz detector circuit is shown in Figure 6.15.

The important parameter to measure is S_{11}. Its value should be on the order of –20 dB. Copper tape can be used to fine tune the circuit if necessary. When the proper frequency is placed on the input of the circuit, a DC voltage should appear on the output after the diode. With an input power of 0 dBm, an output voltage of at least 0,7 V is expected for both detectors.

6.5 Simulation Results

The design of broadband microwave planar square law diode detectors is presented. Here a passive network is connected to the diode cathode then this combination is modeled with an RLC load. Using the explicit formulas of the filter approximations such as Chebyshev, a broadband matching network is synthesized between the RLC load and a resistive element which is not equal to the characteristic impedance. Then another matching network (filter) is synthesized between the resistive element and the characteristic impedance (50 ohm). This method produces diode detectors over its useful broad bandwidth without using diode biasing (self bias) and without using resistance in the diode matching network. Two broadband square law Schottky barrier diode detectors designed using this method with their analysis are presented. These detectors have suitable topologies for realization using microstrip line

The microstrip substrate used in these examples is RT Duroid 5880™, which have the following parameters: {dielectric constant ε_r = 2.2, dielectric thickness h = 1.575 mm, conductor (copper) thickness for each side t = 0.036 mm}. Two chosen diodes for the detectors are beam lead Schottky barrier diodes (HSCH-5338™) with low parasitic from 1 GHz to 26 *GHz*. Self bias is chosen for all the diodes in order to achieve the highest possible voltage sensitivity.

A diode detector is used to convert AC voltage to DC voltage. We will use this in our WLAN to measure the voltage after the filters to determine which frequency was passed (corresponding to a 1 or a 0). With an input power of 0 dBm, an output voltage of at least 0.7V is expected for both detectors. When it is given it a frequency of 2.4 or 2.6 GHz (depending on the detector) and a magnitude of 0 dBm, we see DC voltages of about 1.373 V or 0.7724 V on the DC voltmeter for 2.4 GHz and 2.6 GHz respectively as shown in Table 6.1.

Table 6.1 Output voltages of detectors

	at 2.4 GHz	at 2.6 GHz
2.4GHz Detector	1.373V	0.1327V
2.6GHz Detector	0.2175V	0.7724V

Linear detection is used in power monitors. In some applications the linearity is important because the detected voltage is a measure of power input. At low power levels (receiver applications) detected voltage is proportional to input power. This is the square law detection region.

The important parameter to measure is S_{11}. Its value should be on the order of –20 dB. As shown in Figure 6.13 and 6.16, the values are on the order of –35 dB and –28 dB at 2.4 GHz and 2.6 GHz respectively.

