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أنا الموقع أدناه مقدم الرسالة التي تحمل العنوان:

Design and Development of Filtering Microstrip Antenna for 2.4GHz Applications

تصميم وتطوير مرشحات شريحية هوائية دقيقة لتطبيقات تردد 2.4 جيجا هرتز

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Student's name: Abdusslam M. A. Elastal

اسم الطالب: عبدالسلام محمد عبدالسلام الأسطل

Signature:

التوقيع:

Date: 31/10/2015

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Electrical Engineering Department

Communication Systems Engineering

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Abdussalam M. Elastal

Supervised by

Dr. Talal F. Skaik

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بناءً على موافقة شئون البحث العلمي والدراسات العليا بالجامعة الإسلامية بغزة على تشكيل لجنة الحكم على أطروحة الباحث/ **عبد السلام محمد عبد السلام الأسطل** لنيل درجة الماجستير في **كلية *الهندسة*** قسم **الهندسة الكهربائية – أنظمة الاتصالات** وموضوعها:

تصميم وتطوير مرشحات شريحية هوائية دقيقة لتطبيقات تردد 2.4 جيجا هرتز

Design and Development of Filtering Microstrip Antenna for 2.4GHz Applications

وبعد المناقشة التي تمت اليوم الاثنين 29 رجب 1436هـ، الموافق 2015/05/18م الساعة الثانية عشرة ظهراً بمبنى القدس، اجتمعت لجنة الحكم على الأطروحة والمكونة من:

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واللجنة إذ تمنحه هذه الدرجة فإنها توصيه بتقوى الله ولزوم طاعته وأن يسخر علمه في خدمة دينه ووطنه.

والله ولي التوفيق ،،،

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ج – 15

د.

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أ.د. فؤاد علي العاجز

P.O. Box 108, Rimal, Gaza, Palestine fax: +970 (8) 286 0800 فاكس Tel: +970 (8) 286 0700 هاتف ص.ب. 108 الرمال، غزة، فلسطين
public@iugaza.edu.ps www.iugaza.edu.ps

بسم الله الرحمن الرحيم

"قل إن صلاتي ونسكي ومحياي ومماتي لله رب العالمين"

صدق الله العظيم

الأنعام –آية 162

DEDICATION

To

My Parents

My Wife

and

My Family

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ABSTRACT

A microstrip antenna is considerable narrowband antenna and a method to increase its bandwidth as well as adding filtering property is studied here. In this research three models are developed for 2.4 GHz band applications using filtering antenna concept that increases the bandwidth of the antenna and allows it to work as a filter to improve its frequency selectivity.

In the first model, the antenna and the filter are designed separately then they are integrated into one unit. The rectangular micristrip antenna is designed with quarter-wave transformer for matching and the filter is a third order Chebyshev filter. Integrating the filter and antenna together makes the filter acts as a part of the antenna and the antenna acts as a new resonator in the filter. This model enhanced the bandwidth by about 145% and improved the frequency selectivity of the antenna.

Other models are also designed with reduced size in comparison to the first model. The quarter-wave transformer is removed in these models and the matching is achieved by connecting the filter and antenna directly by a transmission line with adjustable length and width. Then the matching is achieved by varying the length and width of the transmission line or by adding a tunable shunt stub.

Keywords:

2.4 GHz band, Antenna, Bandwidth, Filter, Filtering antenna, Frequency selectivity, Hairpin filter, Rectangular microstrip antenna, Rectangular patch antenna.

ARABIC ABSTRACT

ملخص الدراسة

النطاق الترددي للهوائيات الشريحية الدقيقة صغير. وهذه دراسة لطريقة لزيادة النطاق الترددي مع إضافة خاصية التصفية للهوائي. في هذا البحث تم تطوير ثلاث نماذج لتطبيقات تردد 2.4 جيجا هرتز باستخدام مبدأ المرشحات الهوائية وذلك لزيادة النطاق الترددي للهوائي والسماح له بالعمل كمرشح لتحسين الانتقائية الترددية.

في النموذج الأول تم تصميم الهوائي والمرشح كل على حدة ثم تم دمجهما معا كوحدة واحدة. بداية تم تصميم الهوائي الدقيق مستطيل الشريحة مع محول ربع موجي لغرض التوافق الترددي، أما المرشح من نوع Chebyshev من الدرجة الثالثة. دمج الهوائي والمرشح معا جعل المرشح يعمل كجزء من الهوائي بينما عمل الهوائي كرنّان جديد في المرشح. هذا النموذج أظهر تطويرا في النطاق الترددي بحوالي 145% كما حسن من انتقائية الهوائي.

النماذج الأخرى تم تصميمها بتصغير الحجم بعد إزالة المحول الربع موجي ثم إعادة التوافق بتوصيل المرشح والهوائي معا بواسطة خط نقل قابل لتعديل طوله أو عرضه. بعدها تم إضافة أغصان لتحقيق التوافق. هذه الأغصان قابلة لتعديل طولها وعرضها.

كلمات مفتاحية:

مرشح هوائي، هوائي، هوائي شريحي مستطيل دقيق، هوائي الرقعة المستطيلة، مرشح، مرشح دبوس الشعر، الانتقائية الترددية، النطاق الترددي، نطاق 2.4 جيجا هرتز.

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LIST OF ABBREVIATIONS

BPF	Band Pass Filter
BSF	Band Stop Filter
BW	Band Width
CST	Computer Simulation Technology
DECT	Digital European Cordless Telephone
DMS	Defected Microstrip Antenna
FBW	Fractional Band Width
FM	Frequency Modulation
FNBW	First Null Beam Width
HFSS	High Frequency Structure Simulator
HPBW	Half Power Beam Width
HPF	High Pass Filter
IEEE	Institute of Electrical and Electronics Engineers
ISM	Industrial, Scientific and Medical instrumentation
LH	Left Hand
LPF	Low Pass Filter
MSA	Micro-Strip Antenna
PAN	Personal Area Network
RF	Radio Frequency
RH	Right Hand
RLC	Resistance Inductance Capacitance
TEM	Transverse Electro-Magnetic
UWB	Ultra Wide-Band
Wi-Fi	Wireless Fidelity

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Chapter 1 : Introduction

The focus of this thesis is on the design and development of filtering microstrip antenna for 2.4 GHz applications. The antenna will have filtering features in addition to bandwidth enhancement by using a filter and an antenna as one device.

1.1 Communication system

Data communication can be generally defined as "The exchange of information by two devices via some transmission link." [1]. This exchange can be done directly like speaking and talking or using tools like electrical communication systems that use electrical signal to transmit and receive data, this system can be modeled as shown in Figure 1.1 [2]. The transmission link is also called channel. This channel can be wire, air, radio link, optical fibers and other medias. Communication systems can be classified according to channel type to wired channel or a kind/type of wireless channels, like mobile systems, satellite systems, broadcasting systems(like TVs and radios) . . . etc.

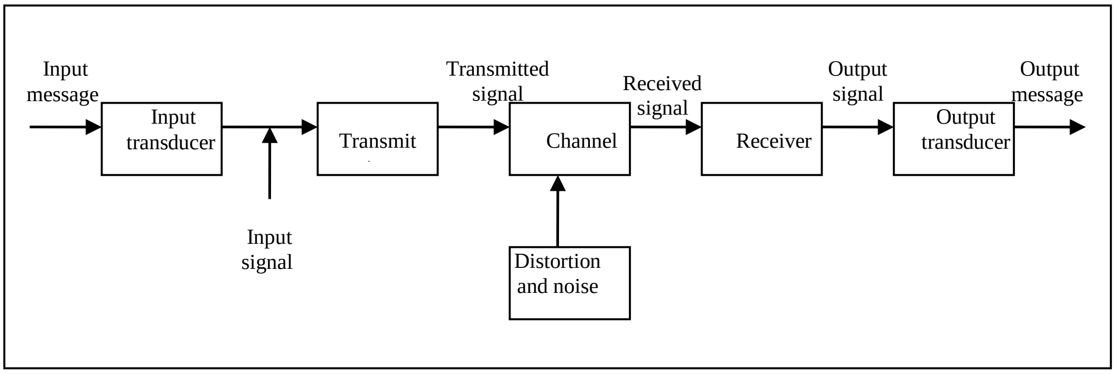

Figure (1.1) A communication system.

1.2 ISM band "2.4 GHz band"

A 2.4 GHz band is a term for one of unlicensed bands of frequencies from 2400 MHz to 2483.5 MHz that is also called ISM band. It is referred to "Industrial-instrumentation- Scientific and Medical” band.

ISM band can be used as unlicensed frequency that is used for a lot of services like Bluetooth, cordless telephones (DECT), baby monitoring, car alarms, microwave ovens, wireless personal area networks PAN, Wi-Fi … etc. However, Wi-Fi can also use the band 2300-2500 MHz.

A 2.4 GHz band channelization is shown in Figure 1.2 using IEEE 802.11b protocol.

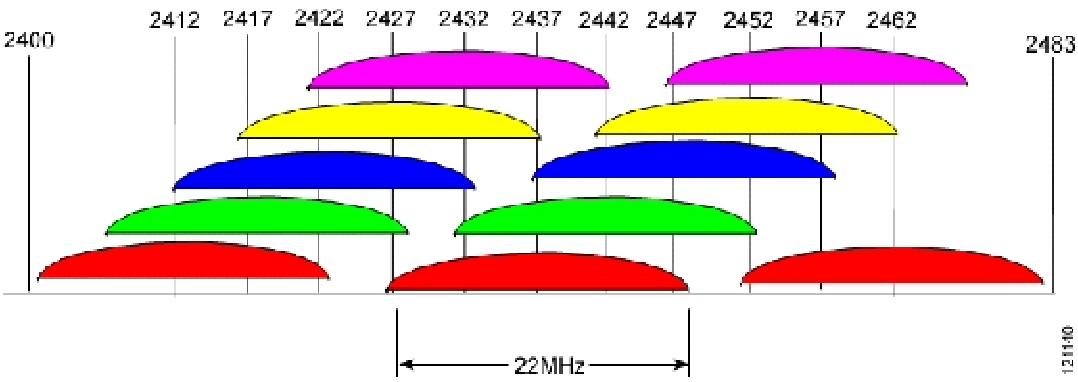

Figure (1.2): North American Channelization Scheme [4]

1.3 Antennas

The antenna is a device that transforms electrical signals into radio waves and vice versa. The IEEE standard definitions of terms of antenna (IEEE Std 145-1983) defines the antenna or aerial as "a means for radiating or receiving radio waves."[5].

Antennas can be classified into six major types[5]:

1. **Wire antennas** which are the most famous and popular kind of antennas that it can be seen almost everywhere (cars, buildings, ships, TVs, radios, some mobiles and so on. It can be in different shapes like straight wire (dipole), loop, and helix as shown in Figure 1.3

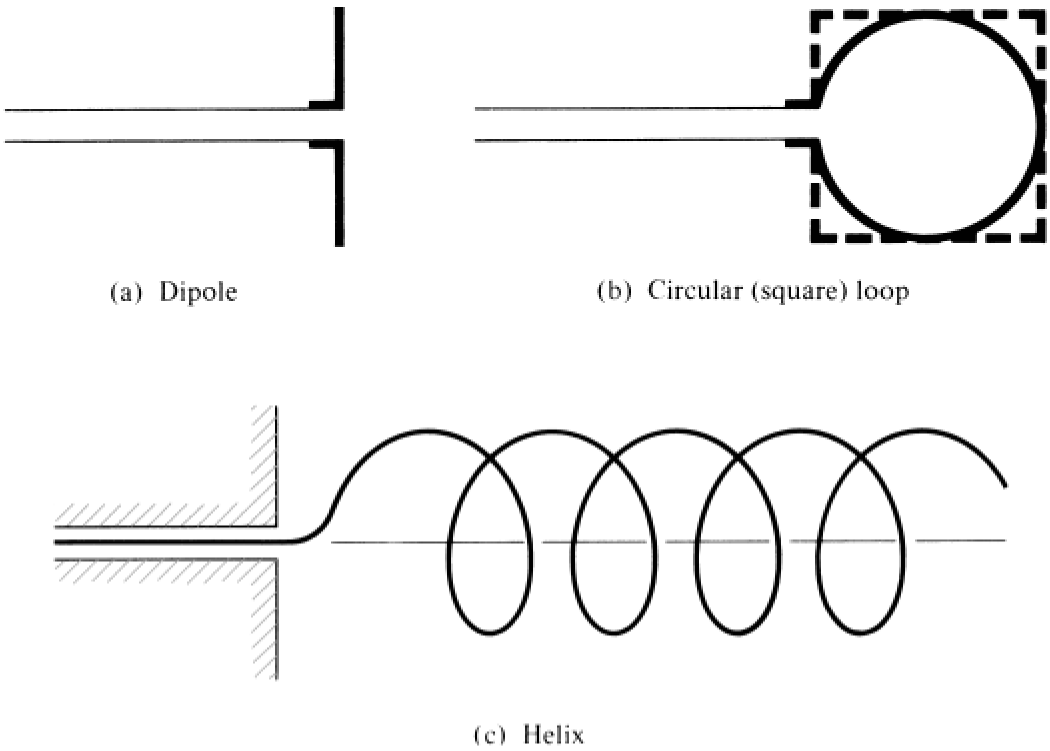

Figure (1.3): Wire antenna configuration [5].

2. **Aperture antennas** that are developed because of the need to more complex forms and usage for higher frequencies than wired antennas as shown in Figure 1.4. Waveguides is the famous aperture antenna. These antennas are very useful in aircraft and spacecraft applications because they can be covered with a dielectric material to protect them from dangerous conditions in high speed environments.

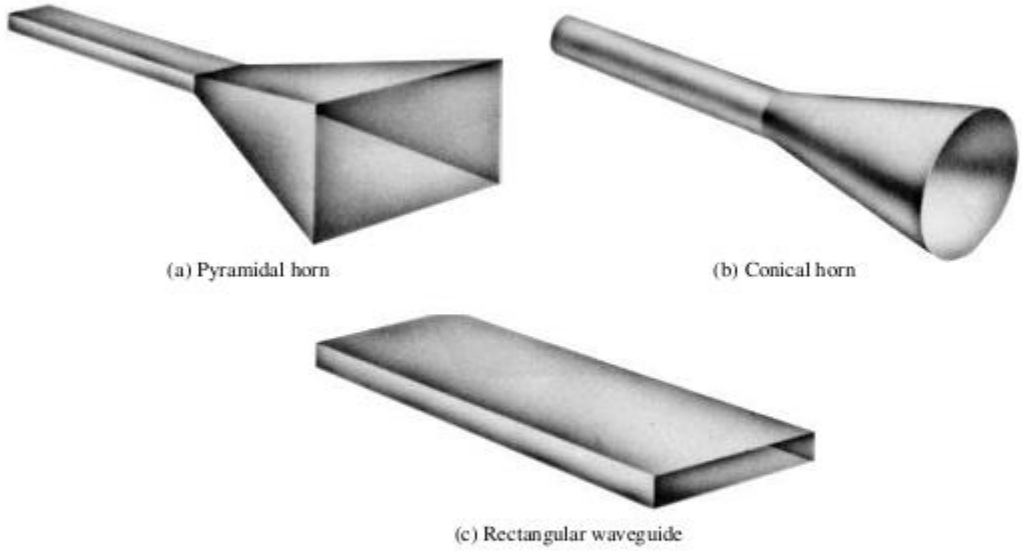

Figure (1.4): Aperture antenna configurations [5].

3. **Microstrip antennas** or patched antenna. A microstrip antenna "MSA" basically consists of two metallic sheets "patches" that are separated by a substrate, one of them is the ground and the other acts as a radiator as shown in Figure 1.5. This antenna becomes widely used because of low cost, easy to analyze and fabricate and their small size . It can be used on the surface of high performance aircraft, satellites, mobiles, missiles, and spacecrafts.

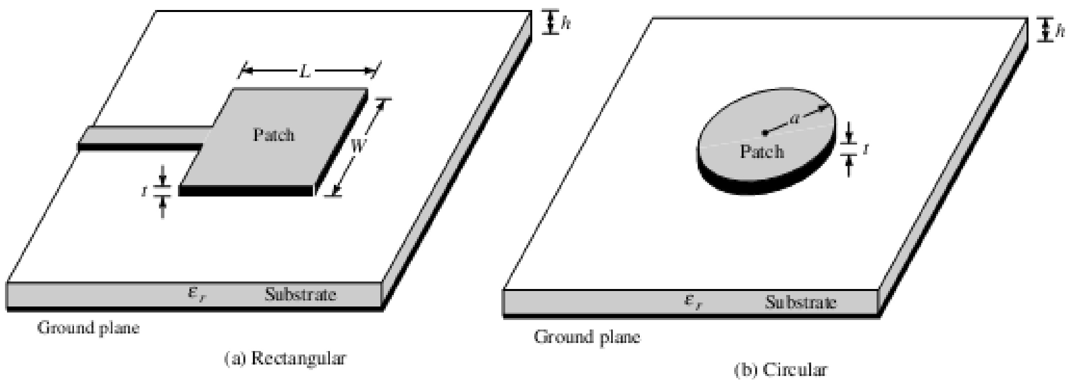

Figure (1.5): Rectangular and circular microstrip (patch) antennas [5].

4. **Array antennas** are a group of radiating elements arranged in a certain electrical and geometrical arrangement (an array) as shown in Figure 1.6 to achieve a certain radiation characteristic which may not be achieved in a single radiating element. This array also may increase radiation in to maximum in a particular direction or directions and decreasing in others as desired.

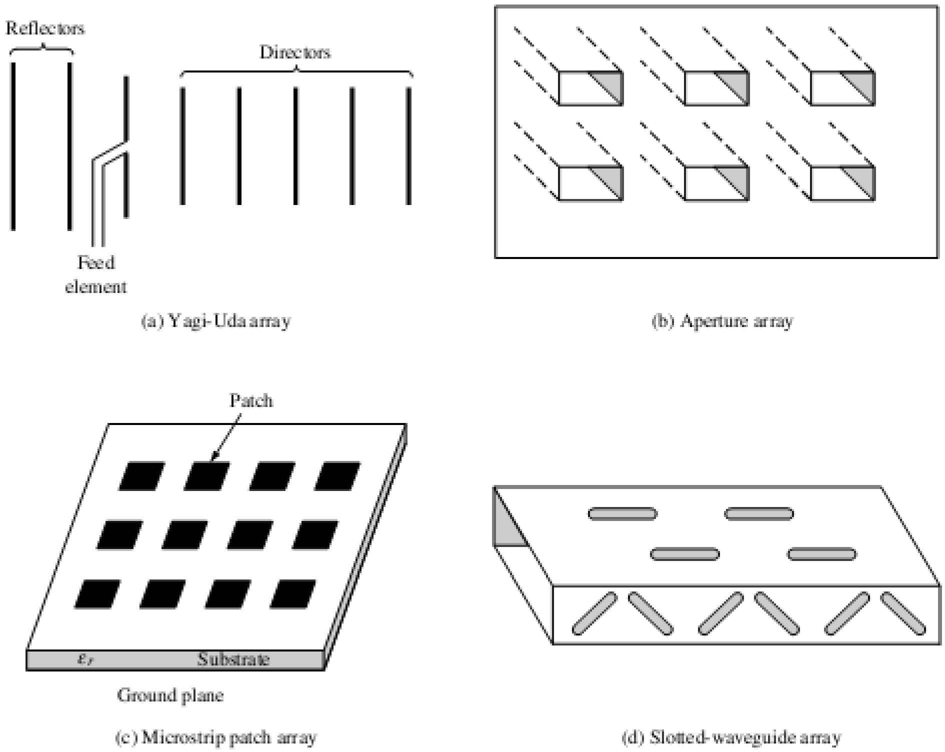

Figure (1.6): Typical wire, aperture, and microstrip array configurations [5].

5. **Reflector antennas** are primarily used for satellite communications because of the great distances. The parabolic antenna (reflector) is the most poplar form for this type of antennas. In this type, the larger antenna diameter, the higher gain can be achieved for transmitting and receiving. Their shapes are shown in Figure 1.7.