6.5.1 For 2.4 GHz Detector

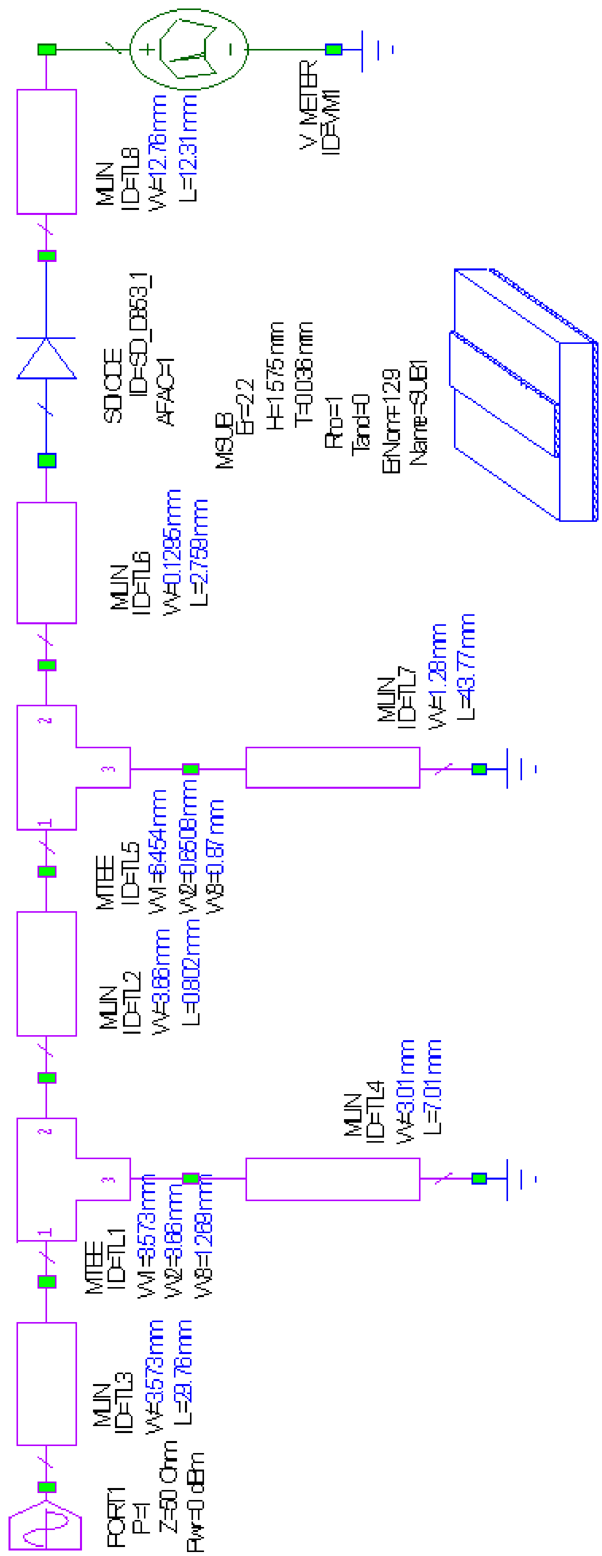


Figure 6.12 - 2.4 GHz Detector circuit

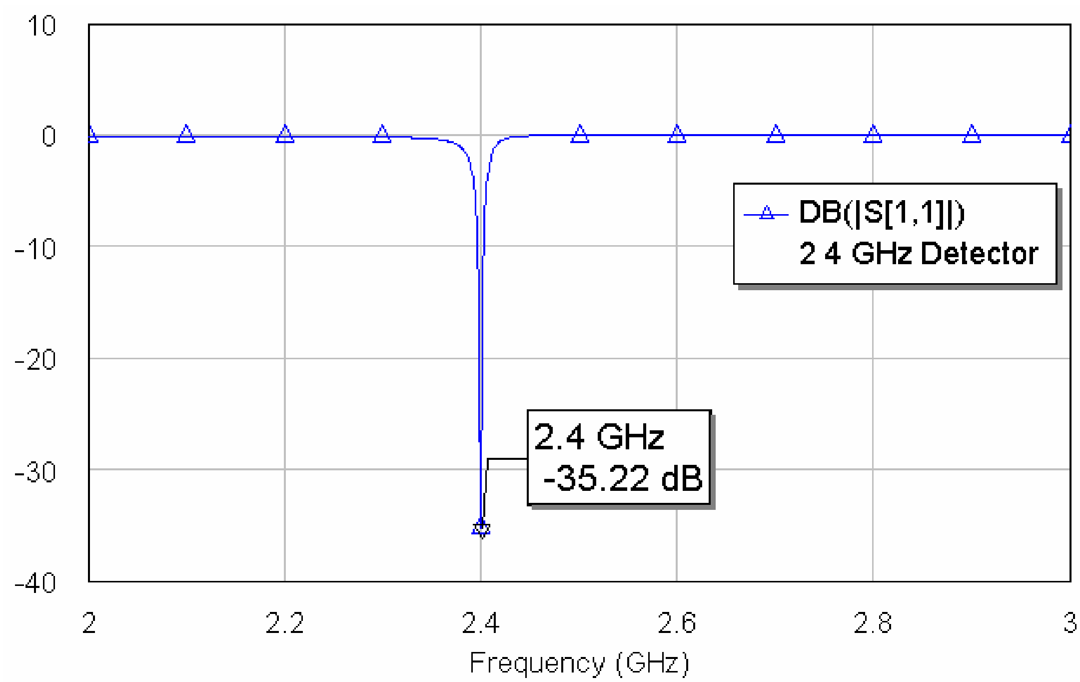


Figure 6.13 Reflection coefficient of 2.4 Ghz detector circuit

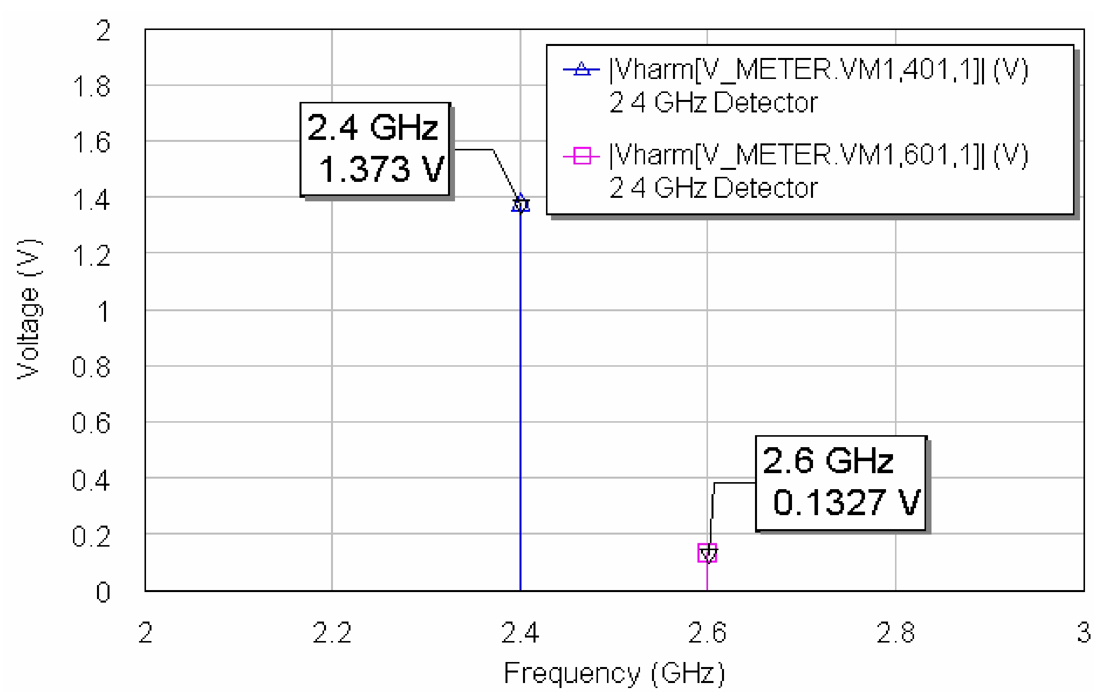


Figure 6.14 Output voltage of 2.4 Ghz detector circuit

6.5.2 For 2.6 GHz Detector

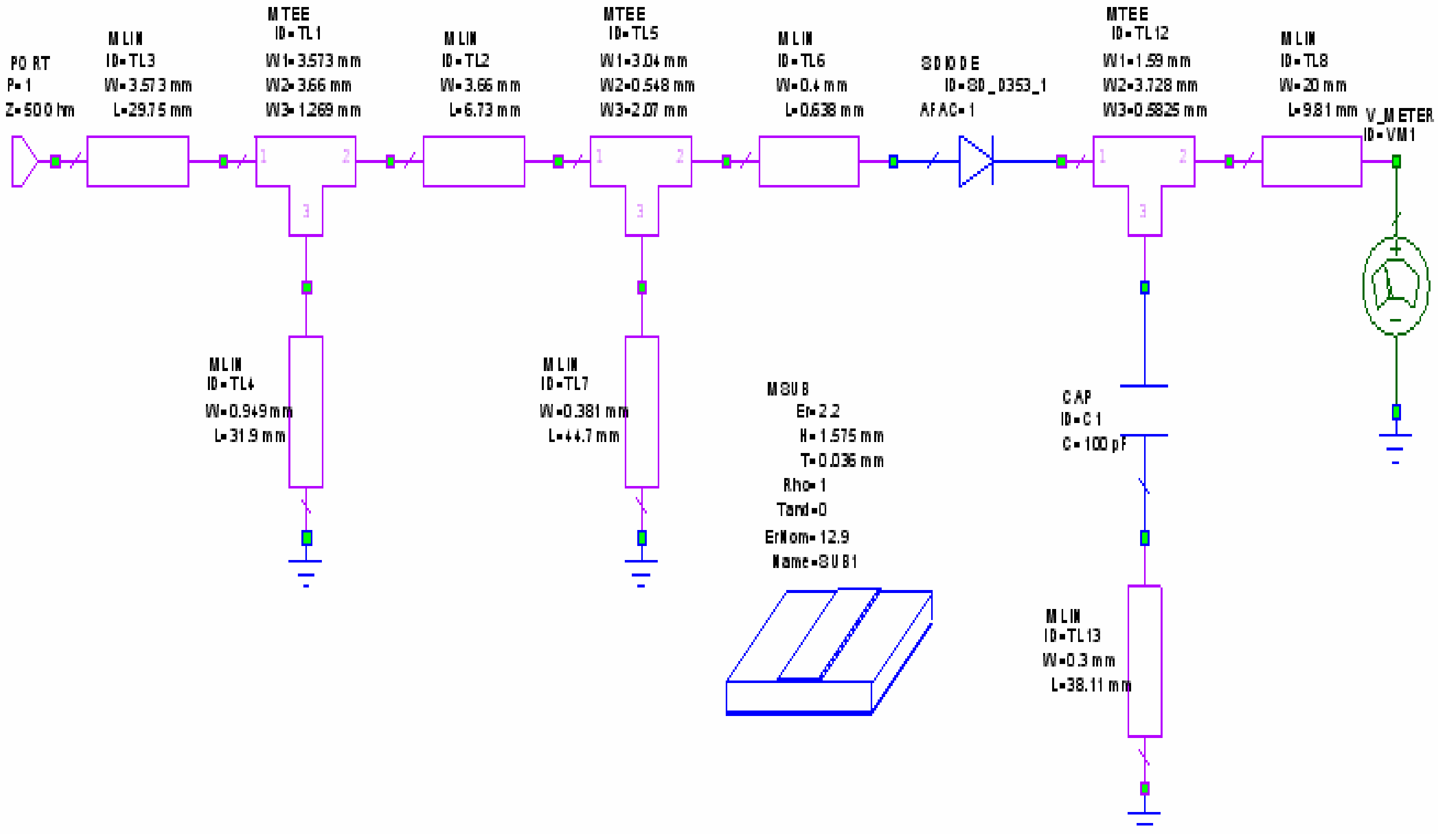


Figure 6.15 - 2.6 GHz Detector circuit

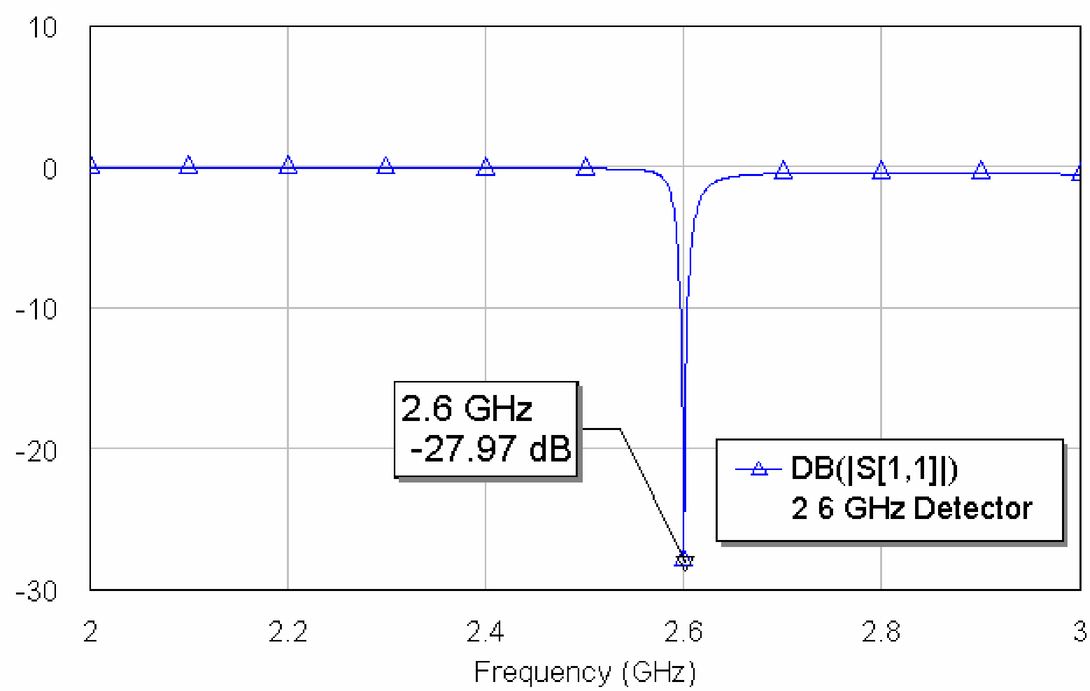


Figure 6.16 Reflection coefficient of 2.6 Ghz detector circuit

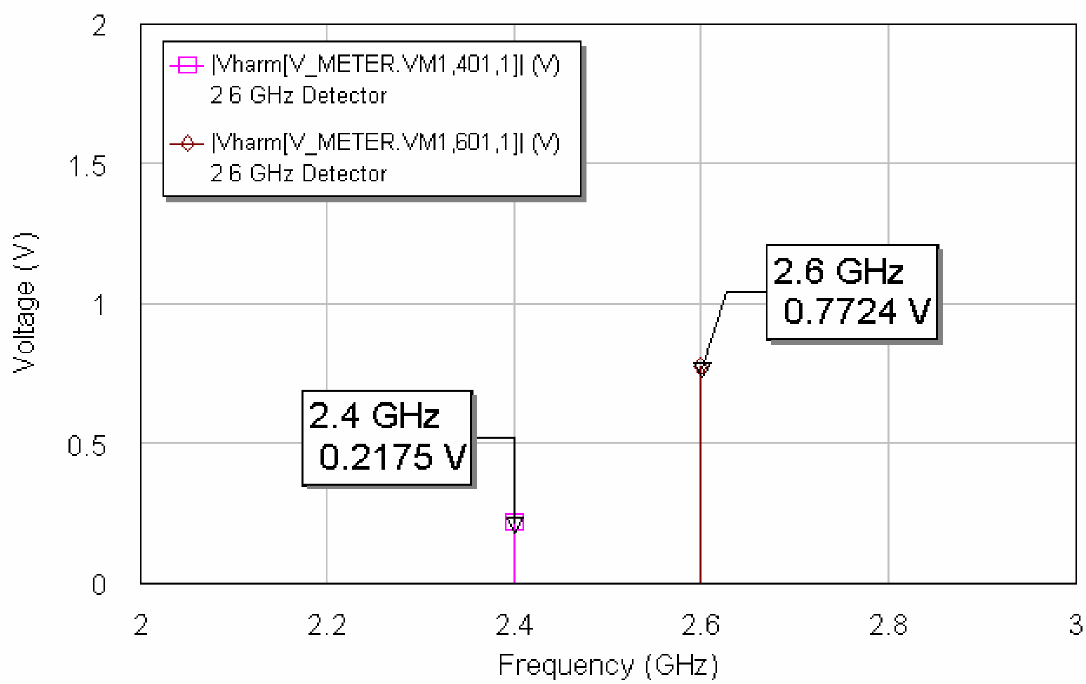


Figure 6.17 Output voltage of 2.6 Ghz detector circuit

CHAPTER SEVEN
COMPARATOR CIRCUIT

7.1 Introduction

A comparator compares two voltages in its input, if one of the non-inverting input is higher than the inverting input, the output is driven to one of the positive supply rail. The output of the comparator will have to be tie to high with a resistor. A comparator will be used in this communication system to detect a shift between the outputs of the quadrature detector when a shift in frequency occurs.

7.1.1 Comparator Operation

The following drawing shows the two simplest configurations for voltage comparators. The diagrams below the circuits give the output results in a graphical form. For these circuits the REFERENCE voltage is fixed at one-half of the supply voltage while the INPUT voltage is variable from zero to the supply voltage.

In theory the REFERENCE and INPUT voltages can be anywhere between zero and the supply voltage but there are practical limitations on the actual range depending on the particular device used. (Web_24, 2005)

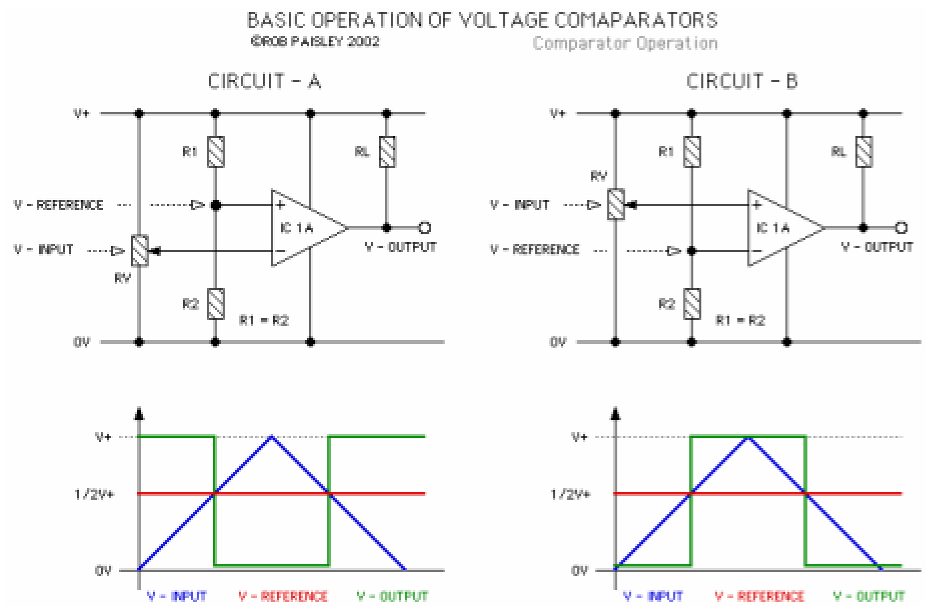


Figure 7.1 Basic operation of voltage comparators (Web_24, 2005)

7.1.2 Input vs. Output Results

1. Current WILL flow through the open collector when the voltage at the PLUS input is lower than the voltage at the MINUS input.
2. Current WILL NOT flow through the open collector when the voltage at the PLUS input is higher than the voltage at the MINUS input. (Web_24, 2005)

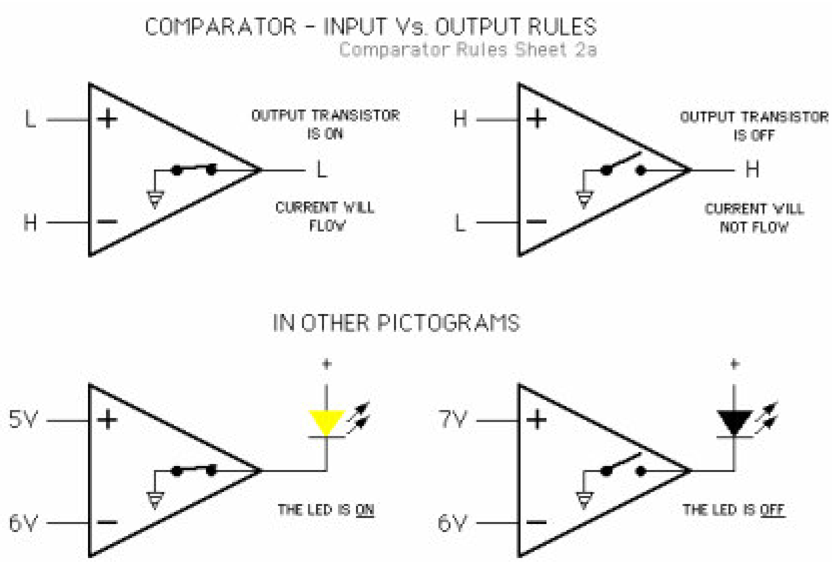


Figure 7.2 Input vs. output rules (Web_24, 2005)

7.2 LM 311 Voltage Comparator

A comparator is used in this communication system to detect a shift between the outputs of the detector when a shift in frequency occurs. LM311 was chosen as a comparator. LM311 was chosen because of its price and its response time is sufficient for maximum bit rate of 200kbps. The LM311 is voltage comparator that has input currents nearly a thousand times lower than devices like the LM106 or LM710. They are also designed to operate over a wider range of supply voltages: from standard ±15V op amp supplies down to the single 5V supply used for IC logic. Their output is compatible with RTL, DTL and TTL as well as MOS circuits. Further, they can drive lamps or relays, switching voltages up to 50V at currents as high as 50 mA. Both the inputs and the outputs of the LM111, LM211 or the LM311 can be isolated from system ground, and the output can drive loads referred to ground, the positive supply or the negative supply. (Web_7, 2005)

A problem with circuits based on a comparator like the LM311 that is able to change state rapidly is that they are likely to oscillate between high and low output states when the two inputs vary slowly and are close to one another in magnitude. To get round this problem, it is common to use what is known as *hysteresis*, whereby the transition voltage is made to change slightly depending whether the output of the comparator is high or low.

Features:

- Operates from single 5V supply
- Input current: 150 nA max. over temperature
- Offset current: 20 nA max. over temperature
- Differential input voltage range: ±30V
- Power consumption: 135 mW at ±15V
- The LM311 has a temperature range of 0°C to +70°C.

LM 311 detector circuit and its simulation result are designed with Electronics Workbench Multisim V7.0 software. The outputs of each detector circuit are connected to the input terminals of a comparator (LM 311). Outputs of detectors are DC voltage, so LM 311 compares these DC voltages. Output voltages of detectors were 1.37V DC at 2.4GHz for 2.4GHz detector and 0.77V DC at 2.6GHz for 2.6GHz detector. These detector outputs are connected with legs of number 2 and 3 of LM311. As shown in Figure 7.4, the output is the digital data, which will be uploaded to the PC. The circuit will effectively produce a 1 or 0 depending on whether a frequency of 2.4 GHz or 2.6 GHz is received and detected in the circuit. For our device to communicate with the PC, it must be connected to the serial port or parallel port or USB. The serial port was chosen because it could be implemented more easily by connecting two wires. Also the maximum theoretical speed of 200kbps for our device can be handled by RS232. The data transmitted from the serial port uses RS-232 and this requires a 5v-power supply. Once again the program for serial communications is used to receive the data that is detected by the circuit. The serial communications software operates on both the transmitting and receiving computers and proper functionality of the circuit can be verified.

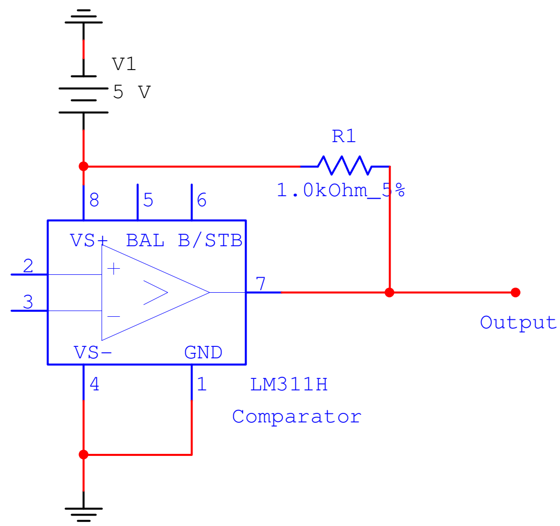


Figure 7.3 LM 311 comparator circuit (by Electronics Workbench Multisim V7.0)