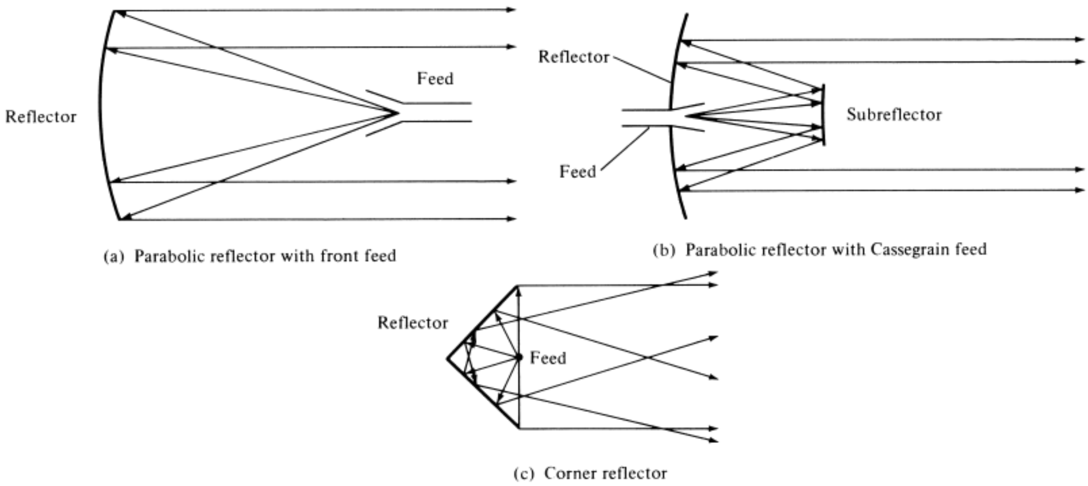

Figure (1.7): Typical reflector configurations [5].

6. **Lens antennas** are constructed of a dielectric material and can be used as a reflector antenna, to concentrate radiated energy in a certain direction or directions. They can be placed in front of a dipole on horn antenna to increase their directivity in the desired direction. They can be classified according to their material or their shapes. They are almost used for high frequencies. Some forms are shown in Figure 1.8.

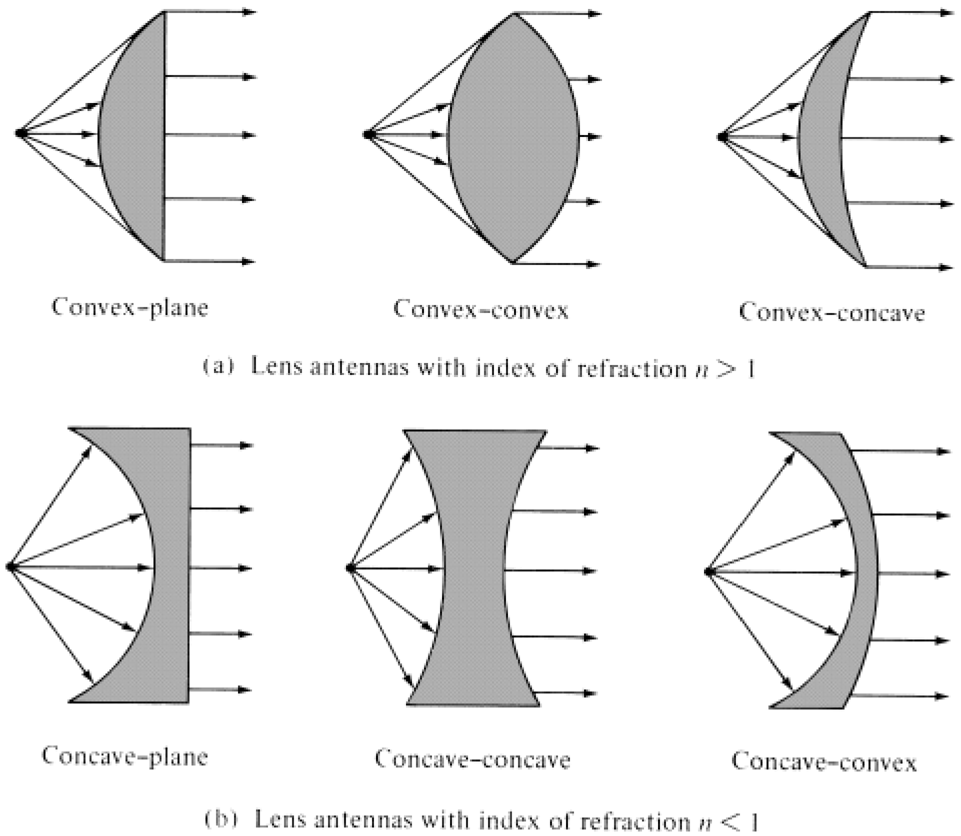

Figure (1.8): Typical lens antenna configurations [5].

1.4 Filters

Filters are microwave devices that have two ports one for input and the other for output. They allow some frequencies to pass from input to output and block others. Filters are commonly used in transceiver circuits.

They can be classified according to their frequency selection, their response, their technology and their frequency band. The classification of filters can be according to one of the following:

a- Selectivity: low pass filters (LPF), high pass filters (HPF), band pass filters (BPF) or band stop filters (BSF) as shown in Figure 1.9.

b- Response: Chebyshev, maximally flat, Elliptic, ...etc. as shown in Figure 1.10.

c- Technology: lumped, waveguide, microstrip, ...etc.

d- Frequency band: Narrow band or broadband.

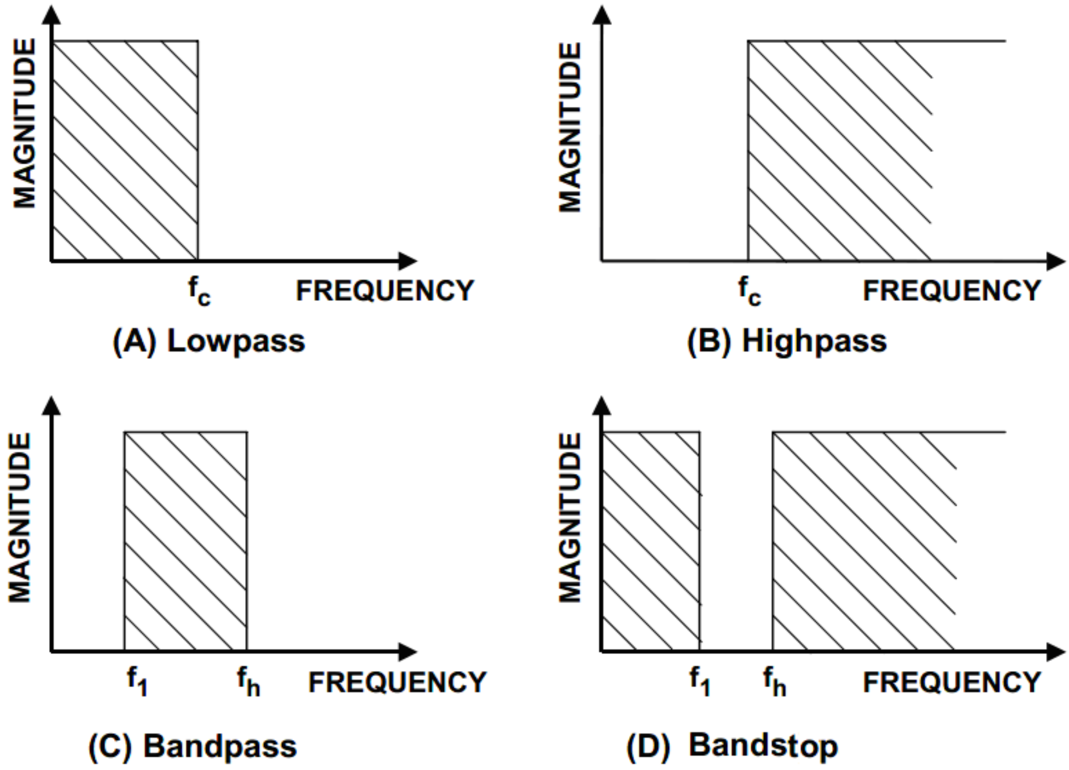

Figure (1.9): Typical Filter types according to selectivity [8].

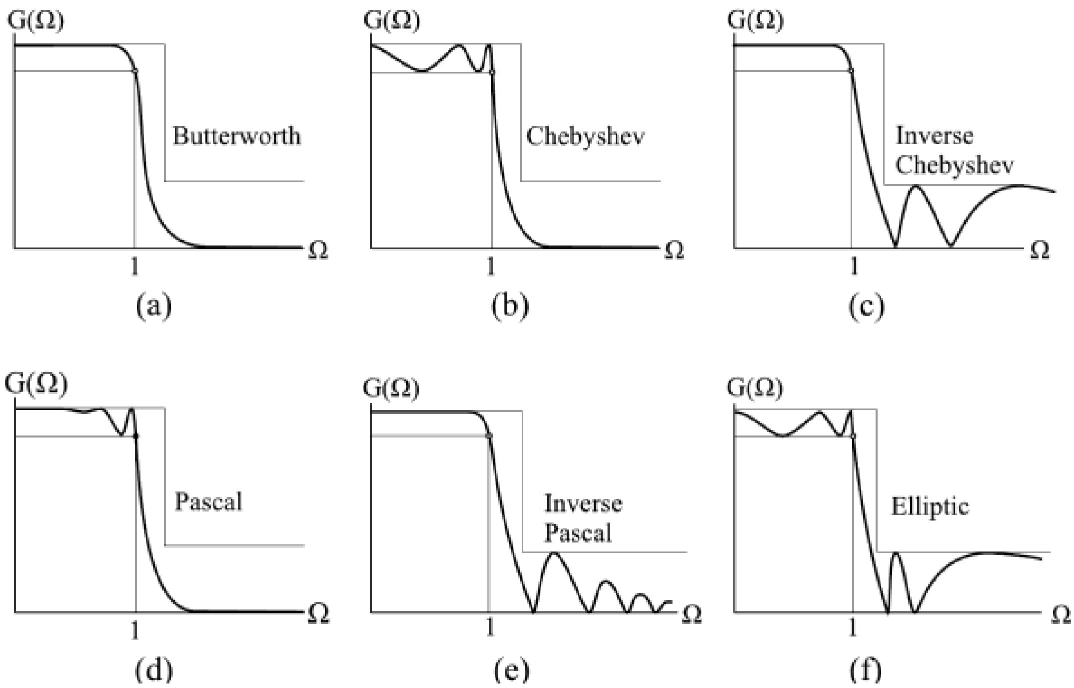

Figure (1.10): Typical Filter types according to filter response [9].

1.5 Filtering Antenna "Filtenna"

Finally filtering antenna is the integration between filters and antennas, where the antenna acts as the last resonator of the filter. Filtering antennas can be used to reject a certain band of frequencies or to enhance the bandwidth and selectivity. Filtering antennas can be designed in two ways: design the filter and antenna independently or using co-design both filter and antenna in the same time. Figure 1.11 shows a general block diagram for a filtering antenna structure in which the proposed design is based on. Bandpass filters are usually designed by using resonators coupled together as will be explained later in chapter 3.

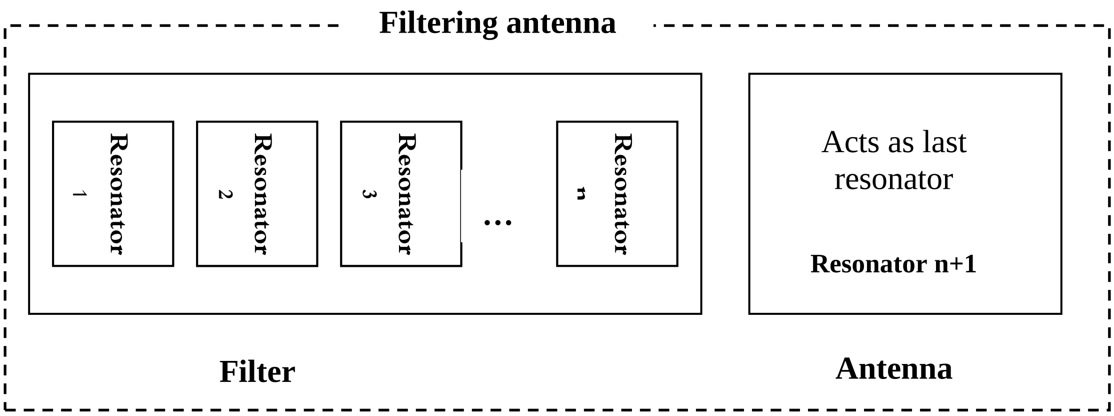

Figure (1.11): Block diagram of filtering antenna structure

1.5.1 Literature Review

Filtering antenna is an antenna with filtering features like rejecting certain band as shown in [10] that proposed low cost Ultra Wide Band (UWB) printed dipole antenna with filtering feature. They designed an UWB antenna and integrated a (4.9-5.9 GHz) band reject filter to avoid the interference with 5GHz Wi-Fi band. Filtering antennas can also be used to increase the bandwidth as in [11] the authors proposed a compact UWB filtering antenna to reduce the overall component size using coplanar waveguide as a band pass filter and a U-shape antenna.

Another design is presented in [12] using a meander line antenna and a shunt quarter wavelength resonator for 2.45 GHz to make bandwidth selectivity enhancement. The resulting filtering antenna is second order Chebyshev band pass filter with 0.1 dB equal ripple response.

In [13], a compact filtering antenna using defect ground resonator based on co-design method for a Γ-shape antenna with defect ground resonator are used and integrated to be a filtering antenna for 2.45 GHz second order Chebyshev filter with 0.1 dB equal ripple response.

Another compact simple structured filtering antenna in [14] is composed of inverted L antenna and quarter wavelength transmission line that is modeled by series RLC resonator and shunt LC resonator for 2.45 GHz with 20% fractional bandwidth (FBW).

Moreover, a printed filtering antenna is presented in [15]. The structure consists of an inverted L antenna and parallel coupled microstrip line sections as resonators of the filter for 2.45 GHz third order Chebyshev band pass filter.

Another compact printed filtering antenna is proposed in [16] using ground intruded coupled line resonator integrated with a Γ-shape antenna at 2.45 GHz.

In [17], authors presented design, fabrication and measurement of a compact filtering microstrip antenna with quasi-elliptic broadside antenna gain response using U-shape patch antenna and T-shape resonator through an inset coupling structure.

In [18], a filtering microstrip antenna array is presented, where the antenna elements together with very compact feed network function as third order bandpass filter. The feeding network consists of one power divider and two baluns provide the first two stages, and the micostrip antenna elements provide the last stage in the filter design. This will provide a third order filtering microstrip antenna array for 5 GHz with 3% fractional bandwidth and Chebyshev 0.3 dB equal ripple broadside antenna gain response.

Also in [19], a filtering antenna using coplanar waveguide CPW bandpass filter integrated with coplanar waveguide CPW wideband antenna is proposed.

In [20], an inset rectangular antenna is cascaded with a hairpin bandpass filter using a classical 50 ohm transmission line. The design procedure is based on the general coupling matrix theory to transform the antenna as a resonator in the filter, but the design bandwidth is 50 MHz that does not match the whole 2.4 GHz band that is described in section 1.2.

As in [11], [21,22] presented other designs to use filtering antenna as a band-stop for UWB antennas.

In [21], a reconfigurable band pass filter is integrated with an UWB antenna. The filter is T-shaped defected microstrip structure (DMS) and the reconfigurability is achieved by incorporating switches with in the filter structure.

In [22], a microstrip open loop resonator in integrated into UWB antenna to design a band-notched "stop" UWB antenna.

A filtering antenna has been designed at 2.45 GHz with a structure consisting of a rectangular patch antenna integrated with U-shaped resonators to form a fourth order filtering antenna.

1.6 Thesis Motivation

The importance of antennas and filters in communication systems leads to find methods to combine both into one part that may be called filtering antenna. This new combination "filtering antenna" needs many researches to assure the required characteristics or enhance the original ones. This thesis will study the possibility of using filtering antenna techniques to increase the bandwidth of the original antenna. The selected antenna is the rectangular patch antenna that is a very narrowband antenna, and the selected filter is a U-shape "Hairpin" filter that is an easy and compact design.

1.7 Thesis Overview

Chapter 2 discusses the theory of a rectangular microstrip "patch" antenna and the design equations and procedure.

Chapter 3 discusses the theory of a microwave filters and their design equations and procedure steps.

Chapter 4 describes filtering antennas and their design procedure and the required equations, tools and programs then shows the simulation results of the designed filtering antenna.

Chapter 5 shows the conclusion of the thesis and future work.

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Chapter 2 Microstrip Patch Antenna

This chapter will discuss antenna theory, antenna parameters and patch antenna design, equations and techniques.

2.1Antenna Theory

As shown in previous chapter, the antenna can be defined as a usually metallic device which radiates and receives electromagnetic waves (EM), more specifically, radio waves [1]. Another explanation says that the antenna is the transition between a guided EM wave and a free-space EM wave and vice-versa [2], this definition can be shown in Figure 2.1.

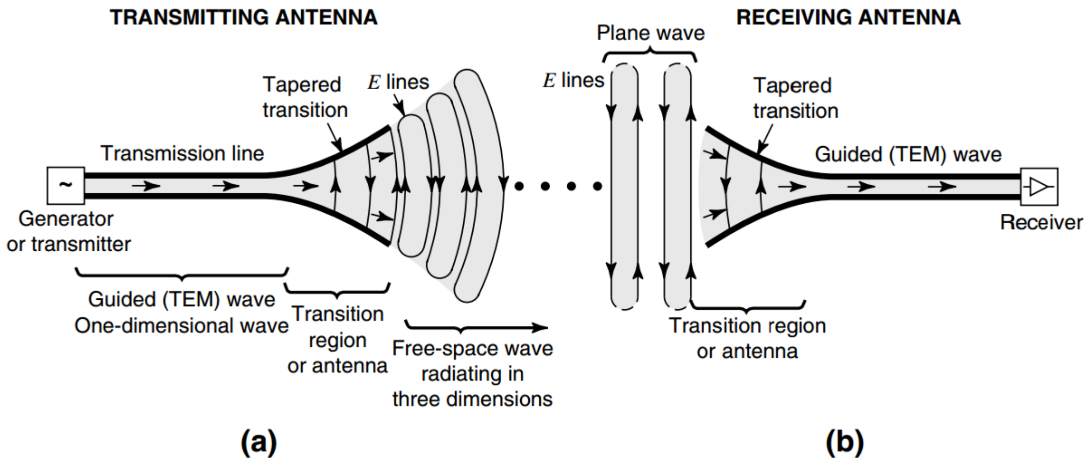

Figure (2.1): The antenna as a transition structure, for a transmitting antenna and for a receiving antenna [3].

As shown above, for both antennas, the transmission line has the form of a coaxial line or a waveguide. When the transmission line is connected to a transmitter that generates radio-frequency (RF), the antenna will be a transmitting antenna, where energy is guided through the uniform part of the line as a plane Transverse Electromagnetic (TEM) wave with little loss, transformed into a signal that is amplified, modulated and applied to the antenna. On the other hand, when the transmission line is connected to a receiver, the antenna will be a receiving antenna. The antenna will collect the alternating currents that resulted from the transformation process of the received radio waves by the antenna [3].

If a time-changing current or an acceleration (or deceleration) of charge occurs, the radiation will be created in a certain length of current element. This can be described by [1]:

$$l.\frac{dI}{dt} = l.q_l.\frac{dv}{dt}(A.m/s) \tag{2.1}$$

where:

l - Length of the current element in meters (m).

${}^{dI}/_{dt}$ - Time-changing current in ampere per second (*A/s*).

q_l - Charge per unit length (*coulombs/m*). Note that $q=I$ and $q_{electron}=1.602\times10^{-19}$ *coulombs*.

Furthermore, the radiation is always perpendicular to the acceleration and its power is proportional to the square of both parts of the equation (2.1). It is important to refer that the spacing between the two wires of the transition line is just a small part of a wavelength; therefore, the more the transition curve of the antenna opens out the more the order of a wavelength or more is reached; consequently, the more the wave tends to be radiated and launched into the free-space [3].