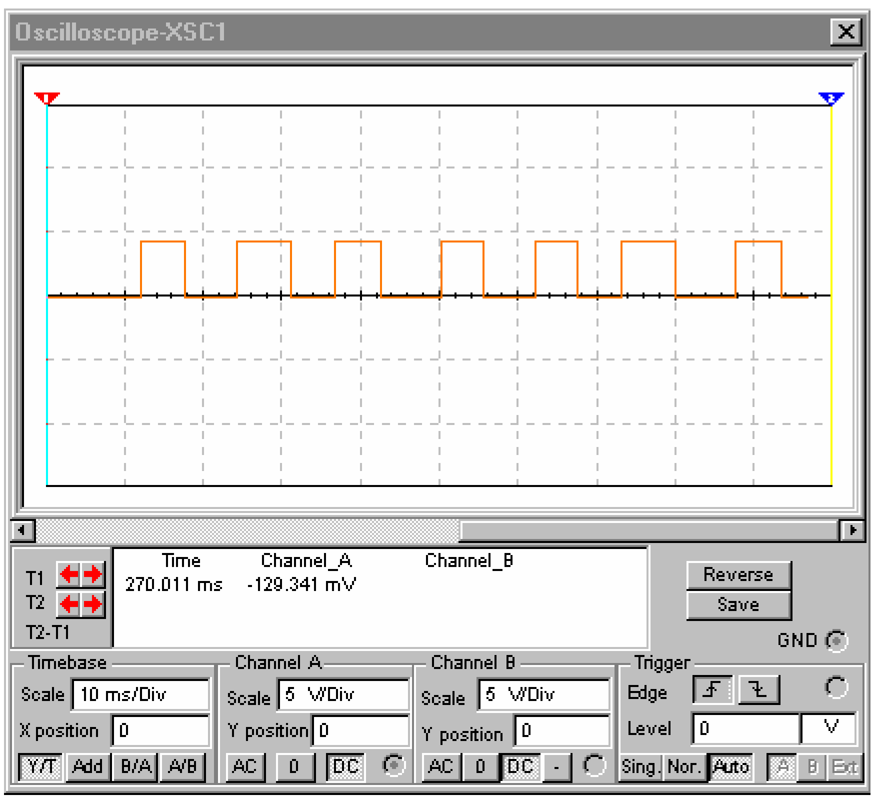


Figure 7.4 Typical output of LM311 comparator circuit

Absolute Maximum Ratings for the LM311

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

Total Supply Voltage (V_{84})	36V
Output to Negative Supply Voltage (V_{74})	40V
Ground to Negative Supply Voltage (V_{14})	30V
Differential Input Voltage	±30V
Input Voltage (Note 13)	±15V
Power Dissipation (Note 14)	500 mW
ESD Rating (Note 19)	300V
Output Short Circuit Duration	10 sec
Operating Temperature Range	0° to 70°C
Storage Temperature Range	−65°C to 150°C
Lead Temperature (soldering, 10 sec)	260°C
Voltage at Strobe Pin	V^{+}−5V
Soldering Information	
Dual-In-Line Package	
Soldering (10 seconds)	260°C
Small Outline Package	
Vapor Phase (60 seconds)	215°C
Infrared (15 seconds)	220°C

See AN-450 "Surface Mounting Methods and Their Effect on Product Reliability" for other methods of soldering surface mount devices.

Electrical Characteristics

for the LM311

Parameter	Conditions	Min	Typ	Max	Units
Input Offset Voltage (Note 16)	T_A=25°C, R_S≤50k		2.0	7.5	mV
Input Offset Current(Note 16)	T_A=25°C		6.0	50	nA
Input Bias Current	T_A=25°C		100	250	nA
Voltage Gain	T_A=25°C	40	200		V/mV
Response Time (Note 17)	T_A=25°C		200		ns
Saturation Voltage	V_{IN}≤−10 mV, I_{OUT}=50 mA T_A=25°C		0.75	1.5	V
Strobe ON Current (Note 18)	T_A=25°C		2.0	5.0	mA
Output Leakage Current	V_{IN}≥10 mV, V_{OUT}=35V T_A=25°C, I_{STROBE}=3 mA V^- = Pin 1 = −5V		0.2	50	nA
Input Offset Voltage (Note 16)	R_S≤50K			10	mV
Input Offset Current (Note 16)				70	nA
Input Bias Current				300	nA
Input Voltage Range		−14.5	13.8,−14.7	13.0	V
Saturation Voltage	V^+≥4.5V, V^-=0 V_{IN}≤−10 mV, I_{OUT}≤8 mA		0.23	0.4	V
Positive Supply Current	T_A=25°C		5.1	7.5	mA
Negative Supply Current	T_A=25°C		4.1	5.0	mA

Figure 7.5 Electrical characteristics of LM311 (Web_7, 2005)

CHAPTER EIGHT
CONCLUSION

8.1 Highlights

In recent years more and more users expect to have the ability to operate their portable computer globally while remaining connected to the communications network and service providers. So far there are various WLAN, which have been implemented to meet these demands. When designing such a network a number of factors have to be considered. These are range, maximum bit rate, portability, power consumption and cost. This thesis covers the specifications, design and simulation of a system that addresses these factors.

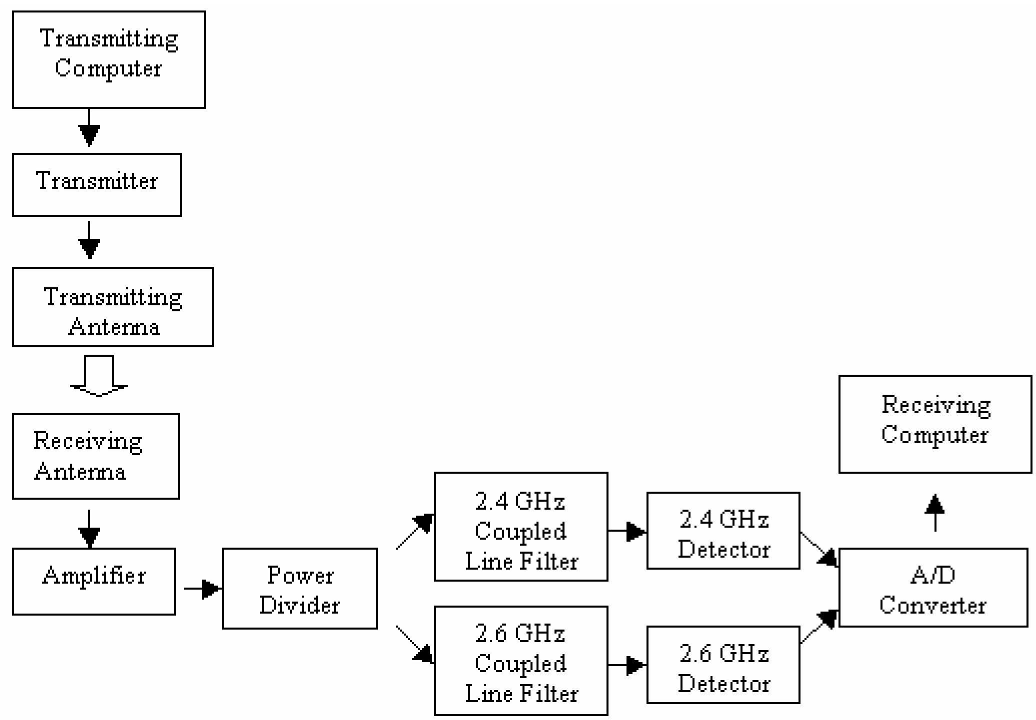


Figure 8.1 System level diagram a wireless local area network (Web_4, 2005)

The FSK receiver design with microstrip structure is described in this project. A binary 0 is represented by a frequency of 2.4 GHz while a binary 1 is represented by a frequency of 2.6 GHz. Microstrips are widely used because of its simplicity and its low fabrication cost. Microstrip circuits became popular because they are planar, small, easy and fast to make, and cheap. However, they cannot handle very high power levels and they are more lossy than coax or waveguide.

In all microstrip structures, the length and width of the microstrip lines was found by using TX-Line in the AWR-Microwaveoffice software package. Width and length values of each microstrip lines were calculated according to impedance and frequency values.

The system consists of amplifier, power dividers, bandpass filters, detectors, and voltage comparators as shown in Figure 8.2. The functions of the receiver are to demodulate the signal that has been received from the antenna and then transmit this signal to the PC. But before the signal is demodulated, it has to be filtered by the antenna and then amplified. After amplification, power divider is used since equal power is desired on both branches of the output circuit. The unwanted signal is filtered out with a BPF. The signal is then passed into a quadrature detector. A quadrature detector is used to demodulate frequency shift keying (FSK) and other types of FM signals and then voltage comparator compares output voltage of detectors. Finally the data is transmitted to the PC using RS-232. Figure 8.2 shows the physical layout of the receiver system.

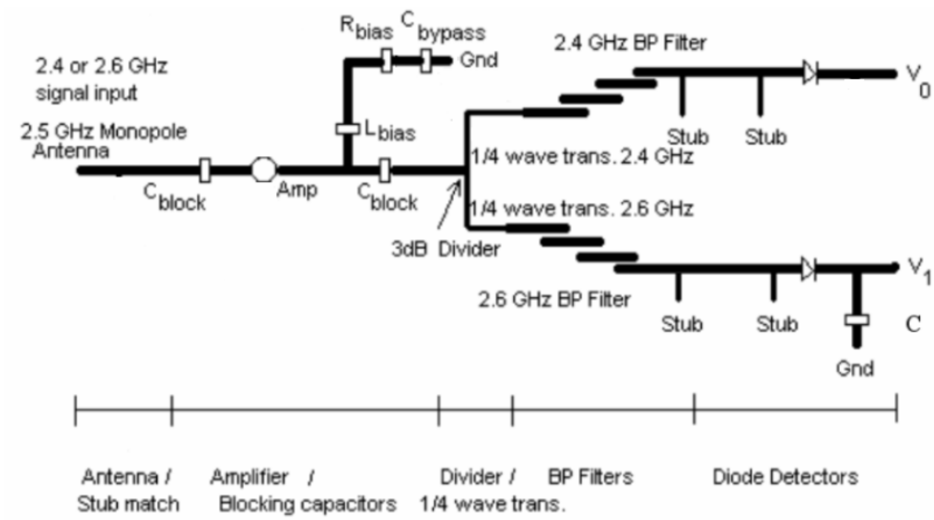


Figure 8.2 Wireless local area network receiver layout (Furse, Woodward, & Jensen, 2004)

Amplification is necessary in a wireless network. The reduction in the signal due to losses during transmission, reception, and power dissipation in circuit components must be compensated by using an amplifier to provide sufficient gain for the receiver circuit. Some real world issues when choosing an amplifier are its cost, size, and gain. The gain of the amplifier element correlates with the range that can be obtained by the link. So the ERA-3SM amplifier from MiniCircuits that has gain of 14 dB at the two operating frequencies was selected. AWR simulation result of S_{21} parameter of amplifier circuit proved that we obtained gain of 14 dB at operating frequency. To avoid affecting the amplifier's tuning and matching characteristics, the blocking capacitor was used. DC blocking capacitors are necessary on the input and output of the amplifier to block DC voltages and allow the RF signal to pass. But it must have a low impedance at all operating frequencies.

The power splitter (3-dB coupler) is the next sequential component in the circuit. The input of the circuit needs to be matched to the characteristic impedance of 50 ohms. Equal power is desired on both branches of the output circuit. As an optional constraint, each branch of the circuit can be tuned to the frequency that will eventually be detected on that branch, 2.4 and 2.6 GHz, by using quarter wave transformers. The 3 dB power splitter is a simple T-divider constructed from microstrip line with a 50Ω input and two

100 Ω output lines. The 100 Ω lines must be matched to the 50 Ω input impedance of the subsequent filters; therefore, quarter-wave transformer sections were used. The impedances of the microstrip lines used to realize the quarter-wave transformers change very little from 2.4 to 2.6 GHz; however, the length (quarter wavelength) changes significantly. Thus, two different transformers were designed, one for 2.4 GHz and one for 2.6 GHz. Each transformer was connected to an arm of the power splitter. The length of the 100 ohm line can be zero since a quarter wave transformer can be placed on each branch immediately after the input line is split. The impedance for the quarter wave transformer is found by taking the square root of the 100 ohm impedance times the 50 ohm impedance producing 70.71 ohms. The values were obtained by a line with the proper width as dictated by using TX-Line in the AWR-Microwaveoffice software package.The parameters of interest are S11, S21, and S31. the reflection coefficient parameter, S11 is a measure of how well the input of the power splitter is matched. The responses were tuned by shortening or lengthening the quarter wave sections until a low (less than or equal to –20 dB) reflection coefficient was attained, it is achieved -36.9 dB and transmission coefficients with values near –3 dB were attained.

Two band-pass filters were required for the project. One must pass 2.4 GHz and reject 2.6 GHz, and the other must pass 2.6 GHz and reject 2.4 GHz. When designing the filters, normalized frequencies and impedances were used. The parameters for fifth order coupled line filters that meet the specifications for this project were selected from Table 5.1. These values were used in the AWR MWO software package to find the lengths, widths, and separations for each of the respective coupled line sections. Each filter section is one quarter wavelength long. According to Δ=5% (fractional bandwidth of the passband), ripple of 0.5 dB, Z_0=50 Ω and N=5, it was obtained the even and odd mode coupled line characteristic impedances Z_{0e} and Z_{0o}. The parameter of interest is the transmission coefficient, S_{21}, from the input port to the output port for each filter. As can be seen in S_{21} parameter, the filters have smooth responses and steep falloffs outside the pass-band. TX-Line Program in AWR was used to calculate the dimensions of the copper strips on substrates. Circuit simulations were performed for two filters. This substrate offered the minimum power loss in the passband.