Looking at the antenna structure as a whole, the transition region of the antenna is like a radiation resistance (R_r) to the transmission line point of view, which represents the radiation that the antenna emits, analyzing it as a circuit. Figure 2.2 shows the complete circuit of an antenna; where the source is an ideal generator with a potential V_g and with an impedance Z_g; the transmission line is a line with characteristic impedance Z_0, and the antenna itself is represented by a load impedance Z_A [$Z_A=(R_L+ R_r) + jX_A$] connected to the transmission line. The load resistance R_L is used to represent the conduction and dielectric losses associated with the antenna structure while R_r, referred to as the radiation resistance, is used to represent radiation by the antenna. The reactance X_A is used to represent the imaginary part of the impedance associated with radiation by the antenna. Therefore, if ideal conditions are applied, the radiation resistance R_r, which is used to represent radiation by the antenna, will get all the energy that is generated by the transmitter (the source) [1].

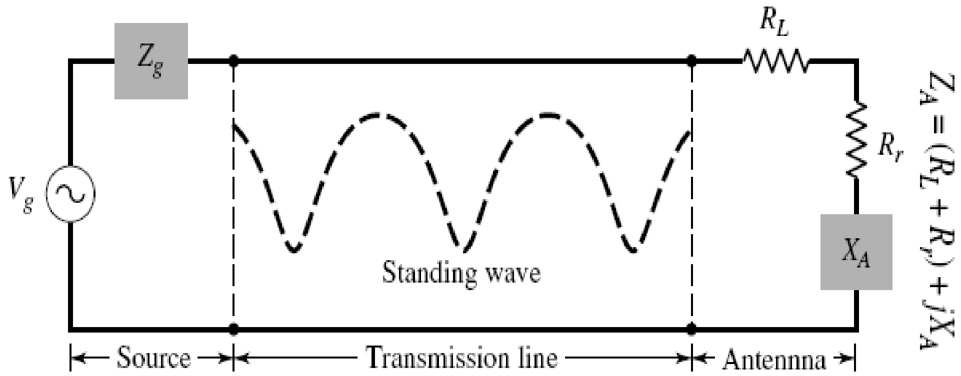

Figure (2.2): Circuit representing antenna as whole structure [1].

2.2Antenna Parameters

Antennas are defined by several parameters according to their constitution and shape. In this section, the most important are considered and explained, and an overview of each is essential to describe antenna’s performance.

2.2.1 Radiation Pattern

An antenna radiation pattern or antenna pattern is defined as “a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates. In most cases, the radiation pattern is determined in the far field region and is represented as a function of the directional coordinates. Radiation properties include power flux density, radiation intensity, field strength, directivity, phase or polarization.” [4].

A mathematical illustration of the radiation properties of an antenna as a function of the space coordinates defined by the spherical coordinates θ and ϕ is shown in Figure 2.3.

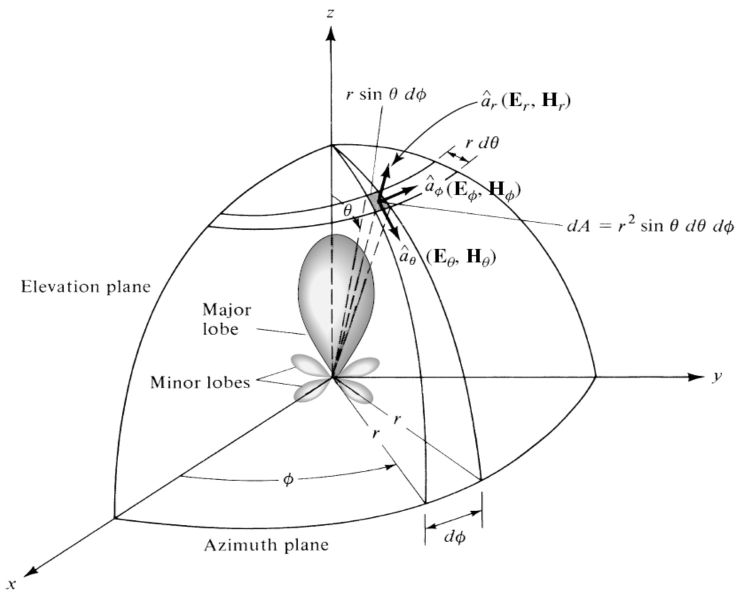

Figure (2.3): Coordinate system for antenna Analysis [4].

A trace of the received electric (magnetic) field at a constant radius is called the amplitude field pattern. On the other hand, a graph of the spatial variation of the power density along a constant radius is called an amplitude power pattern. Often the field and power patterns are normalized with respect to their maximum value, yielding normalized field and power patterns. Also, the power patterns are usually plotted on a logarithmic scale or more commonly in decibels (dB) [4].

Various parts of a radiation pattern are referred to as lobes, which may be sub-classified into major or main, minor, side, and back lobes as shown in Figure 2.4. A radiation lobe is a “portion of the radiation pattern bounded by regions of relatively weak radiation intensity.” [4].

The main lobe (or main beam or major lobe) is the lobe containing the direction of maximum radiation. There is also usually a series of lobes smaller than the main lobe. Any lobe other than the main lobe is called a minor lobe. Minor lobes are composed of side lobes and back lobes. Back lobes are directly opposite the main lobe, or sometimes they are taken to be the lobes in the half-space opposite the main lobe [5].

There are three common radiation patterns that are used to describe an antenna's radiation property:

• **Isotropic**- A hypothetical lossless antenna having equal radiation in all directions.

• **Directional**- An antenna having the property of radiating or receiving electromagnetic waves more effectively in some directions than in others.

• **Omnidirectional**- An antenna having an essentially non-directional pattern in a given plane and a directional pattern in any orthogonal plane.

Directional or omnidirectional radiation properties are needed depending on the practical application. Omnidirectional patterns are normally desirable in mobile and hand-held systems [4].

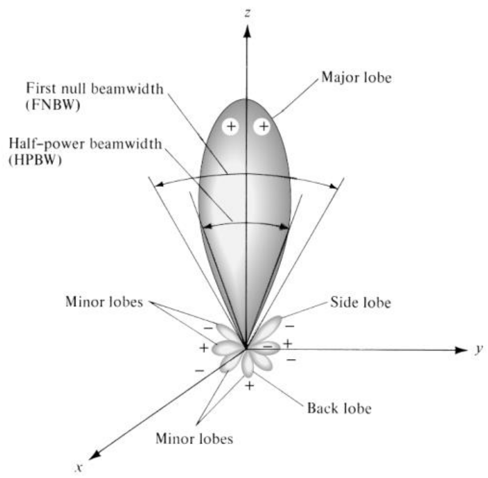

Figure (2.4): Radiation lobes and beamwidths of an antenna pattern [4].

2.2.2 Beamwidth

The beamwidth of a pattern definition is the angular separation between two identical points on opposite side of the pattern maximum. In an antenna pattern, there are a number of beamwidths. One of the most widely used beamwidths is the Half-Power Beamwidth (HPBW), which is defined by IEEE as: "In a plane containing the direction of the maximum of a beam, the angle between the two directions in which the radiation intensity is one-half value of the beam". The angular separation between the first nulls of the pattern is referred to as the First-Null Beamwidth (FNBW) as shown in Figures 2.4 and 2.5 [4].

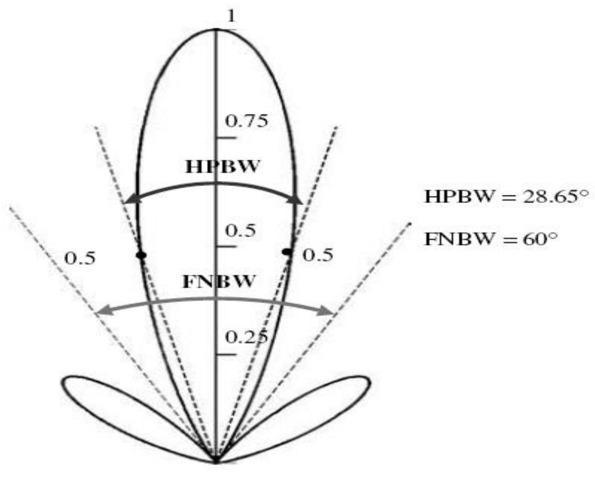

Figure (2.5): beamwidths of a directional antenna power pattern [4].

The beamwidth of an antenna is a very important issue and often is used as a trade-off between it and the side lobe level; that is, as the beamwidth decreases, the side lobe increases and vice versa. In addition, the beamwidth of the antenna is also used to describe the resolution capabilities of the antenna to distinguish between two adjacent radiating sources or radar targets [4].

2.2.3 Directivity

The directivity of an antenna is defined as "the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions." In other words, the directivity of a nonisotropic source is equal to the ratio of its radiation intensity in a given direction, over that of an isotropic source [4,5].

Radiation intensity is the power radiated in a given direction per unit solid angle and has units of watts per square radian (or steradian, *sr*). The advantage of using radiation intensity is that it is independent of distance *r* [5].

$$D = \frac{U}{U_i} = \frac{4\pi U}{P_r} \tag{2.2}$$

where D is the directivity of the antenna; U is the radiation intensity of the antenna; U_i is the radiation intensity of an isotropic source; and P_r is the total power radiated.

Sometimes, the direction of the directivity is not specified. In this case, the direction of the maximum radiation intensity is implied and the maximum directivity is given as

$$D_{\max} = \frac{U_{\max}}{U_i} = \frac{4\pi U_{\max}}{P_r} \tag{2.3}$$

where D_{max} is the maximum directivity and U_{max} is the maximum radiation intensity.

The directivity of an antenna can be easily estimated from the radiation pattern of the antenna. An antenna that has a narrow main lobe would have better directivity than the one that has a broad main lobe; hence, this antenna is more directive.

2.2.4 Antenna Efficiency

Antenna efficiency is the measure of the antenna's ability to transmit the input power into radiation. Antenna efficiency is the ratio between the radiated power (P_r) to the input power (P_{in}) [4,5]:

$$e = \frac{p_r}{p_{in}} \tag{2.4}$$

There are a number of antenna efficiencies; the total antenna efficiency e_0 is used to take into account losses at the input terminals and within the structure of the antenna as shown in Figure 2.6. Such losses may be due to [4]:

1. Reflections because of the mismatch between the transmission line and the antenna.

2. I^2R losses (conduction and dielectric).

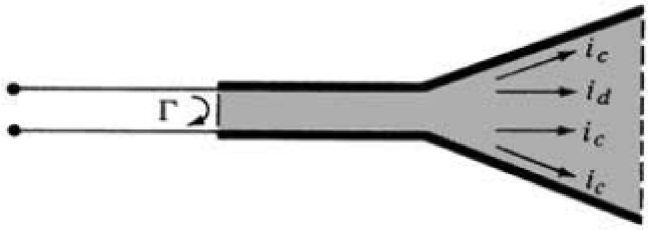

Figure (2.6): antenna losses (Reflection, conduction and dielectric) [4].

In general, the overall efficiency can be written as:

$$e_o = e_r e_c e_d \tag{2.5}$$

where

e_o = total efficiency

e_r = reflection(mismatch) efficiency

e_c = conduction efficiency

e_d = dielectric efficiency

2.2.5 Gain

The gain of an antenna (referred to a lossless isotropic source) depends on both its directivity and its efficiency. If the efficiency is not 100 percent, the gain is less than the directivity [2]. Thus, the gain

$$G = e_0 D \tag{2.6}$$

Gain of an antenna (in a given direction) is defined as "the ratio of the intensity, in a given direction, to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically. The radiation intensity corresponding to the isotropically radiated power is equal to the power accepted (input) by the antenna divided by 4π." [4]. Gain can be expressed as

$$\text{Gain} = 4\pi \frac{\text{radiation intensity}}{\text{total input (accepted) power}} = 4\pi \frac{U(\theta,\phi)}{p_{in}} \tag{2.7}$$

2.2.6 Polarization

The polarization of an antenna is the polarization of the wave radiated in a given direction by the antenna when transmitting. When the direction is not stated, the polarization is taken to be the polarization in the direction of maximum gain. The polarization of a radiated wave is the property of an electromagnetic wave describing the time varying direction and relative magnitude of the electric field vector [4].

If the polarization of the receiving antenna is not the same as the polarization of the incoming (incident) wave, there is polarization mismatch resulting in power loss. The requirement of the antenna polarization depends on the applications [4,5]. Polarization can be categorized as linear, circular and elliptical as shown in Figure 2.7.

• **Linear polarization**: If the electric field vector moves back and forward along a line it is assumed to be linearly polarized. A linearly polarized wave is considered as horizontally polarized if the electric field is parallel to the earth and vertically polarized if the electric field is perpendicular to the earth. For a linearly-polarized antenna, the radiation pattern is taken both for a co-polarized and cross-polarized response.

• **Circular polarization**: If the electric field vector remains constant in length but rotates around in a circular path then it is considered circularly polarized. For circular polarization, the fields components have same magnitude and the phase between two components is *90* degree.

• **Elliptical polarization**: describes an antenna when its electric field vector at a far field point is such that traces elliptical curves constantly with time. Moreover, both the circular and elliptical polarizations are characterized for being right-hand (RH) or left-hand (LH) polarized, depending on the sense of the field. If the field is flowing in the clockwise direction, the field will be right hand polarized; otherwise it will be left hand polarized [4].

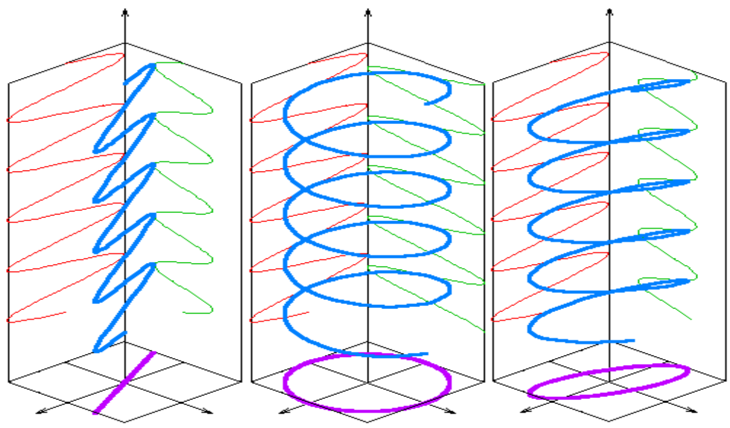

Figure (2.7): Three types of polarization (linear, circular and elliptical) [6].

2.2.7 Input Impedance

Input impedance is defined as “the impedance presented by an antenna at its terminals or the ratio of the voltage to current at a pair of terminals or the ratio of the appropriate components of the electric to magnetic fields at a point.” [4]. The input impedance will be affected by other antennas or objects that are nearby, but we assume that the antenna is isolated. Input impedance is composed of real and imaginary parts [5]:

$$Z_{in} = R_A + jX_A \tag{2.8}$$

The input resistance R_A represents dissipation, which occurs in two ways. Power that leaves the antenna and never returns (i.e., radiation) is a form of dissipation. There are also ohmic losses associated with heating on the antenna structure, but on many antennas ohmic losses are small compared to radiation losses. However, ohmic losses are usually significant on electrically small antennas, which have dimensions much less than a wavelength. The input reactance X_A represents power stored in the near field of the antenna. As a consequence of reciprocity, the impedance of an antenna is identical during reception and transmission [5].

Frequency makes the impedance vary constantly with its value. For lower frequencies (higher wavelengths) the length of the transmission line is not significant when compared to wavelength, so the result is a short line. Although for higher frequencies where the transmission line is slightly a big fraction of a wavelength, it can be a problem because the input impedance will be influenced in a large scale by the length of the transmission line. Therefore, the term impedance matching becomes important, because in this case the length of a transmission line is not significant when compared to the wavelength. The input impedance of an antenna will be matched with a transmission line if both impedances are the same ($Z_A = Z_0$).

If an antenna is mismatched, the loss of power can be very high, because the power generated by the source will be reflected back. As a measure related with power, input impedance is a crucial quantity for the power that an antenna will receive. Thus, as a definition, a maximum power will be delivered from the source to the antenna, if the input impedance is equal to the conjugate of the impedance generated by the source $Z_{in} = Z_S{*}$. Therefore, no power will be transmitted when Z_A is much smaller or much superior than Z_S as shown in Figure 2.8 [4,5].

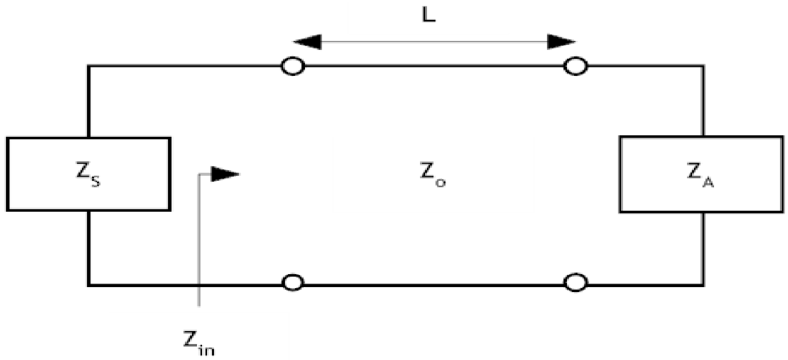

Figure (2.8): Circuit representing input impedance at the entrance terminals of the transmission line[4].

2.2.8 Bandwidth

The term bandwidth specifies the range of frequencies which an antenna can achieve, in order to obtain a desirable behavior of a certain characteristic. The bandwidth can be considered to be the range of frequencies, on either side of a center frequency (usually the resonance frequency for a dipole), where the antenna characteristics such as (input impedance, radiation pattern, beamwidth, polarization, side lobe level, gain, beam direction, radiation efficiency) are within an acceptable value of those at the center frequency (at -10 dB). It is classified as the first required condition before building an antenna, because it is a measure of how acceptable the performance of an antenna can be. However, as a range, two boundaries define the lower and upper frequency limits, and the ratio of its size to the centre frequency as a percentage define the percent bandwidth for a narrowband antenna – thus occupying a small space quantity on the RF spectrum – given by equation (2.9); otherwise, for a broadband (or wideband) antenna the bandwidth is defined as the ratio of the upper to lower frequencies as written in equation (2.10). Both expressions are analytically represented as [4,7]:

$$B_f = \frac{f_H - f_L}{f_c} \times 100 \tag{2.9}$$

$$B_r = \frac{f_H}{f_L} \tag{2.10}$$

where:

B_f – Fractional bandwidth in Hz percentage;

B_r – Bandwidth ratio;

f_H – Upper frequency in Hz;

f_L – Lower frequency in Hz;

f_c – Centre frequency in Hz.

In fact, all the parameters can be influenced by frequency in a different way, what gives a different meaning to the bandwidth of each. Especially analyzing the radiation pattern and input impedance bandwidth differences, the useful bandwidth of a designed antenna can be related to both, despite of the disparity on those differences. In some cases the satisfactory bandwidth for radiation pattern goes above the one for input impedance, or vice-versa [4,7].

2.3 Microstrip or Patch Antenna

Microstrip "patch" antenna is an antenna that made from patches of conducting material on a dielectric substrate above a ground plane [6], it is shown in Figure 2.9. Microstrip antennas received considerable attention starting in the 1970s[1].

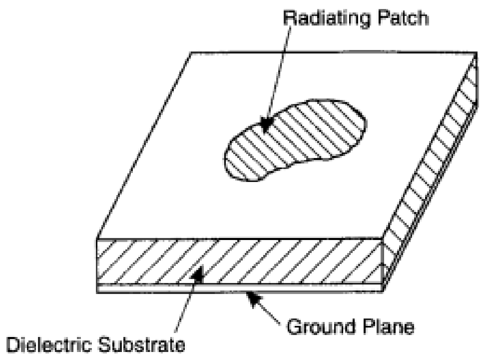

Figure (2.9): Microstrip antenna configuration [7].

This patch can take one of several forms to achieve different design requirements. Typical shapes are rectangular, square, circular and circular ring [8], but it can take any other shape as shown in Figure 2.10.

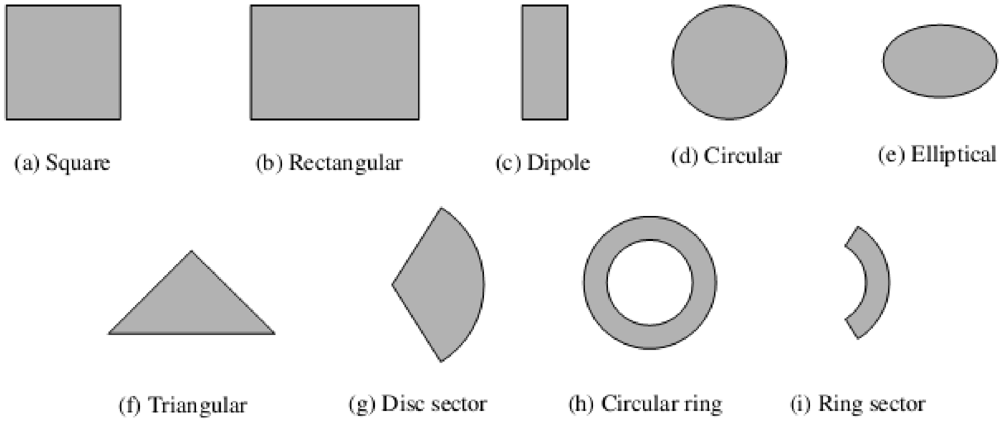

Figure (2.10): Representative shapes of microstrip patch elements [1].

2.3.1 Microstrip Antenna Advantages

Microstrip antenna has many advantages compared to other microwave antennas types. So many applications cover the frequency range from 100 MHz to 100 GHz. These advantages are [6,7]:

1. Low weight, low size, and thin profile configuration.
2. Ease to design and fabricate.
3. Low fabrication cost.
4. With simple feed, linear and circular polarization can be made.
5. Dual frequency and dual polarization can be achieved easily.
6. It is easy to be integrated with microwave integrated circuits.
7. Feed lines and matching networks can be fabricated simultaneously with antenna structure.
8. No cavity backing are required.
9. Easy to be mounted on surfaces when aerodynamic profile is constraint, as on the surface of aircraft.

2.3.2 Disadvantages and Limitations of Microstrip Antenna

Also microstrip antennas have many limitations and disadvantages compared to other antenna types. These disadvantages are[1,6,7,8]:

1. Low efficiency that comes from:
    a. Losses from conducting dielectric and surface waves
    b. Extraneous radiation from feeds and junctions.
    c. Excitation of surface waves
    d. Large ohmic loss in the feed structure arrays.

2. Low power handling capability which means that it is not suitable for high power applications more than 100 watts.

3. Poor polarization purity

4. Narrow frequency bandwidth that just only a fractional of a percent or at most a few percent.

5. Somewhat lower gain (~6 dB).

6. Complex feed structure for high performance array.

2.3.3 Basic Characteristics

As shown in Figure 2.11 a microstrip antenna consists of a very thin conducting patch ($t \ll \lambda_0$ where λ_0 is the free space wavelength) of any planar or non planar geometry on one side of a small fraction of wave length dielectric ($h \ll \lambda_0$ usually $0.003\ \lambda_0 < h < 0.05\ \lambda_0$) with a conducting ground plane on the other side. The substrate relative permittivity ε_r is normally between 2-24[8], some of designers let ε_r between 2.2 - 12 [1].

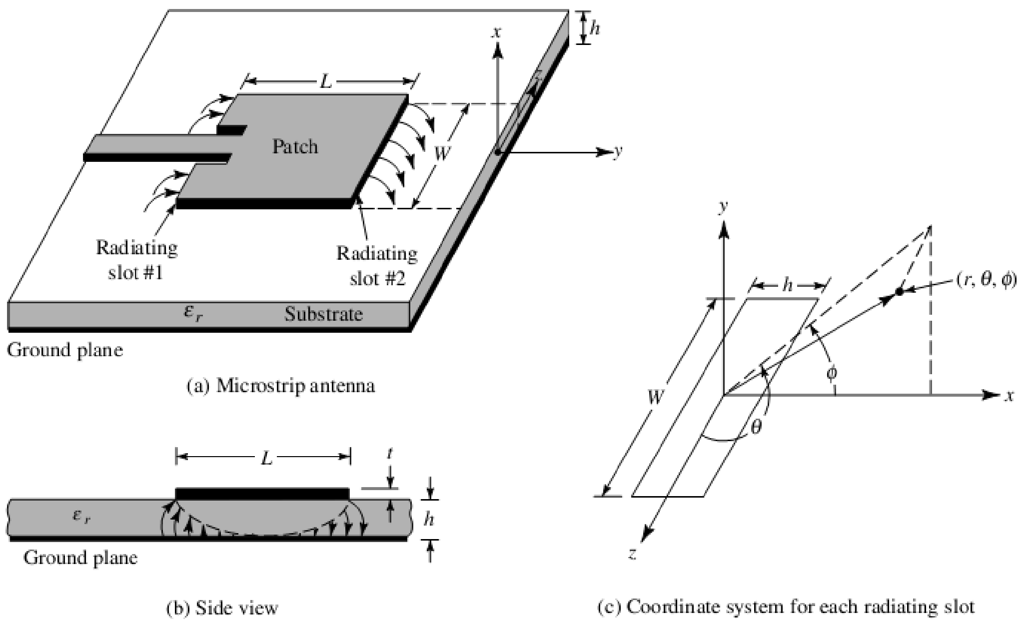

Figure (2.11): Microstrip antenna and its dimensions [1].

Radiation characteristics have been calculated for a large number of patch antennas. Their radiation characteristics are similar despite of the differences in geometrical shape, because they behave like a dipole [7]. A patch antenna has a gain between 5-6 dB and exhibit 3dB beamwidth between 70º and 90º [7].

The microstrip line structure is similar to microstrip antenna. The radiation from microstrip line is sometimes undesired. The microstrip line radiation can be reduced if the substrate is thin and has higher relative dielectric constant, so it can be used for microstrip lines[7] and microwave circuitry because they require tightly bound fields to minimize undesired radiation and coupling, and lead to smaller element sizes; however, because of their greater losses, they are less efficient and have relatively smaller bandwidths[1]. On the other hand thick substrates with low permittivity is used for microstrip antennas because the radiation is the target of the antenna[7].

The dimensions of rectangular microstrip antenna are shown in Figure 2.11 where patch length *L* is between $\lambda_0/3$ and $\lambda_0/2$ and its width *W* is smaller than λ_0 but it can not be too small otherwise the antenna becomes a microstrip line which is not a radiator [8].

When rectangular antenna is connected to a microwave source, the excitation of the patch will make a charge distribution on the upper and lower surfaces of the patch as shown in Figure 2.12. It can be modeled as a transmission line resonant cavity with two open ends where the fringing fields from the patch to the ground are exposed to the upper half space and are responsible for the radiation. This radiation mechanism is the same as the slot line, thus there are two radiating slots on a patch antenna, as indicated in Figure 2.13(a)[8].

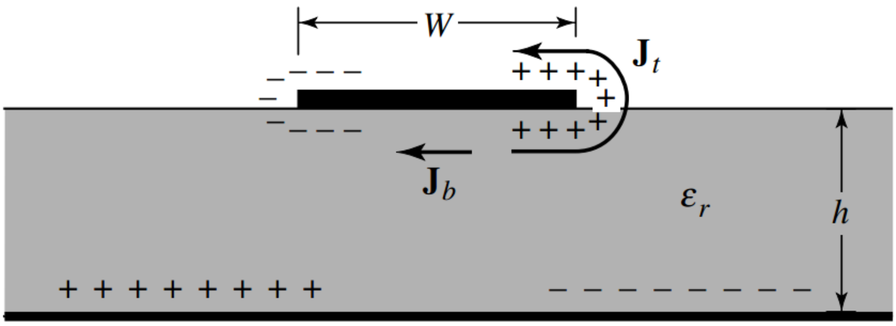

Figure (2.12): charge distribution and current density on a microstrip antenna[1].

2.3.4 Feeding methods

There are different ways to feed the microstrip antenna. The most popular ways are four configurations. They can be listed as coaxial probe, microstrip line, aperture coupling and proximity coupling [1], as shown in Figure 2.13. Their equivalent circuits are shown in Figure 2.14.

In this work, the focus will be on microstrip line feed.

2.3.4 Ground Plane

The ground plane is usually the conducting patch on the other side of the substrate which is ideally should be infinite as for monopole antenna. But practically a small ground plane is needed as shown in Figure 2.13. Fringing fields between the patch and the ground plane generate the radiation of the microstrip antenna[8].

2.3.5 Analyzing Rectangular Microstrip Antenna

Microstrip antenna can be analyzed and modeled in several ways. It can be analyzed as a transmission line model, a cavity model, wire grid model [7] and full wave model which includes primarily integral equations and moment methods.

The transmission line model is the easiest of all methods and it gives a good physical insight, but it has two disadvantages, that it is less accurate and it is difficult to model coupling by it.

The cavity model is more complex, but it is more accurate than transmission line model. It also gives a good physical insight and it is less difficult to model coupling, so it can be successfully used.

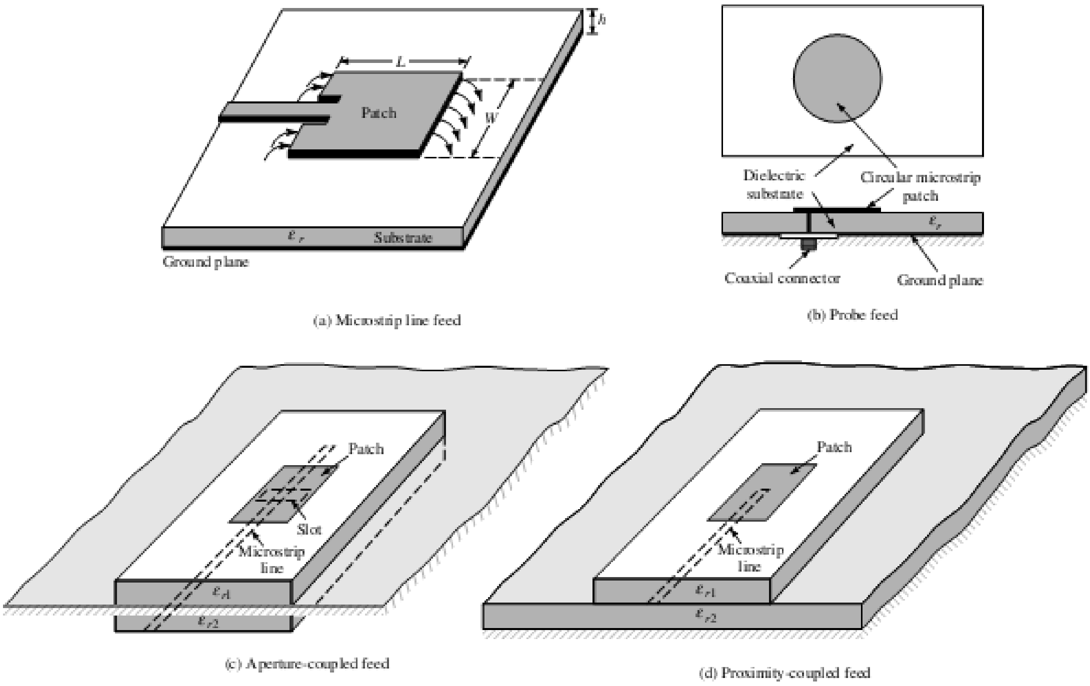

Figure (2.13): Microstrip antenna and its most popular feeding types[1].

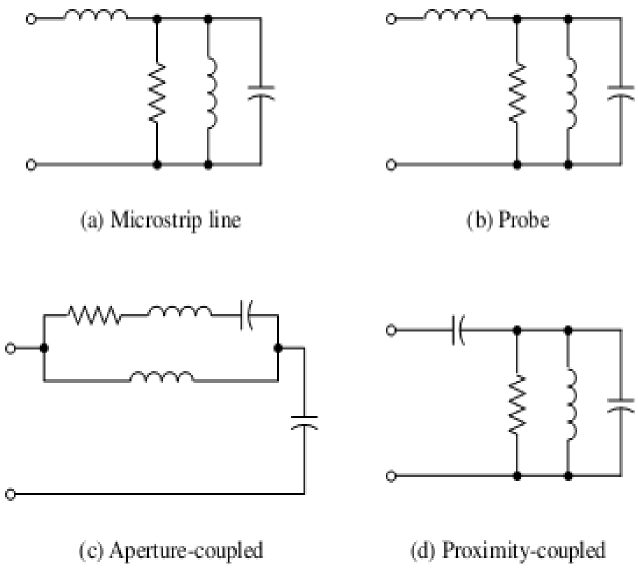

Figure (2.14): Equivalent circuits for typical feed of Figure 2.13[1].

The full wave model is the most complex model and usually gives less physical insight, but it is very accurate, very multilateral, and can treat single elements, finite and infinite arrays, stacked elements, arbitrary shaped elements and coupling [1].

For a rectangular microstrip antenna, the transmission line model represents the patch by two slots that are separated by a low impedance transmission line of length *L*. the cavity model represents it as an array of two radiating narrow apertures each of width *W* and height *h* that are separated by a distance *L*.

2.3.6 Designing Rectangular Microstrip Antenna

To design a rectangular microstrip antenna, it is necessary to choose the substrate material with thickness *h*, dielectric permittivity "constant" ε_r and the target center "resonant" frequency f_r in Hz, then determine the antenna dimensions: antenna width *W* and length *L*.

This can be done by the following steps:

step 1: For efficient radiator, calculate the width *W* from the equation (2.11) [1,8]:

$$W = \frac{1}{2 f_r \sqrt{\varepsilon_o \mu_o}} \sqrt{\frac{2}{\varepsilon_r + 1}} = \frac{c}{2 f_r} \sqrt{\frac{2}{\varepsilon_r + 1}} \tag{2.11}$$

where *c* is the free space light velocity.

step 2: Determine the effective dielectric constant ε_{eff} from the equation (2.12)[8]:

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2\sqrt{1 + 12\frac{h}{W}}} \tag{2.12}$$

step 3: Determine the incremental length ΔL produced by the fringing fields from equation (2.13)[8]:

$$\Delta L = 0.412h \frac{(\varepsilon_{eff} + 0.3)(\frac{W}{h} + 0.264)}{(\varepsilon_{eff} + 0.258)(\frac{W}{h} + 0.8)} \tag{2.13}$$

step 4: Determine the effective length L_{eff} of the patch by equation (2.14) then determine the actual length *L* of the patch by equation (2.15)[7]:

$$L_{eff} = \frac{c}{2 f_r \sqrt{\varepsilon_{eff}}} \tag{2.14}$$

$$L = L_{eff} - 2\Delta L \tag{2.15}$$

After that the antenna can be simulated using any simulation tool and its parameters can be optimized to obtain optimum characteristics.

2.4 Summary

This chapter discussed antenna theory and parameters then described the microstrip antenna by discussing their advantages and disadvantages, basic character, feeding methods, analyzing them and finally their designing procedure.

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Chapter 3 Microwave Bandpass Filters

This chapter will discuss resonators, filter theory, filter types and filter design procedure.

3.1 Introduction

Filters play very important role in many radio frequency microwave applications. They are commonly used in transceiver circuits to separate or combine different frequencies. The electromagnetic spectrum is limited and has been shared. Filters are used to select or confine the radio frequency microwave signals within assigned spectral limits [1].

Filters simply can be defined as two port networks that allow some frequencies to go through while block the others.

Microwave bandpass filters basically consist of a number of resonators between two ports, this number can identify the filter's order. These resonators can be coupled electrically or magnetically or both. Microwave resonators are tunable circuits used in microwave oscillators, filters and frequency meters. Their operation is very similar to that of lumped element resonators (such as parallel and series RLC circuits) of circuit theory.

3.2 Microwave Resonators

This section will talk about the microwave resonators because they are the basic components in bandpass filters. A microwave resonator is a tunable circuit that is used in microwave oscillators, filters and frequency meters. Resonators can be made of closed sections of waveguides (cavity) or transmission lines such as microstrip resonators. Cavity resonator is defined in English dictionary as " A hollow chamber or cavity with dimensions chosen to permit internal resonant oscillation of electromagnetic or acoustical waves of specific frequencies."[4]. A microstrip resonator is defined as" A microstrip resonator is any structure that is able to contain at least one oscillating electromagnetic field."[1]. These definitions let that the resonator has the effect to identify the filter's frequencies.

3.2.1 RLC Resonant Circuits

Resonance is the tendency of a system to oscillate at maximum amplitude, so at the resonant (tuned) frequency the average energies stored are equal. This means that the energy stored in electric field (or in a capacitor) W_E will equal the energy stored in magnetic field (or in a inductor) W_M , then the impedance becomes purely real.

The operation of microwave resonators is very similar to that of the lumped element resonators of circuit theory, near resonance, a microwave resonator can usually be modeled by either a series or parallel RLC lumped-element equivalent circuit.

First in the series RLC resonant circuits that is shown in Figure 3.1, the input impedance is

$$Z_{in} = R + j\omega L - j\frac{1}{\omega C} \tag{3.1}$$

where R is the resistance (real impedance)

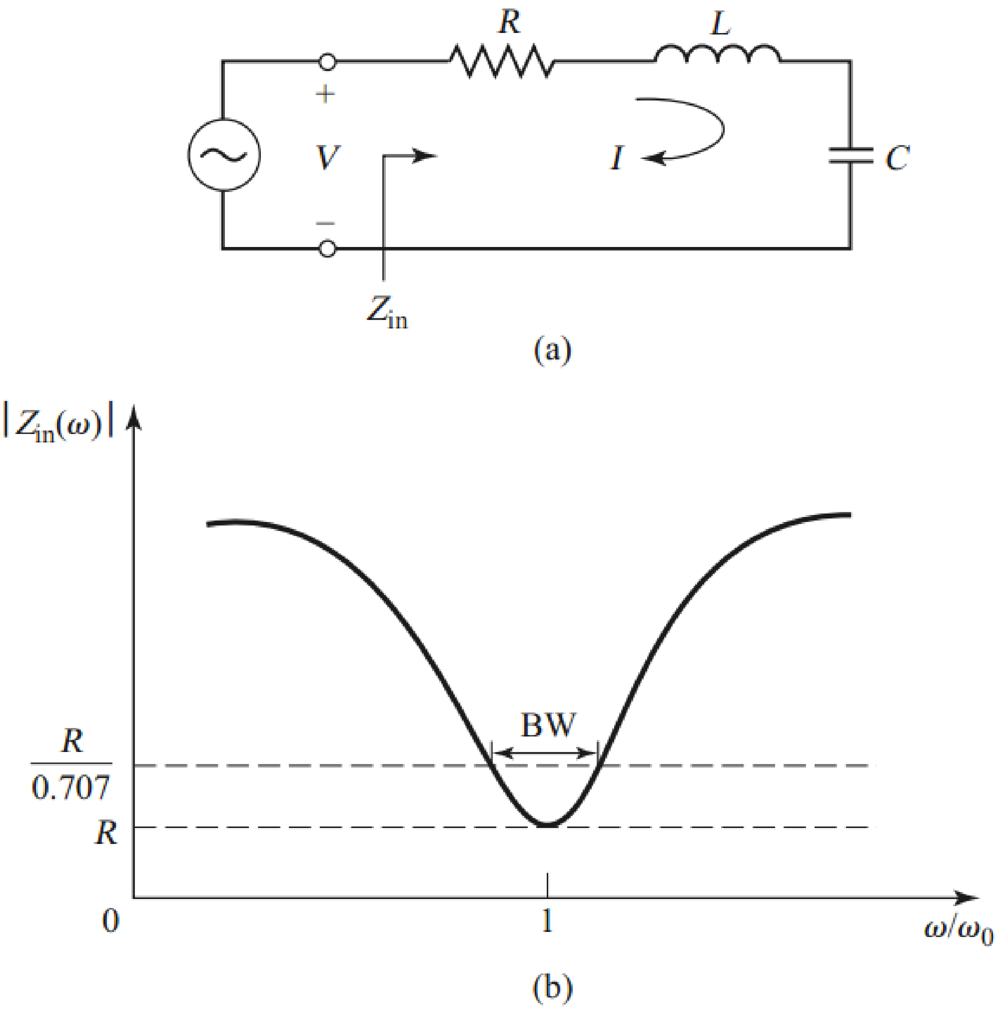


Figure (3.1): A series RLC circuit and its response (a) The series RLC circuit (b) the input impedance magnitude versus frequency [5].