The goal of the detector circuit is to convert a high frequency signal into a DC voltage. We have used this in our WLAN to measure the voltage after the filters to determine which frequency was passed (corresponding to a 1 or a 0). In this case, a diode rectifier and integrating capacitor are used to meet this goal. An effective matching circuit is needed to prevent reflections and maximize the power input to the diode detector circuit. Detector design was realized using low cost, plastic packaged, two silicon Schottky diodes (HSCH-5338^{TM}). The module needs to have minimum component count and extremely low power new developments in low-barrier Schottky diodes suggested the viability of a detector with zero DC bias. The mirostrip substrate used in this circuit is RT Duroid 5880^{TM}, which have the following parameters: {dielectric constant ε_r = 2.2, dielectric thickness h = 1.575 mm, conductor (copper) thickness for each side t = 0.036 mm}. Self bias was chosen for all the diodes in order to achieve the highest possible voltage sensitivity. Input matching requirements were realised by using a microstrip line with a short circuit stub located the diode (the initial chosen length for the open circuit stub is 12.31 mm and the initial width is 12.76 mm for 2.4 GHz). With an input power of 0 dBm, an output voltage of at least 1V is expected for both detectors. When it is given it a frequency of 2.4 or 2.6 GHz (depending on the detector) and a magnitude of 0 dBm, we saw DC voltages of about 1.373 V and 0.772 V on the DC voltmeter for 2.4 GHz and 2.6 GHz respectively. The important parameter to measure is S_{11}. Its value should be on the order of –20 dB. As shown in Figure 6.13 and 6.16, the values are on the order of –35.22 dB and –27.97 dB at 2.4 GHz and 2.6 GHz respectively.

A comparator was used in this communication system to detect a shift between the outputs of the detector when a shift in frequency occurs. LM311 was chosen as a comparator. LM311 was chosen because of its price and its response time is sufficient for maximum bit rate of 200kbps. LM 311 detector circuit and its simulation result were designed with Electronics Workbench Multisim V7.0 software. The outputs of each detector circuit were connected to the input terminals of a comparator (LM 311). Outputs of detectors are DC voltage, so LM 311 compares these DC voltages. Output voltages of integrated system are 1.21V DC at 2.4GHz for 2.4GHz detector and 1.87V DC at 2.6GHz for 2.6GHz detector. These detector outputs were connected with legs of number 2 and 3 of LM311. the output is the digital data, which will be uploaded to the PC. The circuit will effectively produce a 0 or 1 depending on whether a frequency of 2.4 GHz or 2.6 GHz is received and detected in the circuit. For our device to communicate with the PC, it must be connected to the serial port or parallel port or USB. The serial port was chosen because it could be implemented more easily by connecting two wires. Also the maximum theoretical speed of 200kbps for our device can be handled by RS232. The data transmitted from the serial port uses RS-232 and this requires a 5V power supply. Once again the program for serial communications is used to receive the data that is detected by the circuit. The serial communications software operates on both the transmitting and receiving computers and proper functionality of the circuit can be verified.

8.2 System Integration and Simulation

All of the parts have been integrated and simulated. Integration and simulation is done by AWR-Microwaveoffice software package. Simulation results are explained by S-parameters. Figure 8.3 shows integrated system schematic.