At resonance the average magnetic energy W_M stored in the inductor L will equal the average electric energy W_E stored in the capacitor C.

$$W_M = W_E \tag{3.2}$$

where

$$W_M = \frac{1}{4}|I|^2 L \tag{3.3}$$

$$W_E = \frac{1}{4}|I|^2 \frac{1}{\omega^2 C} \tag{3.4}$$

This implies

$$\omega_0 = \frac{1}{\sqrt{LC}} \tag{3.5}$$

and

$$Z_{in} = R \tag{3.6}$$

So at the resonance the impedance will be pure resistance

Quality factor Q (see section 3.3.2) is a measurement of the loss of a resonant circuit that is generally defined as

$$Q = \omega \frac{W_E + W_M}{P_{loss}} \tag{3.7}$$

where P_{loss} is the power dissipated by the resistor R that is equal to

$$P_{loss} = \frac{1}{2}|I|^2 R \tag{3.8}$$

This implies that lower loss implies higher quality factor. Then the quality factor at resonance will be

$$Q = \omega_0 \frac{2W_M}{P_{loss}} = \frac{\omega_0 L}{R} = \omega_0 \frac{2W_E}{P_{loss}} = \frac{1}{\omega_0 CR} \tag{3.9}$$

but at the near resonance $\omega = \omega + \Delta\omega$ where $\Delta\omega$ is small. the input impedance will be

$$Z_{in} = R + j\omega L - j\frac{1}{\omega C} = R + j\omega L\left(1 - \frac{1}{\omega^2 LC}\right)$$

$$Z_{in} = R + j\omega L\left(\frac{\omega^2 - \omega_0^2}{\omega^2}\right) \approx R + j2L\Delta\omega \approx R + j\frac{2RQ\Delta\omega}{\omega_0} \tag{3.10}$$

Since

$$\omega^2 - \omega_0^2 = (\omega - \omega_0)(\omega + \omega_0) = \Delta\omega(2\omega - \Delta\omega) \approx 2\omega\Delta\omega \tag{3.11}$$

By the same way, the parallel resonant RLC circuit that is shown in Figure 3.2, can be analyzed as shown next.

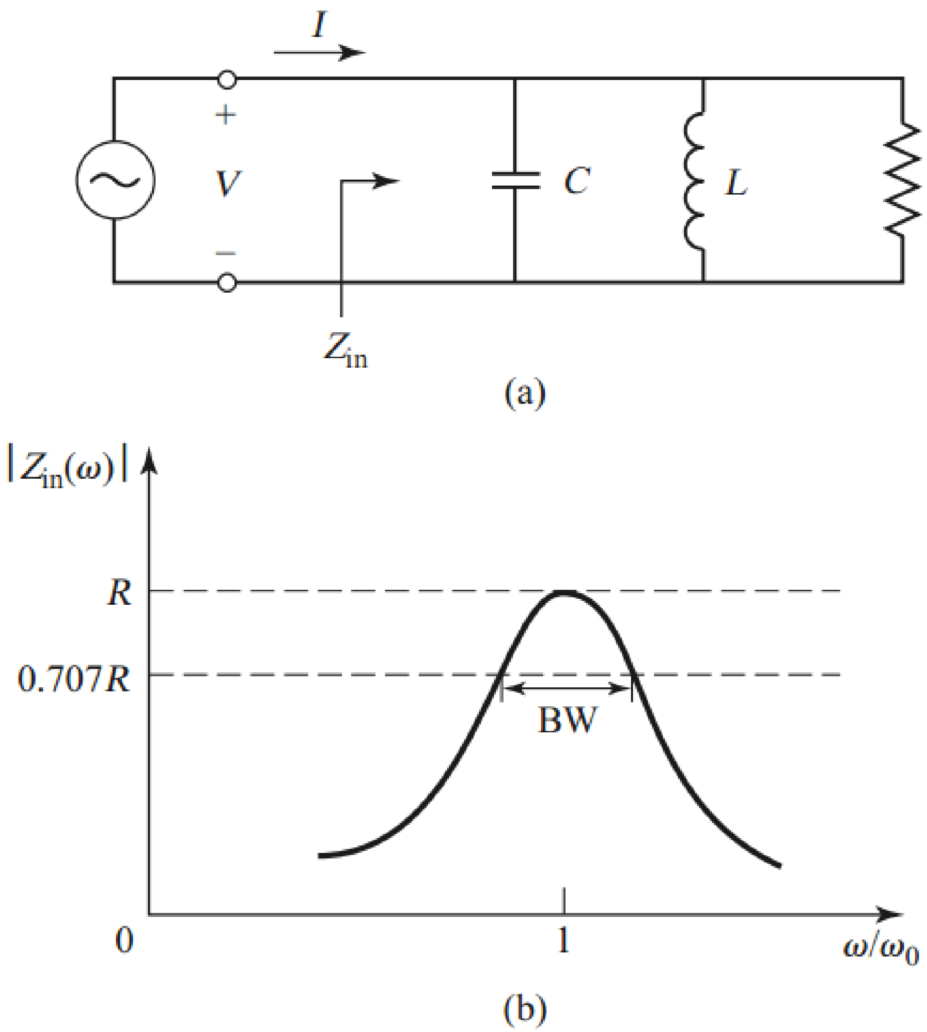


Figure (3.2): A parallel RLC circuit and its response (a) The parallel RLC circuit (b) the input impedance magnitude versus frequency [5].

For parallel resonant RLC circuit, the input impedance Z_{in} is

$$Z_{in} = \left(\frac{1}{R} + j\omega C - j\frac{1}{\omega C} \right)^{-1} \tag{3.12}$$

and the power loss is

$$P_{loss} = \frac{1}{2}\frac{|V|^2}{R} \tag{3.13}$$

The energy stored in inductor and capacitor are

$$W_M = \frac{1}{4}\frac{|V|^2}{\omega^2 L} \tag{3.14}$$

$$W_E = \frac{1}{4}|V|^2 C \tag{3.15}$$

the resonance occurs when $W_E=W_M$ then

$$\omega_0 = \frac{1}{\sqrt{LC}}$$

$$Z_{in}=R$$

So at the resonance

$$Q = \omega_0 \frac{2W_M}{P_{loss}} = \frac{R}{\omega_0 L} = \omega_0 \frac{2W_E}{P_{loss}} = \omega_0 CR \tag{3.16}$$

but at the near resonance $\omega = \omega + \Delta\omega$ where $\Delta\omega$ is small. the input impedance will be

$$Z_{in} = \left(\frac{1}{R} + j\omega C - j\frac{1}{\omega L}\right)^{-1} = \left(\frac{1}{R} + j\omega C\left(1 - \frac{1}{\omega^2 LC}\right)\right)^{-1}$$

$$Z_{in} = \left(\frac{1}{R} + j\omega C\left(\frac{\omega^2 - \omega_0^2}{\omega^2}\right)\right)^{-1} \approx \left(\frac{1}{R} + j2C\Delta\omega\right)^{-1}$$

$$Z_{in} \approx \frac{R}{1 + j2CR\Delta\omega} \approx \frac{R}{1 + j2Q\frac{\Delta\omega}{\omega_0}} \tag{3.17}$$

since from equation 3.11 $\omega^2 - \omega_0^2 \approx 2\omega\Delta\omega$

3.2.2 Loaded and Unloaded Quality Factor

Quality factor Q is the characteristic of the resonant circuit itself with no loading effects. It is a measurement of the loss of a resonant circuits. It can be calculated by equations (3.9) or (3.16). If the circuit is effected by a load from external circuitry then the external circuitry effect is defined as external quality factor Q_e. So quality factor Q is defined as unloaded quality factor [5]. The loaded quality factor Q_L takes the account for characteristic of the resonant circuit and these loading effects because in practice a resonant circuit is invariably coupled to other circuitry, so Q_L is lower than Q. This can be modeled as shown in Figure 3.3 where the external circuitry is modeled by external load resistor R_L. In the model if the resonant RLC circuit is series, then the effective resistance is $R+R_L$, but if the circuit is parallel then the effective resistance is $R||R_L = RR_L / (R+R_L)$ [5]. This results that Q, Q_e and Q_L can be expressed as

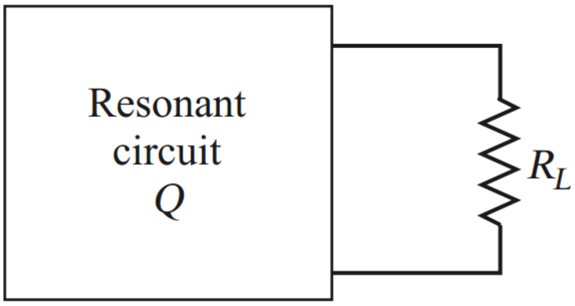


Figure (3.3): A resonant circuit connected to an external load RL [5].

$$Unloaded \quad Q = \begin{cases} \dfrac{\omega_0 L}{R} & series \quad circuit \\ \\ \dfrac{R}{\omega_0 L} & parallel \quad circuit \end{cases}$$

$$External \quad Q_e = \begin{cases} \dfrac{\omega_0 L}{R_L} & series \quad circuit \\ \\ \dfrac{R_L}{\omega_0 L} & parallel \quad circuit \end{cases} \qquad (3.18)$$

$$Loaded \quad Q_L = \begin{cases} \dfrac{\omega_0 L}{R + R_L} & series \quad circuit \\ \\ \dfrac{R \parallel R_L}{\omega_0 L} & parallel \quad circuit \end{cases} \qquad (3.19)$$

$$\frac{1}{Q_L} = \frac{1}{Q} + \frac{1}{Q_e} \qquad (3.20)$$

Table 3.1 summarizes the above results for series and parallel resonant circuits.

3.2.3 Resonator Types

Resonators can be classified to three major types:

a. Transmission line resonators.

b. Cavity resonators.

c. Dielectric resonators.

3.2.3.1 Transmission line resonators

Transmission line usually consists of two parallel conductors as shown in Figure 3.4(a). it can be coaxial cable, two wire transmission line, parallel plate wave guide, microstrip line and coplanar wave guide [5].

Table (3-1): Summary of Results for Series and Parallel Resonator[5].

Quantity	Series Resonator	Parallel Resonator
Input impedance/admittance	$Z_{in} = R + j\omega L - j\frac{1}{\omega C}$	$Y_{in} = \frac{1}{R} + j\omega C - j\frac{1}{\omega L}$
	$\simeq R + j\frac{2RQ_0\Delta\omega}{\omega_0}$	$\simeq \frac{1}{R} + j\frac{2Q_0\Delta\omega}{R\omega_0}$
Power loss	$P_{loss} = \frac{1}{2}\|I\|^2 R$	$P_{loss} = \frac{1}{2}\frac{\|V\|^2}{R}$
Stored magnetic energy	$W_m = \frac{1}{4}\|I\|^2 L$	$W_m = \frac{1}{4}\|V\|^2\frac{1}{\omega^2 L}$
Stored electric energy	$W_e = \frac{1}{4}\|I\|^2\frac{1}{\omega^2 C}$	$W_e = \frac{1}{4}\|V\|^2 C$
Resonant frequency	$\omega_0 = \frac{1}{\sqrt{LC}}$	$\omega_0 = \frac{1}{\sqrt{LC}}$
Unloaded Q	$Q_0 = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 RC}$	$Q_0 = \omega_0 RC = \frac{R}{\omega_0 L}$
External Q	$Q_e = \frac{\omega_0 L}{R_L}$	$Q_e = \frac{R_L}{\omega_0 L}$

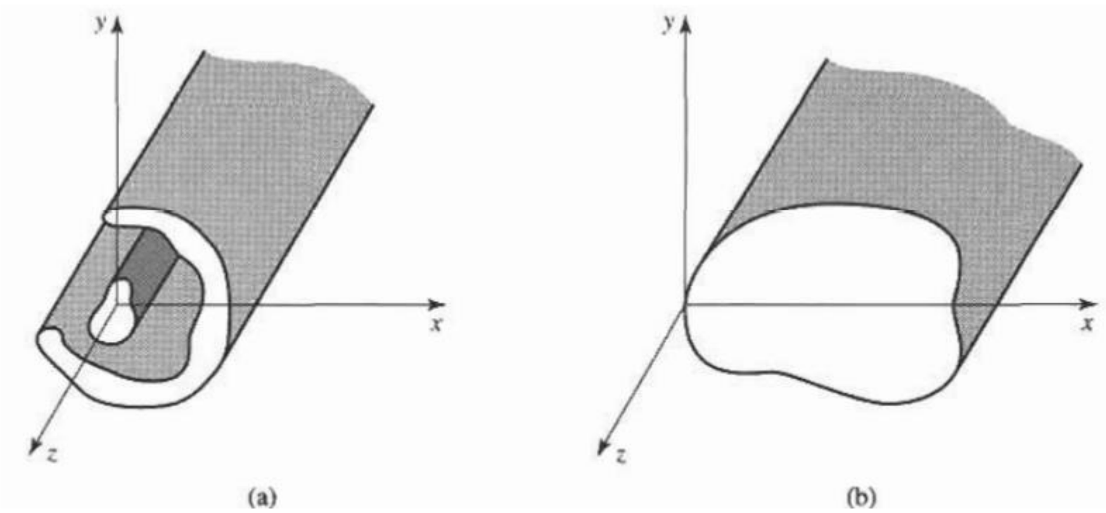


Figure (3.4): (a) General two-conductor transmission line (b) closed waveguide [5].

Transmission line sections can be used in various lengths and termination (usaually open or short circuited) to form resonators. At near resonance these sections can be modeled by either a series or parallel RLC lumped element equivalent circuits [5]. Figure 3.5 shows equivalent lumped element model for some transmission line resonators.

3.2.3.2 Cavity Resonators

Resonators can also be constructed from closed sections of waveguide as in Figure 3.4(b). Waveguide resonators are usually short circuited at both ends because of radiation loss from open ended waveguide. These shorted ends (closed ends) are forming a closed box or cavity, and electric and magnetic energy is stored within the cavity. Coupling to the resonator can be done by a small aperture or a small probe or loop. Power can be dissipated in the metallic walls of the cavity as well as in the dielectric filling the cavity [5]. Cavity resonators can be either rectangular or circular waveguide cavity as shown in Figure 3.6.

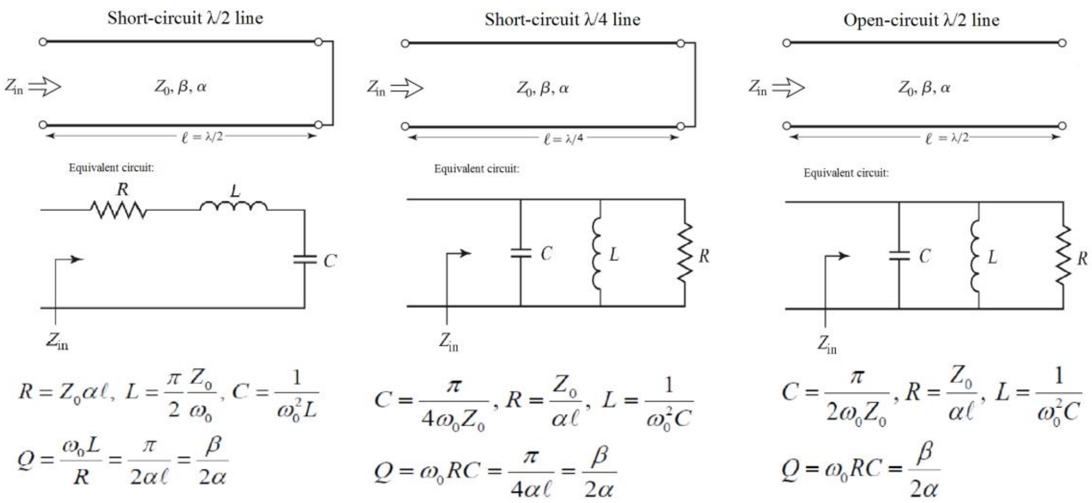


Figure (3.5): Resonators' types and their modeling.

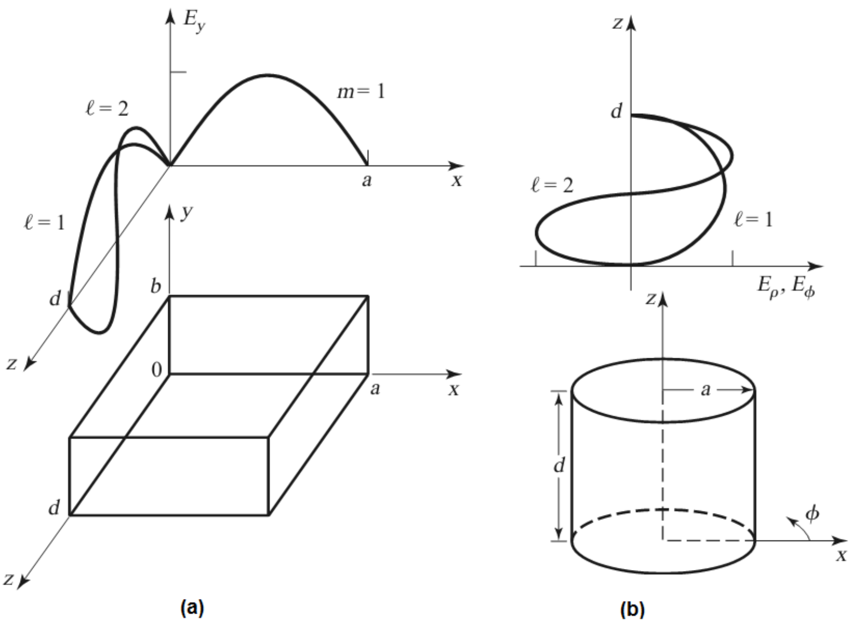


Figure (3.6): (a) A rectangular resonant cavity, and the electric field distributions for the TE_{101} and TE_{102} resonant modes. (b) A cylindrical resonant cavity, and the electric field distribution for resonant modes with l=1 or l=2 [5].

3.2.3.3 Dielectric Resonators

The microwave resonator also can be constructed from a small disc or cube of low-loss high dielectric constant material [5]. Dielectric resonators are similar in principle to the rectangular or cylindrical waveguide cavities. The high dielectric constant of the resonator ensure that most of the fields are contained within the dielectric, but unlike metallic cavities, there is some field fringing or leakage from the sides and ends of the dielectric resonator. Dielectric resonator has many advantages that can be listed as [5]:

1. Smaller in cost, size and weight than an equivalent metallic cavity.
2. It can very easily be incorporated into microwave integrated circuits.
3. It can very easily be coupled to planar transmission lines.
4. Materials with dielectric constants $10 \leq \varepsilon_r \leq 100$ are generally used.
5. Conductor losses are absent.
6. Q can be achieved up to several thousands.
7. The resonant frequency can be mechanically tuned by using adjustable metal plate above the resonator.

3.2.4 Microstrip Resonators

A microstrip resonator is any structure that is able to contain at least one oscillating electromagnetic field[1]. This definition means that there are many forms of microstrip resonators. Microstrip resonators for filter designs can generally classified as lumped-element or quasilumped-element resonators and distributed line or patch resonators. Figure 3.7 shows some typical configurations of microstrip resonators.

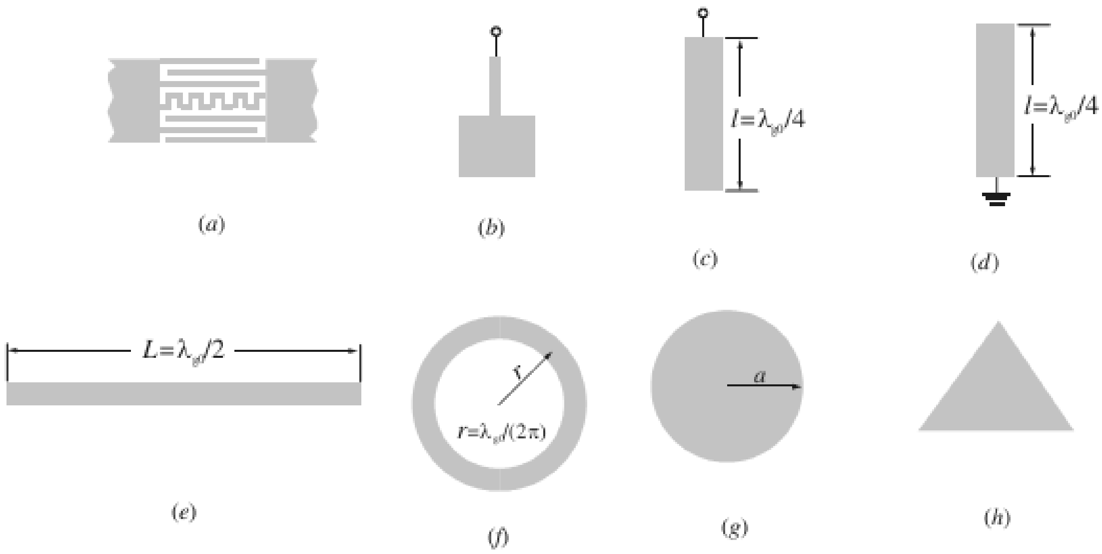


Figure (3.7): Some typical microstrip resonators: (a) lumped-element resonator; (b) quasilumped-element resonator; (c) $\lambda_{g0}/4$ line resonator (shunt series resonance); (d) $\lambda_{g0}/4$ line resonator (shunt paral-lel resonance); (e) $\lambda_{g0}/2$ line resonator; (f) ring resonator; (g) circular patch resonator; (h) triangular patch resonator.[1]

Lumped-element or quasilumped-element resonators are shown in Figure 3.7(a) and (b), they are formed by the lumped or quasilumped inductors and capacitors. They will resonate at $\omega_0 = 1/\sqrt{LC}$.

The distributed line resonators as shown in Figure 3.7(c) and (d) may be termed quarter-wavelength resonators, since they are $\lambda_{go}/4$ long, where λ_{go} is the guided wavelength at the fundamental resonant frequency f_0. They can also resonate at other higher frequencies when $f \approx (2n\text{-}1)\ f_0$ for $n=2,3,...$. Another typical distributed line resonator is the half-wavelength resonator, as shown in Figure 3.7(e), which is $\lambda_{go}/2$ long at its fundamental resonant frequency, and can also resonate at $f \approx n\ f_0$ for $n=2,3,...$.

The ring resonator as shown in Figure 3.7(f) is another type of distributed line resonator, where r is the median radius of the ring. The ring will resonate at its fundamental frequency f_0 when its median circumference $2\ \pi r \approx \lambda_{\ g0}$. The higher resonant modes occur at $f \approx nf_0$ for $n=2,3,...$. But in this configuration, the symmetrical geometry applies the resonator to resonate in either of two orthogonal coordinates that will give this type of resonator an important feature, that it can support a pair of degenerate modes that have the same resonant frequencies but orthogonal field distributions. This feature can be utilized to design dual-mode filters. The ring resonator also can be square or meander loops.

Patch resonators are shown in Figure 3.7(g) and (h). These resonators are important to increase the power handling capability for microstrip filters design. The main advantage of microstrip patch resonators is their lower conductor losses as compared with narrow microstrip line resonators. For filter applications, patch resonators are usually enclosed in metal housings to minimize the radiation loss because they have stronger radiation than other resonator types. Patch resonators usually have a larger size, but this would not be a problem for the applications in which the power handling or low loss have a higher priority. Also the size may not be a problem for the filters operating at very high frequencies., Patches may take different shapes, such as circular, triangular, rectangular, . . .etc. depending on the applications. These microstrip patch resonators can be analyzed as waveguide cavities with magnetic walls on the sides.

3.3 Chebyshev Filters

This section will discuss chebyshev filters, chebyshev response, and chebyshev transfer function. Then it will discuss designing lowpass prototype filters then filter transformation to chebyshev bandpass filter.

3.3.1 Chebyshev Filters

Chepyshev filters are used to separate one band of frequencies from another[6]. Their response can be defined as "The Chebyshev response is a mathematical strategy for achieving a faster roll-off by allowing ripple in the frequency response[6]". Cheyshev filters can be analog or digital. They can be one of two types as shown in Figure 1.10 (b) and (c):

Type I: Chebyshev filters that have ripples in the passband and no ripples in the stopband.

Type II: Inverse Chebyshev filters that have ripples in the stopband and no ripples in the passband.

3.3.2 Transfer Function

The amplitude-squared transfer function that describes the response of type I Chebyshev filters is [1,5]

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\varepsilon^2 T_n^2(\Omega)} \tag{3.21}$$

where the ripple constant ε is related to a given passband ripple L_{Ar} in dB by

$$\varepsilon = \sqrt{10^{\frac{L_{Ar}}{10}} - 1} \tag{3.22}$$

$T_n(\Omega)$ is a Chebyshev function of type I of order n, which is defined as

$$T_n(\Omega) = \begin{cases} \cos(n\cos^{-1}\Omega) \\ \cosh(n\cosh^{-1}(\Omega) \end{cases} \tag{3.23}$$

3.3.3 Lowpass Prototype Filters

A lowpass prototype filter is in general defined as the lowpass filter whose element values are normalized to make the source resistance or conductance equal to one, denoted by g_0=1, and the cutoff angular frequency to be unity, denoted by Ω_c=1 (rad/s). There are two possible forms of an n-pole lowpass prototype for realizing an all-pole filter response, including Butterworth, Chebyshev, and Gaussian responses. Lowpass prototype filters are demonstrated in Figure 3.8. They both are dual from each other and give the same response, so either form may be used [1].

It should be noted that in Figure 3.8, g_i for i = 1 to n represent either the inductance of a series inductor or the capacitance of a shunt capacitor; therefore, n is also the number of reactive elements. If g_1 is the shunt capacitance or the series inductance, then g_0 is defined as the source resistance or the source conductance. Similarly, if g_n is the shunt capacitance or the series inductance, g_{n+1} becomes the load resistance or the load conductance. Unless otherwise specified these g-values are supposed to be the inductance in henries, capacitance in farads, resistance in ohms, and conductance in mhos.

This type of lowpass filter can serve as a prototype for designing many practical filters with frequency and element transformations. For Chebyshev lowpass prototype filter with a passband ripple L_{Ar} dB and the cutoff frequency $\Omega_c = 1$, the element values may be computed using the following formulas [1]:

$$g_0 = 1 \tag{3.24}$$

$$g_1 = \frac{2}{\gamma}\sin\left(\frac{\pi}{2N}\right) \tag{3.25}$$

$$g_i = \frac{1}{g_{i-1}} \frac{4\sin\left[\frac{(2i-1)\pi}{2N}\right]\cdot\sin\left[\frac{(2i-3)\pi}{2N}\right]}{\gamma^2 + \sin^2\left[\frac{(i-1)\pi}{N}\right]} \qquad \text{for} \qquad i = 2,3,\cdots N \tag{3.26}$$

$$g_{N+1} = \begin{cases} 1 & for \quad N \quad odd \\ \coth^2\left(\frac{\beta}{4}\right) & for \quad N \quad even \end{cases} \tag{3.27}$$

$$where \qquad \beta = \ln\left[\coth\left(\frac{L_{Ar}}{17.37}\right)\right] \tag{3.28}$$

$$\gamma = \sinh\left(\frac{\beta}{2N}\right) \tag{3.29}$$

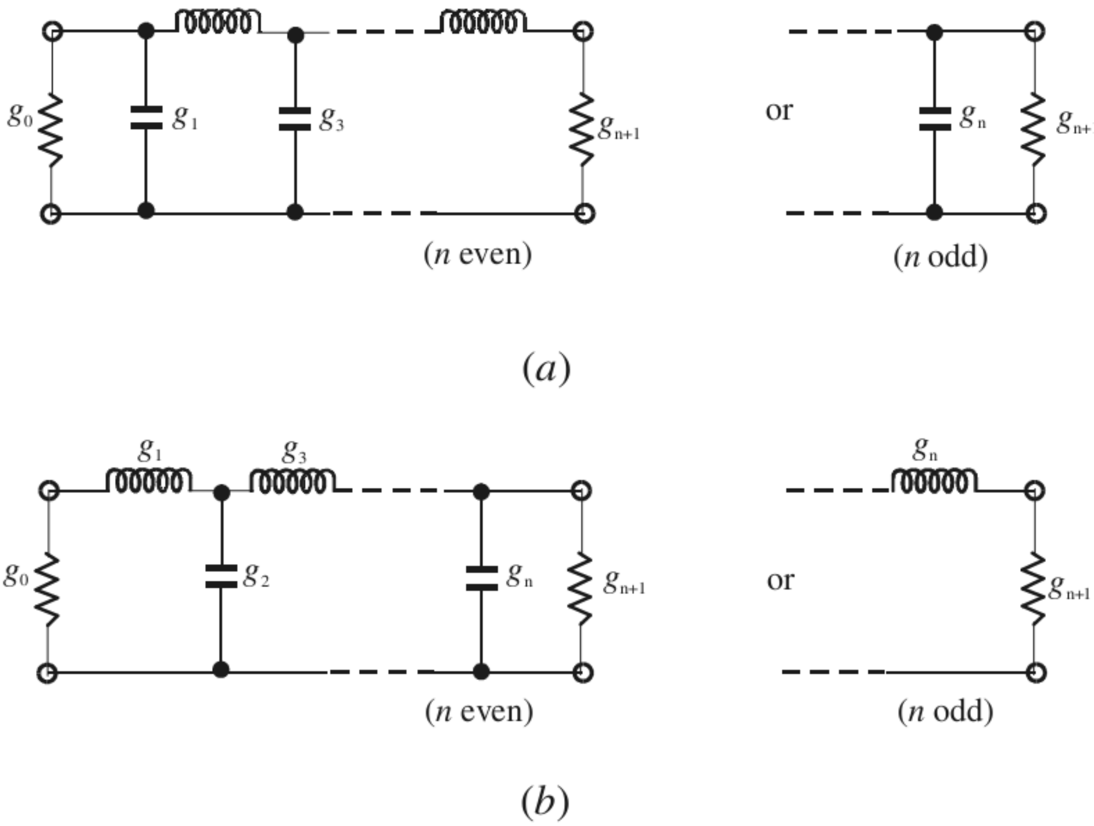


Figure (3.8): Lowpass prototype filters for all-pole filters with (a) a ladder network structure and (b) its dual.[1]

Some typical element values for such filters are tabulated in Table 3.2 [1] for various passband ripples L_{Ar} , and for the filter degree of n = 1 to 9. For a given passband ripple L_{Ar} dB, minimum stopband attenuation L_{As} dB at $\Omega=\Omega_s$, the degree of a Chebyshev lowpass prototype, which will meet this specification, can be found by [1]

$$n \geq \frac{\cosh^{-1}\sqrt{\dfrac{10^{0.1L_{As}}-1}{10^{0.1L_{Ar}}-1}}}{\cosh^{-1}\Omega_s} \tag{3.30}$$

Table (3-2): Element values for Chebyshev lowpass prototype filters (g0= 1.0, Ωc= 1) [1]

For passband ripple L_{Ar} = 0.01 dB

n	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}
1	0.0960	1.0								
2	0.4489	0.4078	1.1008							
3	0.6292	0.9703	0.6292	1.0						
4	0.7129	1.2004	1.3213	0.6476	1.1008					
5	0.7563	1.3049	1.5773	1.3049	0.7563	1.0				
6	0.7814	1.3600	1.6897	1.5350	1.4970	0.7098	1.1008			
7	0.7970	1.3924	1.7481	1.6331	1.7481	1.3924	0.7970	1.0		
8	0.8073	1.4131	1.7825	1.6833	1.8529	1.6193	1.5555	0.7334	1.1008	
9	0.8145	1.4271	1.8044	1.7125	1.9058	1.7125	1.8044	1.4271	0.8145	1.0

For passband ripple L_{Ar} = 0.04321 dB

n	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}
1	0.2000	1.0								
2	0.6648	0.5445	1.2210							
3	0.8516	1.1032	0.8516	1.0						
4	0.9314	1.2920	1.5775	0.7628	1.2210					
5	0.9714	1.3721	1.8014	1.3721	0.9714	1.0				
6	0.9940	1.4131	1.8933	1.5506	1.7253	0.8141	1.2210			
7	1.0080	1.4368	1.9398	1.6220	1.9398	1.4368	1.0080	1.0		
8	1.0171	1.4518	1.9667	1.6574	2.0237	1.6107	1.7726	0.8330	1.2210	
9	1.0235	1.4619	1.9837	1.6778	2.0649	1.6778	1.9837	1.4619	1.0235	1.0

For passband ripple L_{Ar} = 0.1 dB

n	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}
1	0.3052	1.0								
2	0.8431	0.6220	1.3554							
3	1.0316	1.1474	1.0316	1.0						
4	1.1088	1.3062	1.7704	0.8181	1.3554					
5	1.1468	1.3712	1.9750	1.3712	1.1468	1.0				
6	1.1681	1.4040	2.0562	1.5171	1.9029	0.8618	1.3554			
7	1.1812	1.4228	2.0967	1.5734	2.0967	1.4228	1.1812	1.0		
8	1.1898	1.4346	2.1199	1.6010	2.1700	1.5641	1.9445	0.8778	1.3554	
9	1.1957	1.4426	2.1346	1.6167	2.2054	1.6167	2.1346	1.4426	1.1957	1.0

Sometimes, the minimum return loss L_R in the passband is specified instead of the passband ripple L_{Ar}. If the minimum passband return loss is L_R dB ($L_R<0$), the corresponding passband ripple is [1]:

$$L_{Ar} = -10\log\left(1-10^{0.1L_R}\right) \quad dB \tag{3.31}$$

$$L_R = 10\log\left(1-10^{-0.1L_{Ar}}\right) \quad dB \tag{3.32}$$

3.3.4 Frequency and Element Transformation

Since in the lowpass prototype filters, that have a normalized source resistance/conductance g_0= 1 and a cutoff frequency Ω_c= 1. It is important to get frequency characteristics and element values for practical filters based on the lowpass prototype, one may apply frequency and element transformations or frequency mapping [1,5].

The lowpass prototype filter is to be transformed to bandpass response having a passband ω_2-ω_1 where ω_1 and ω_2 indicate the passband-edge angular frequency. The required frequency transformation is [1,5]

$$\Omega = \frac{\Omega_c}{FBW}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \tag{3.33}$$

with

$$FBW = \frac{\omega_2 - \omega_1}{\omega_0} \tag{3.34}$$

$$\omega_0 = \sqrt{\omega_1 \omega_2} \tag{3.35}$$

where ω_0 denotes the centre angular frequency and FBW is defined as the fractional bandwidth. The transformation for the g values are indicated in Figure 3.9

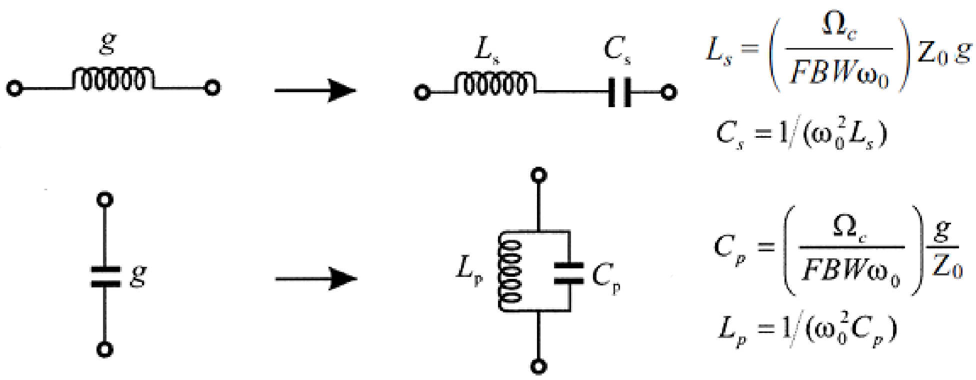


Figure (3.9): Basic element transformation for a lowpass prototype to bandpass transformation[1].

For g representing inductance in the lowpass prototype, (g=L_{LP}),

element transformation is [1,5]:

$$L_s = \frac{\Omega_c}{\omega_0 FBW} Z_0 L_{LP} \tag{3.36}$$

$$Cs = \frac{FBW}{\omega_0 \Omega_c} \frac{1}{Z_0 L_{LP}} \tag{3.37}$$

For g representing capacitance in the lowpass prototype, (g=C_{LP}),

element transformation is [1,5]:

$$L_P = \frac{\Omega_c}{\omega_0 FBW} \frac{C_{LP}}{Z_0} \tag{3.38}$$

$$C_P = \frac{FBW}{\omega_0 \Omega_c} \frac{Z_0}{C_{LP}} \tag{3.39}$$

3.4 Coupled Resonators

This section will talk about coupling, coupled resonators, bandpass filters and hairpin line bandpass filters.

3.4.1 Introduction

It is difficult to implement the bandpass and bandstop filters because they have both series and parallel resonant circuits and the values of some of the components are very different; so coupled resonators may be used.

Coupled resonator circuits are of important to design of RF/microwave filters, specially the narrow-band bandpass filters that essential in many applications. There is a general technique for designing coupled resonator filters which means that it can be applied to any physical structure of the resonator. It has been applied to the design of waveguide filters, dielectric resonator filters, ceramic combline filters, microstrip filters... etc. The design method is based on coupling coefficients of intercoupled resonators and the external quality factors of the input and output resonators.

3.4.2 General Theory of Coupling

After determining the desired filtering characteristic, the next important step is to establish the relationship between the value of every required coupling coefficient and the physical structure of coupled resonators as shown in Figure 3.10 to find the physical dimensions of the filter for fabrication.

In general, the coupling coefficient of coupled RF/microwave resonators, which can be different in structure and can have different self-resonant frequencies, may be defined on the basis of the ratio of coupled energy to stored energy as shown in equation 3.40.

$$k = \frac{\iiint \varepsilon \underline{E}_1 \cdot \underline{E}_2 dv}{\sqrt{\iiint \varepsilon |\underline{E}_1|^2 dv \times \iiint \varepsilon |\underline{E}_2|^2 dv}} + \frac{\iiint \mu \underline{H}_1 \cdot \underline{H}_2 dv}{\sqrt{\iiint \mu |\underline{H}_1|^2 dv \times \iiint \mu |\underline{H}_2|^2 dv}} \tag{3.40}$$

where $\underline{E}$ and $\underline{H}$ represent the electric and magnetic field vectors, respectively. It is clear that all fields are determined at resonance, and the volume integrals are over all effected regions with permittivity of ε and permeability of μ. The first term on the right-hand side represents the electric coupling and the second term the magnetic coupling. The direct evaluation of the coupling coefficient from Equation (3.40) requires knowledge of the field distributions and performance of the space integrals. This is not an easy task unless analytical solutions of the fields exist.

It may be much easier by using full-wave EM simulation or experiment to find some characteristic frequencies that are associated with the coupling of coupled RF/microwave resonators. The coupling coefficient can then be determined against the physical structure of coupled resonators if the relationship between the coupling coefficient and the characteristic frequencies is established.

The coupling coefficient may be generally formalized into Equation 3.41

$$k = \pm \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \tag{3.41}$$

The sign of coupling may only be a matter for cross-coupled resonator filters that does not matter in this work.

Coupled microstrip filters have two typical input/output (I/O) coupling structure are shown in Figure 3.11. The first type is tapped line structure and the second is coupled line structure.

For the tapped line structure, usually a 50 ohm feed line is directly tapped onto the I/O resonator and the coupling or the external quality factor is controlled by the tapping position *t*, as indicated in Figure 3.11.a. The coupling of coupled line structure can be found from the coupling gap *g* and the line width *w* as indicated in Figure 3.11.b. Equation 3.42 extract the external quality factor in the singly loaded resonator.

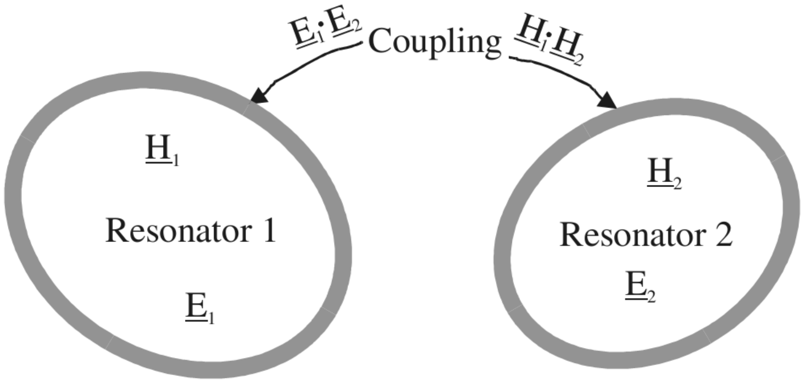


Figure (3.10): General coupled RF/microwave resonators where resonators 1 and 2 can be different in structure and have different resonant frequencies [1].