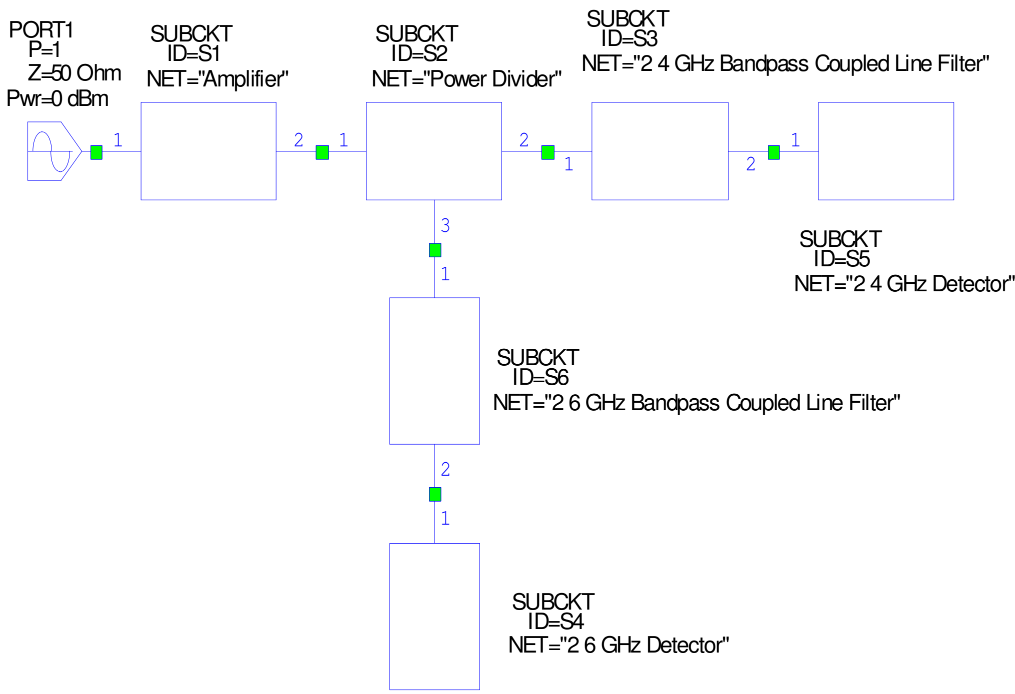


Figure 8.3 Integrated system schematic

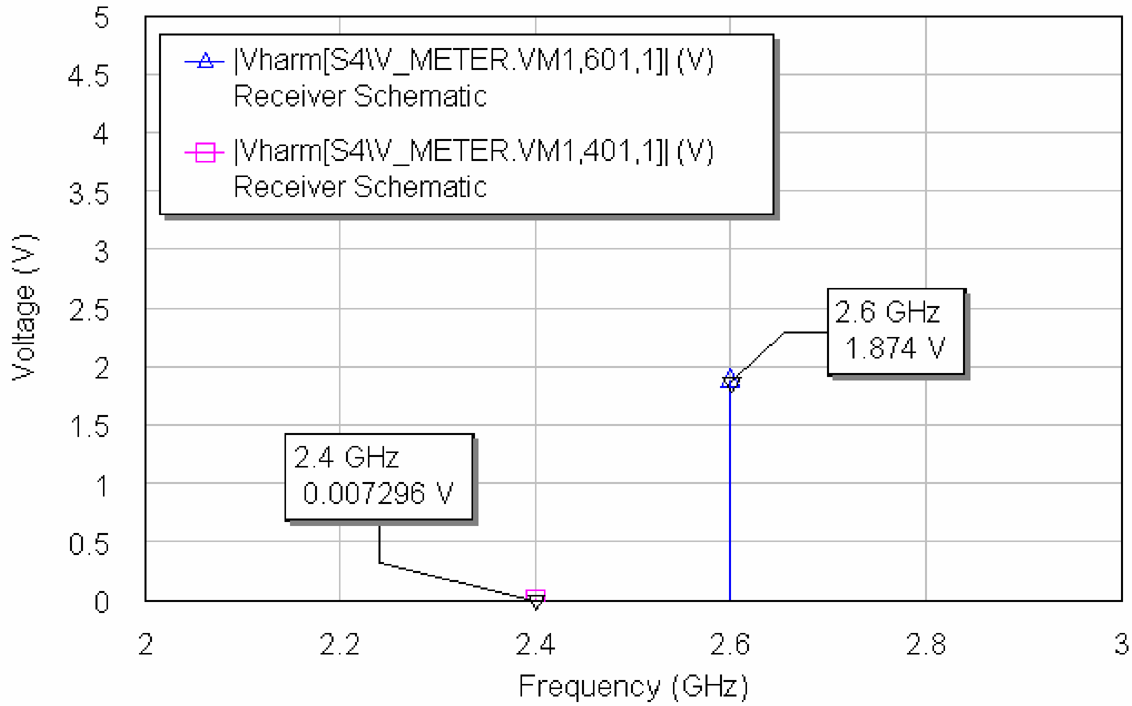


Figure 8.4 Output voltage of integrated system from 2.6 GHz detector

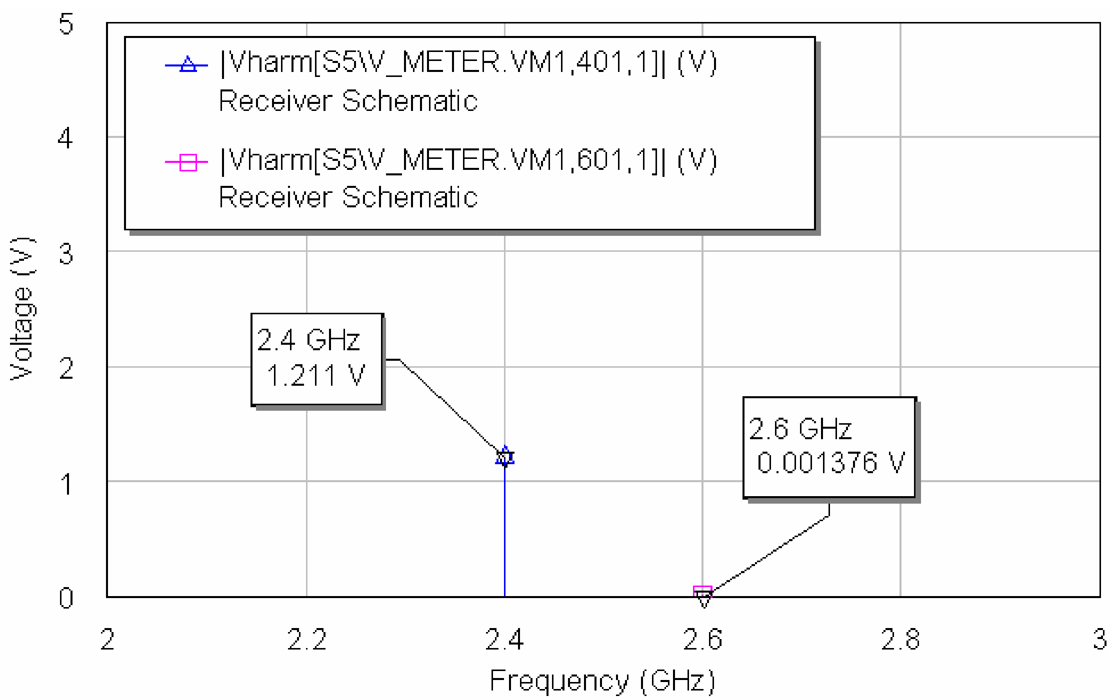


Figure 8.5 Output voltage of integrated system from 2.4 GHz detector

When signal of 0dBm reaches to system input, DC voltage values that are achieved from detector outputs of system are in Figure 8.4 and 8.5. When arriving signal has frequency of 2.4GHz, output voltage was fixed as 1.211V at 2.4GHz detector output and 0.001376V at 2.6GHz detector output. In the same way, with signal of 2.6GHz, 0.007296V at 2.4 GHz detector output and 1.874V at 2.6 GHz detector output were shown. These output values were transmitted voltage comparator and then, digital signal that is sent from transmitter was achieved again from comparator output.

8.3 Prices of Components and Dimensions of Circuits

Table 8.1 Dielectric constants of substrates and dimensions of circuits

	Dielectric Substrate (Er)	Dimensions of circuit (mm)	
		Width	Length
Amplifier	**3.27**	**40.4**	**95.56**
Power Divider	**3.54**	**25.61**	**41.92**
2.4GHz BPF	**3**	**34.92**	**98**
2.6GHZ BPF	**3**	**31.73**	**98**
2.4GHz Detector	**2.2**	**50.22**	**49.93**
2.6GHz Detector	**2.2**	**47.74**	**48.74**

Microstrip structures were used because they are planar, small, easy and fast to make, and cheap. As shown in Table 8.1 and 8.2, Our project has low dimension and cost.

Table 8.2 Prices of components and total cost of receiver (Component References)

AMPLIFIER	Picture	Vendor	Model Number	Quantity	Price
Laminate		Rogers Corporation	TMM3	40,4 x 95,56 mm	$ 6,53
Capacitor	 100pf	The Electronic Goldmine	G206C	2	$1.00
Capacitor Biasing		American Technical Ceramics	SMD100	1	$0.70
Amplifier		Mini-Circuits	ERA-3SM	1	$ 2.15
RF Choke inductor	47 nH	Mini-Circuits		1	$0.770
Resistor	 243ohm	The Electronic Goldmine	G241R	1	$1.00
TOTAL					**$13,15**
POWER DIVIDER	**Picture**	**Vendor**	**Model Number**	**Quantity**	**Price**
Laminate		Rogers Corporation	TMM3	25,61 x 41,92 mm	**$1,81**
2.4GHz BPF	**Picture**	**Vendor**	**Model Number**	**Quantity**	**Price**
Laminate		Rogers Corporation	R03003	34,92 x 98 mm	**$ 3,08**
2.6GHZ BPF	**Picture**	**Vendor**	**Model Number**	**Quantity**	**Price**
Laminate		Rogers Corporation	R03003	31,73 x 98 mm	**$ 2,79**
2.4GHz Detector	**Picture**	**Vendor**	**Model Number**	**Quantity**	**Price**
Laminate		Rogers Corporation	RT/durid 5880	50,22 x 49,93 mm	$ 2,66
Schottky Diode		Agilent Technologies	HSCH 5338	1	$0.22
TOTAL					**$2,88**
2.6GHz Detector	**Picture**	**Vendor**	**Model Number**	**Quantity**	**Price**
Laminate		Rogers Corporation	RT/durid 5880	47,74 x 48,74 mm	$ 2,47
Capacitor		American Technical Ceramics	SMD100	1	$0.70
Schottky Diode		Agilent Technologies	HSCH 5338	1	$0.22
TOTAL					**$3,39**
RECEIVER TOTAL					**$27,1**

8.4 Future Improvements

This project has many possibilities for improvements that can be performed. The antennas that would be used in the project function as designed, but a more directional set of antennas would be desirable. The design is modular enough that antennas can be exchanged easily and tested with the circuitry. Another upgrade is to add an additional amplifier to the circuit. With more gain, the range of the wireless local area network can be improved. Another modification is to use an amplifier between the transmitter and the transmitting antenna to add more power and increase the range of the circuit.

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