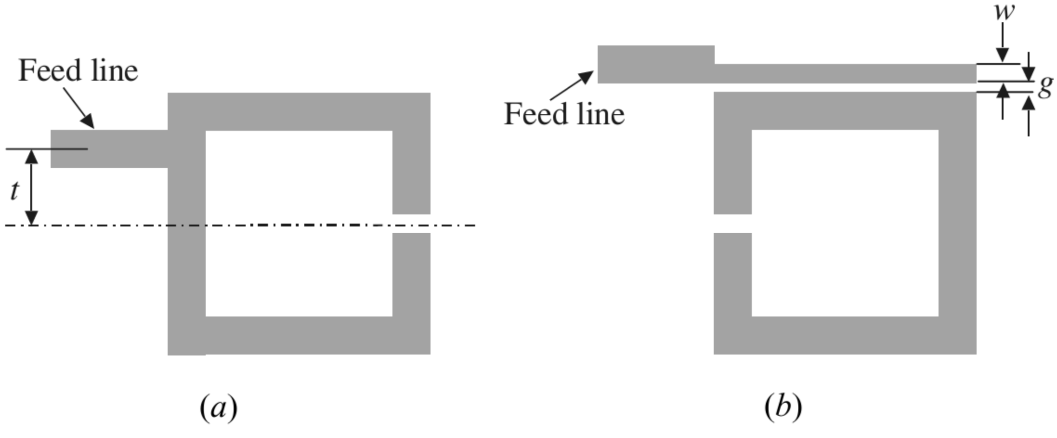


Figure (3.11): Typical I/O coupling structures for coupled resonator filters. (a) Tapped-line coupling. (b) Coupled-line coupling [1].

$$Q_e = \frac{\omega_0}{\Delta\omega_{\pm 90^\circ}} \tag{3.42}$$

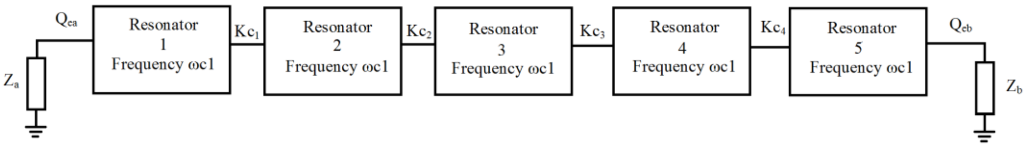
These can be calculated directly from the low pass prototype g-values:

$$Q_{ea} = \frac{g_0 g_1}{FBW} \tag{3.43}$$

$$Q_{eb} = \frac{g_n g_{n+1}}{FBW} \tag{3.44}$$

$$Kc_k = \frac{FBW}{\sqrt{g_k g_{k+1}}} \tag{3.45}$$

where *FBW* is a shortcut for a fractional bandwidth that will equal $FBW=BW/f_0$ where f_0 is the center frequency of any single resonator. This can be shown in Figure 3.12



$$\boxed{Q_{ea} = \frac{g_0 g_1}{FBW}, \quad Q_{eb} = \frac{g_n g_{n+1}}{FBW}, \quad kc_k = \frac{FBW}{\sqrt{g_k g_{k+1}}}}$$

Figure (3.12): The structure of coupled immittance inverter

3.4.3 Hairpin line bandpass filter

As discussed before, the resonator of the bandpass filters can be any shape. One of these shapes is a U-shape as shown in Figure 3.13. If the resonators have a U-shape; then it will be defined as a U-shape resonator or a hairpin resonator. Then the filter will be a hairpin-line bandpass filter.

Hairpin-line bandpass filters are compact structures[1]. They may basically be obtained by folding the resonators of parallel-coupled, half-wavelength resonator filters. So they may use the same design equations for the parallel-coupled, half-wavelength resonator filters[1]. But it is necessary to take into account the reduction of the coupled-line lengths, which reduces the coupling between resonators. Also, if the two arms of each hairpin resonator are closely spaced, they function as a pair of coupled line themselves, which can have an effect on the coupling as well[1].

3.5Filter Design Procedure

To design a bandpass filter it is necessary to follow some steps of design procedure using a simulator program like HFSS, CST ...etc. . The necessity to use the simulator programs come from economically reasons and the ease to test the designs by them without any extra costs compared to the cost of design fabrication and the modifications in the design fabrications.

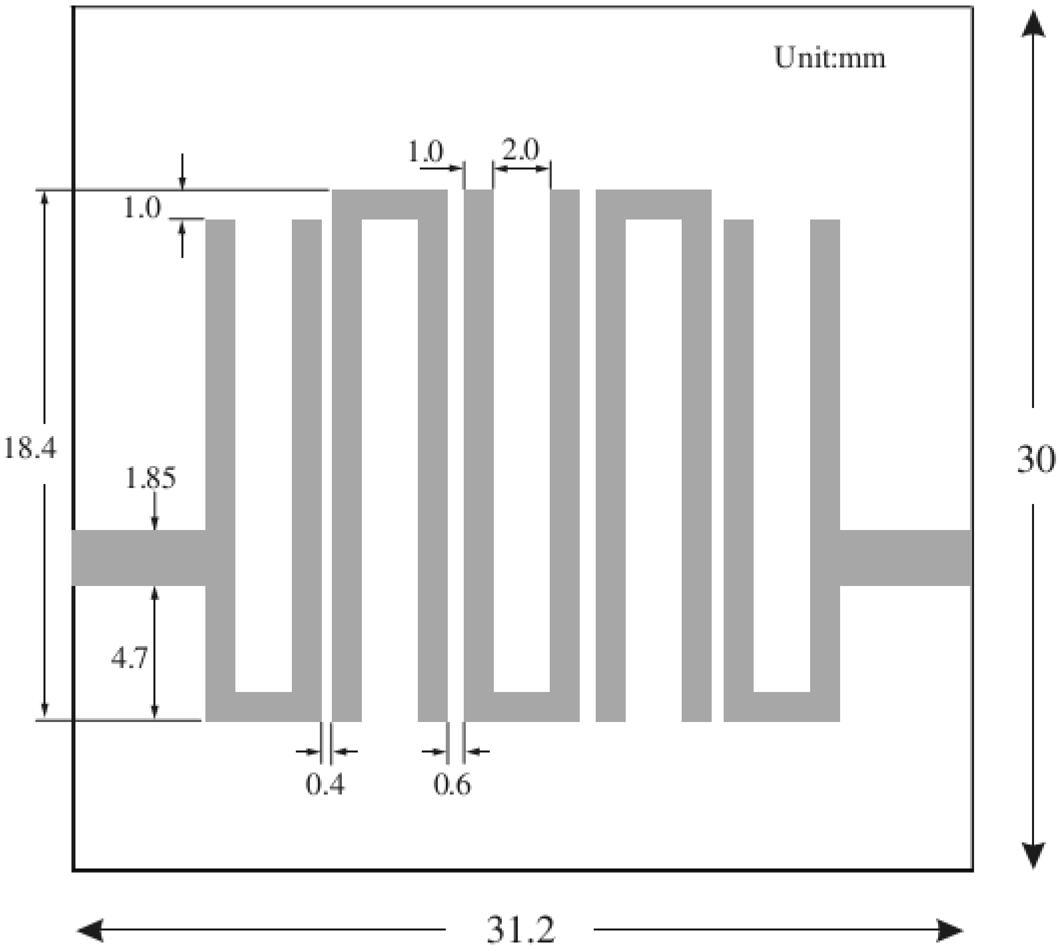


Figure (3.13): Layout of a five-pole, hairpin-line microstrip bandpass filter [1].

The steps of the filter's design procedure can be listed as follow:

1. Determine the filter design requirements like center frequency f_0, filters bandwidth BW, the fractional bandwidth $FBW=BW/f_0$, external quality factors Q_{ea} and Q_{eb} and coupling coefficients K_{ij}.

2. Choose the shape and structure for the resonators and the final structure of the filter, which is chosen as 3rd order hairpin line band pass filter.

3. Choose the simulation program that is chosen as HFSS.

4. Find the dimensions of a single resonator that achieve the proposed center frequency using Eigen mode in HFSS as shown in Figure 3.14.

5. For a single resonator, find the feeders' type, dimensions and location to get the required external quality $Q_{ea,b}$ that is shown in Figure 3.15 from equations (3.43, 3.44) that can be get from equation

$$Q_e=f_0/BW_{3dB} \tag{3.46}$$

by ignoring the effect of other resonators K_c as shown in Figure 3.16.

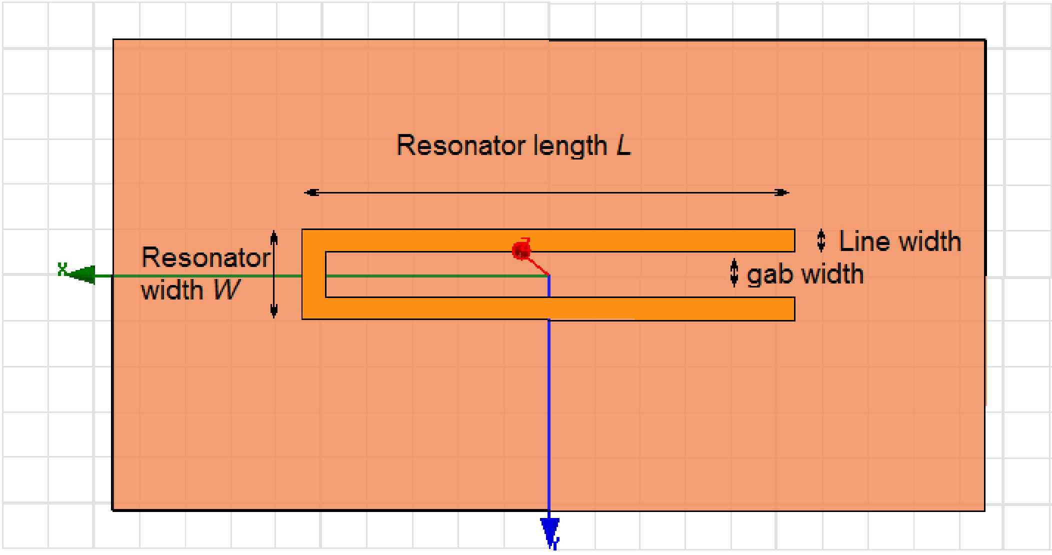


Figure (3.14): A single resonator in Eigen mode

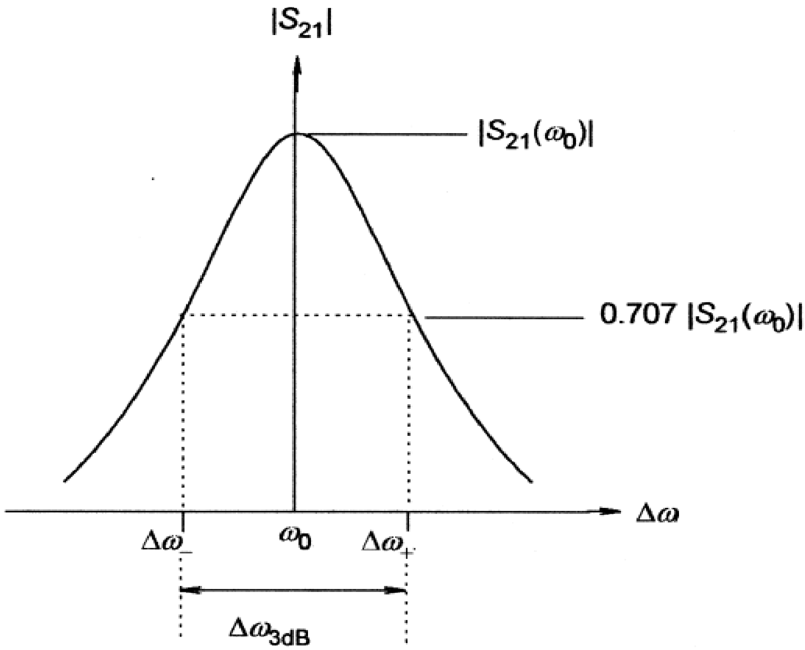


Figure (3.15): External quality factor

6. The required inter resonator coupling from equation (3.45) will be generally as in Figure 3.17. It can be deduced by studying two resonators and make the resonators weakly coupled to the ports as in Figure 3.18. The coupling coefficient can be calculated as

$$kc = \frac{f_1^2 - f_2^2}{f_1^2 + f_2^2} \tag{3.47}$$

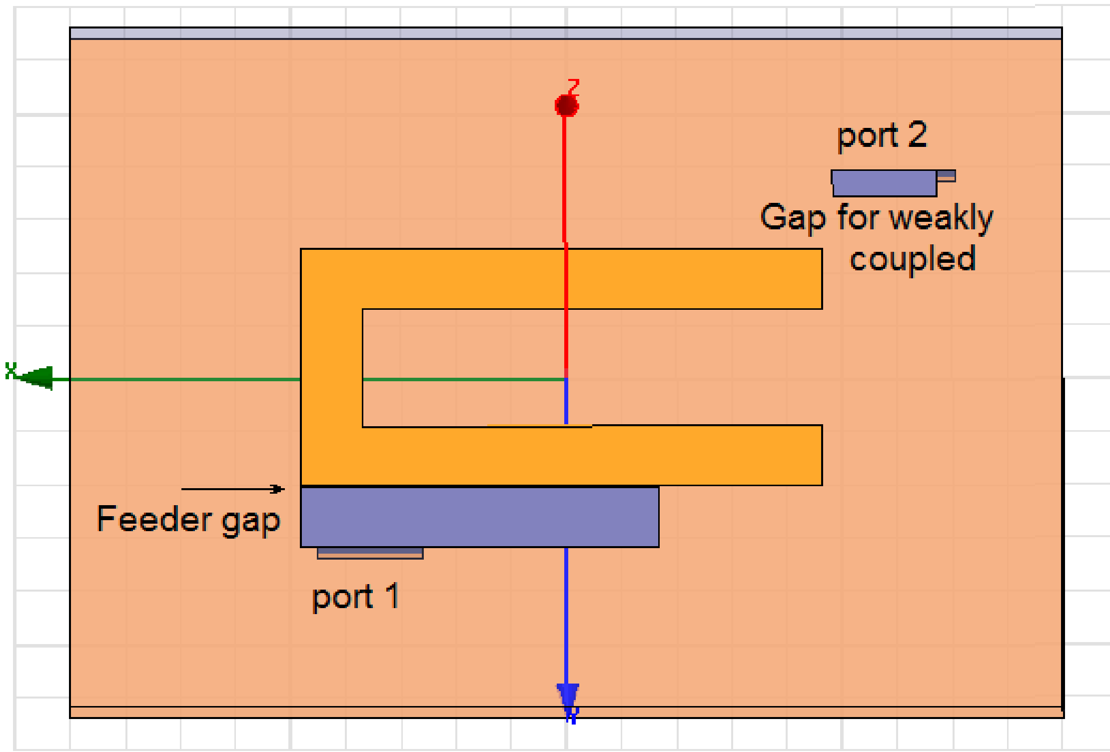


Figure (3.16): Port 2 is weakly coupled and port 1 is coupled

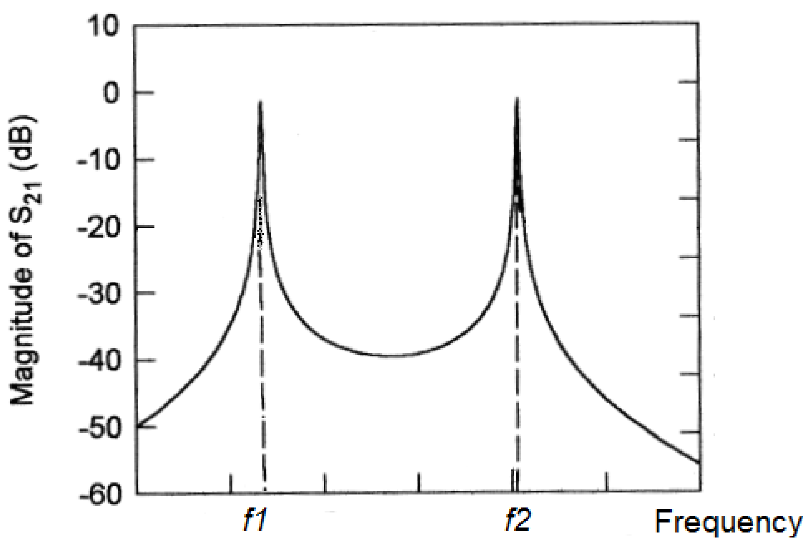


Figure (3.17): General coupling between two resonators

7. Construct the initial design of the whole filter for the required external quality factor in step 5 and required coupling coefficients in step 6, then run the simulation as shown in Figure 3.19.

8. Change the parameters to get the whole design requirements.

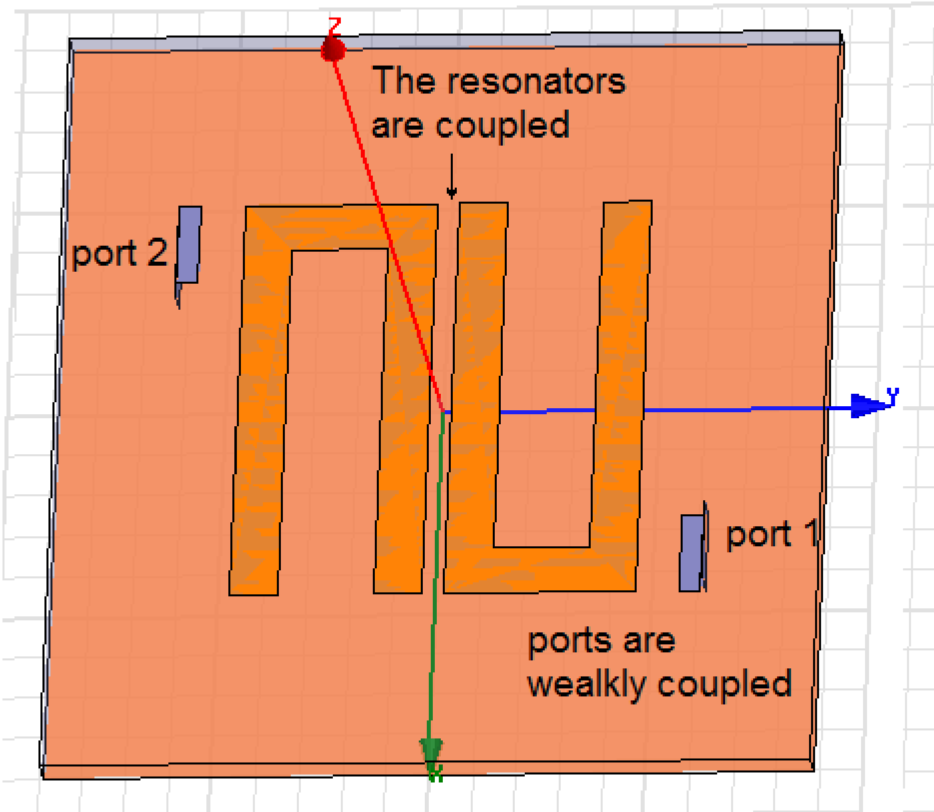


Figure (3.18): Ports are weakly coupled and the resonators are coupled

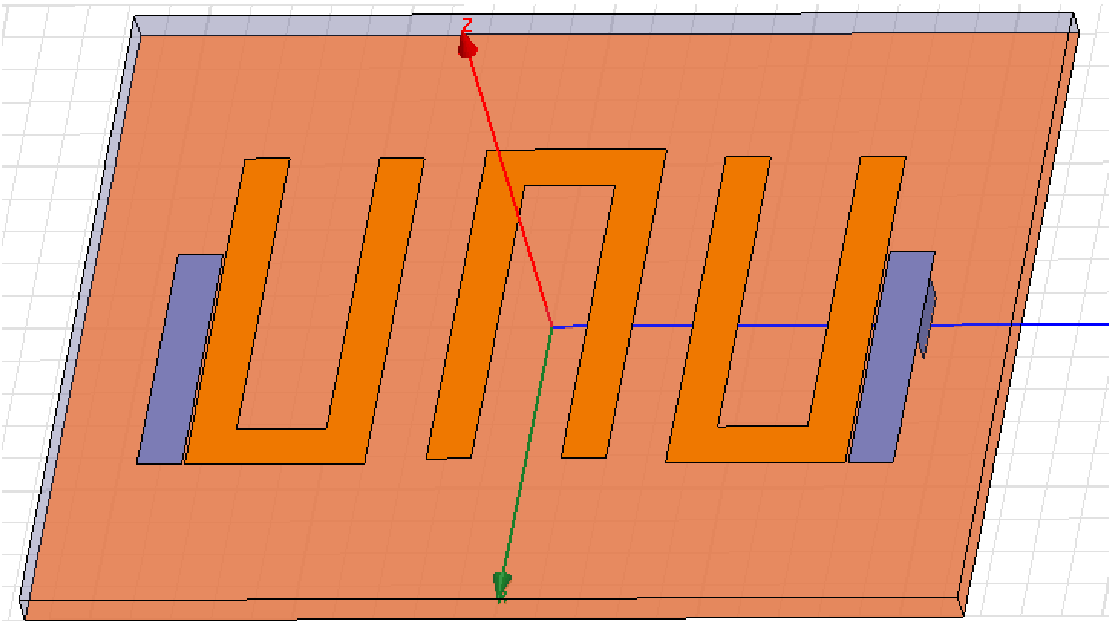


Figure (3.19): The initial filter design

3.6 Summary

This chapter discussed the microwave filters' types and their design principles and parameters. Then it describes the coupling. Then it discussed the hairpin bandpass filters and finally discussed the filters' design procedure.

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Chapter 4 Filtering Antenna Design

This chapter will discuss filtering antenna design procedure. Filtering antenna design begins with designing the antenna (as in Section 4.2) and the filter (as in Section 4.3) separately then uniting the designs (as in Section 4.4) then optimizing the design by varying the parameters to get the overall design specifications.

4.1 Introduction

Filtering antenna design is to design an antenna with filtering features as shown in Section 1.5. But the focus of this work is to use this technology to enhance the antenna parameter specially the bandwidth. The design procedure of filtering antenna begins by identifying the overall specification and selecting the material that will be used in the design.

The proposed overall design is to cover the frequencies of the ISM application. This means that the bandwidth will be from 2400 MHz to 2483.5 MHz. the antenna in the design is rectangular microstrip antenna, and the filter is 3rd order bandpass filter with hairpin shape. The substrate is Rogers 3003 with a dielectric constant of ε_r=3, thickness h=1.52 mm and loss tangent $\tan\delta$=0.0013 [1].

The simulator is High Frequency Structure Simulator HFSS.

4.2 Rectangular Microstrip Antenna Design

This section will discuss the design of a rectangular microstrip antenna and its proposed specifications, equations, calculations simulations and optimizing the results to get optimum return loss.

4.2.1 Microstrip Antenna Design Equations and Calculation

First, before showing the design equations, it is important to know what is needed from the proposed antenna. The antenna must be centered around 2.45 GHz, and the return loss at this point is more than 10 dB, or by other words S_{11} at this point is less than -10 dB and all frequencies that achieve this condition are in the antenna bandwidth.

Now, recall the designing equations from Chapter 2. The antenna width will be calculated from Equation (2.11). The effective dielectric constant ε_{reff} will be calculated from the equation (2.12). The incremental length ΔL that produced by the fringing fields will be calculated from equation (2.13). The actual antenna length L of the patch by equations (2.14) and (2.15)

$$L = \frac{c}{2 f_r \sqrt{\varepsilon_{eff}}} - 2\Delta L \tag{4.1}$$

Now, from Equation (2.11) by substituting f_r = 2.45 GHz and ε_r =3, The width will be W= 43.29 mm, and substitute it in Equation (2.12) to get effective dielectric constant ε_{reff} = 2.839, then Equations (2.13) and (4.1) are used to calculate the antenna length that will be L= 35.09 mm. The 50 ohm transmission line can be calculated from equation (4.2)[2]

$$\frac{W}{h} = \begin{cases} \dfrac{8e^{A}}{e^{2A}-2} & for \quad \dfrac{W}{h} < 2 \\ \dfrac{2}{\pi}\left[B-1-\ln(2B-1)+\dfrac{\varepsilon_r-1}{2\varepsilon_r}\left(\ln(B-1)+0.39-\dfrac{0.61}{\varepsilon_r}\right)\right] & for \quad \dfrac{W}{h} > 2 \end{cases} \tag{4.2}$$

$$where \quad A = \frac{Z_0}{60}\sqrt{\frac{\varepsilon_r+1}{2}} + \frac{\varepsilon_r-1}{\varepsilon_r+1}\left(0.23+\frac{0.11}{\varepsilon_r}\right)$$
$$B = \frac{377\pi}{2Z_0\sqrt{\varepsilon_r}}$$

and vise versa the characteristic impedance can be calculated from equation (4.3) [2,3]

$$Z_0 = \begin{cases} \dfrac{60}{\sqrt{\varepsilon_{eff}}}\ln\left(\dfrac{8h}{W}+\dfrac{W}{4h}\right) & for \quad \dfrac{W}{h} \le 1 \\ \dfrac{120\pi}{\sqrt{\varepsilon_{eff}}\left[\dfrac{W}{h}+1.393+0.667\ln\left(\dfrac{W}{h}+1.444\right)\right]} & for \quad \dfrac{W}{h} > 1 \end{cases} \tag{4.3}$$

The 50 ohm transmission line width will be W_l= 3.82 mm. The initial design is shown in Figure 4.1.

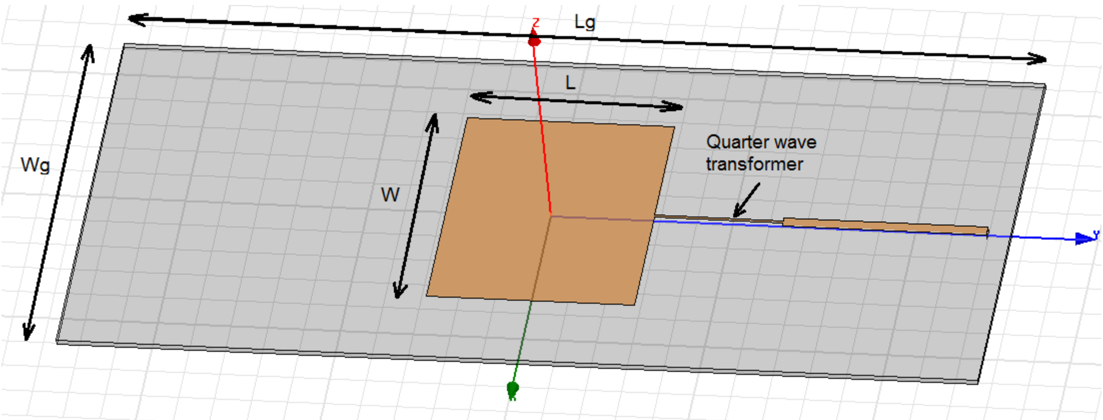

Figure (4.1): The initial design for the rectangular microstrip antenna

The initial simulation results are shown in Figure 4.2.

From Figure 4.2 it is clear that the antenna works at 3.365 GHz as a center frequency and the bandwidth is about 16 MHz and lies between 2.357 and 2.373 GHz. Gain is shown in Figure 4.3 that is 7.1347 dB in maximum.

From the results, it is necessary to modify the antenna to improve the results.

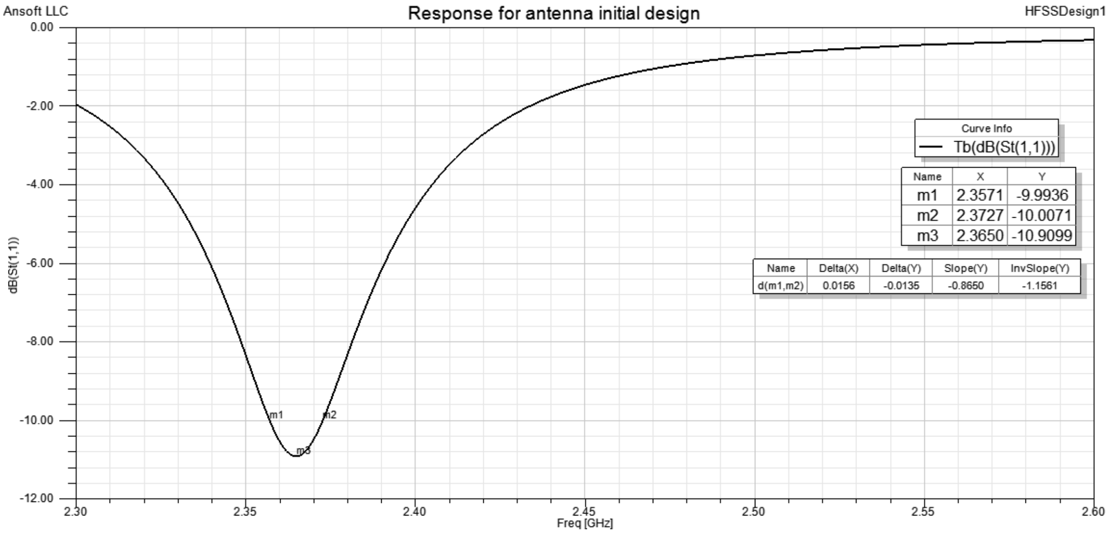

Figure (4.2): Return loss S_{11} for the initial design of rectangular microstrip antenna

4.2.2 Modifications in Antenna Design

It is needed to know the effect of each variable. So, the first step is to study the effects of the size of the ground plane and its corresponding substrate.

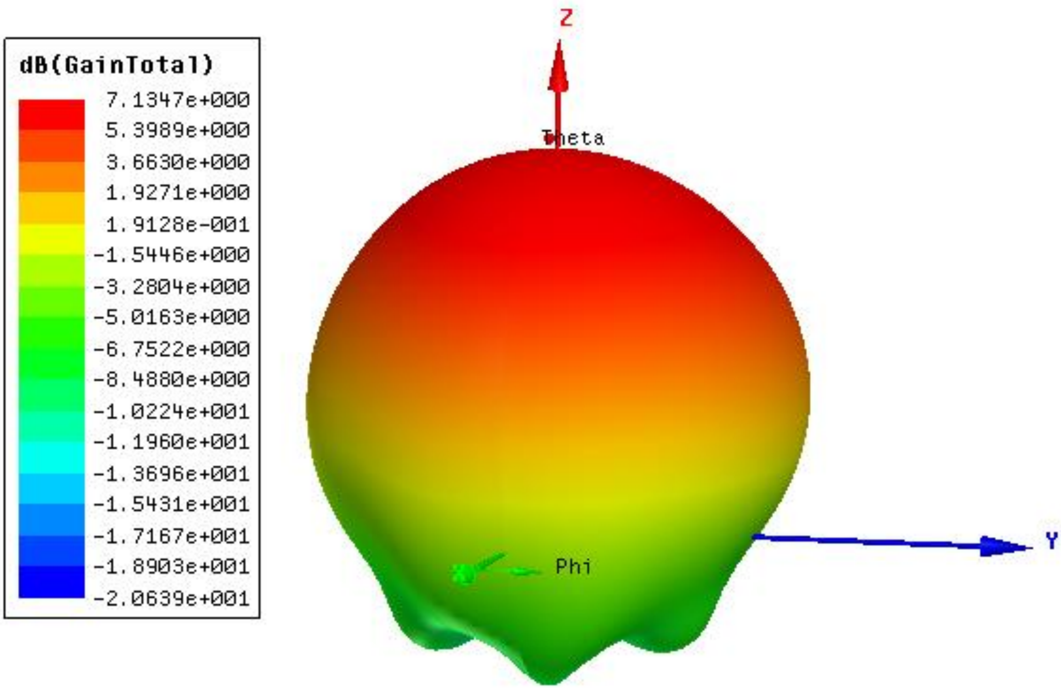

Figure (4.3): Gain for initial design of rectangular microstrip antenna

a. The effect of decreasing L_g

By decreasing the ground plane and the substrate in the direction of L as shown in Figure 4.4 the corresponding return loss is shown in Figure 4.5

It is clear that S_{11} is increased to -8.75 dB that is bad. This makes this antenna less efficeint. This can be referred to decreasing the fringing fields that will decrease the L_{eff}.

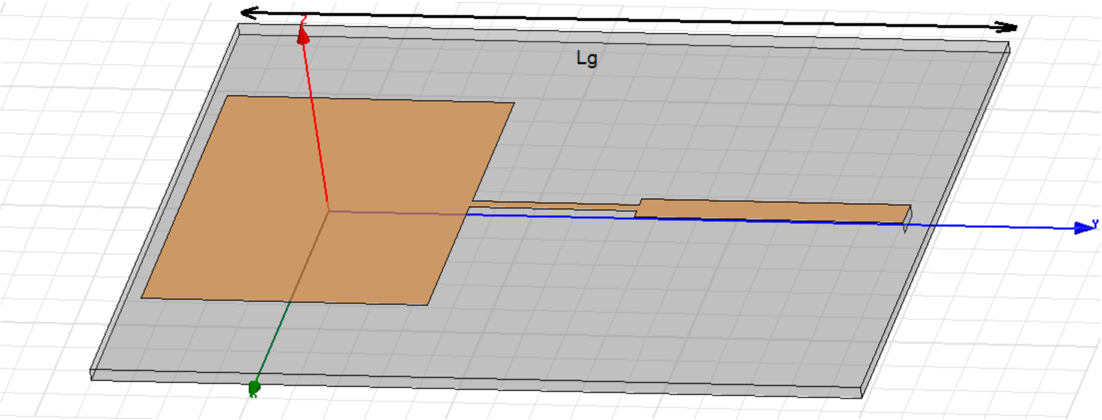

Figure (4.4): The antenna design after decreasing L_g

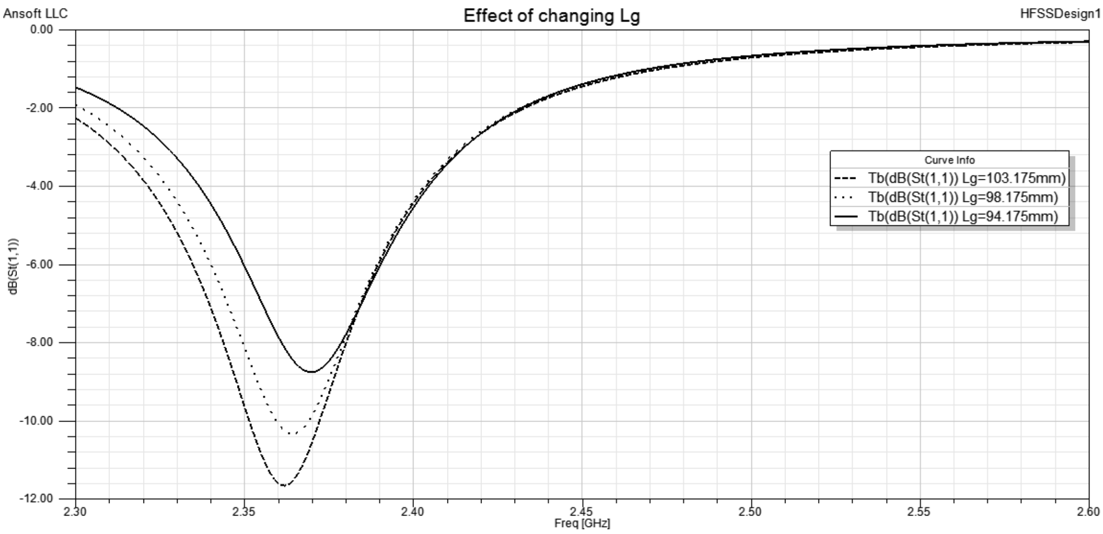

Figure (4.5): S_{11} for the antenna after decreasing L_g

b. The effect of decreasing W_g

By decreasing the ground plane and the substrate in the direction of W as shown in Figure 4.6 the corresponding return loss is shown in Figure 4.7

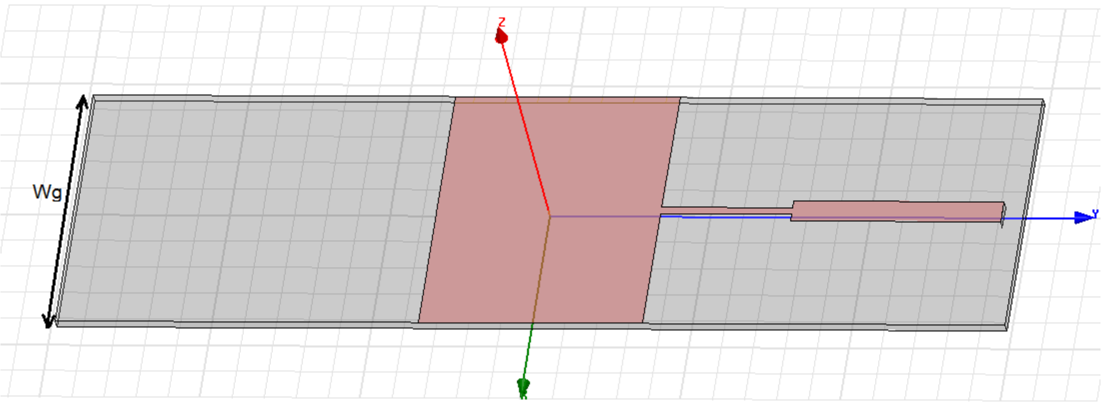

Figure (4.6): The antenna design after decreasing W_g

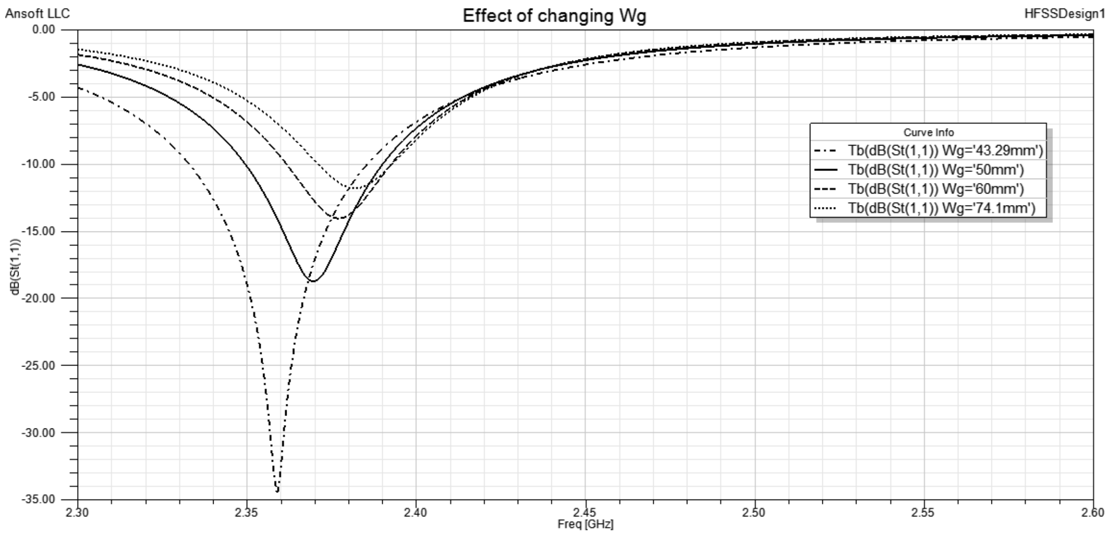

Figure (4.7): S_{11} for the antenna after decreasing W_g

It is clear that the S_{11} is decreased as W_g is decreased that is good. This makes this antenna more efficient. The bandwidth is increased for W_g=W to 52 MHz from 2.333 GHz to 2.385GHz but the center frequency is shifted from 2.382 GHz to 2.359 GHz that is necessary to be shifted to around 2.45 GHz. This can be achieved by changing *W* or *L* or both.

c. The effect of changing *L*

While $Leff=L+2\Delta L$ approximately equals λ/3 to λ/2 this means that *L* can shift the return loss. By decreasing the length of *L* the proposed center frequency should be shifted up. So by changing *L* from 35.09mm to 33.6mm as shown in Figure 4.8 the return loss is shown in Figure 4.9

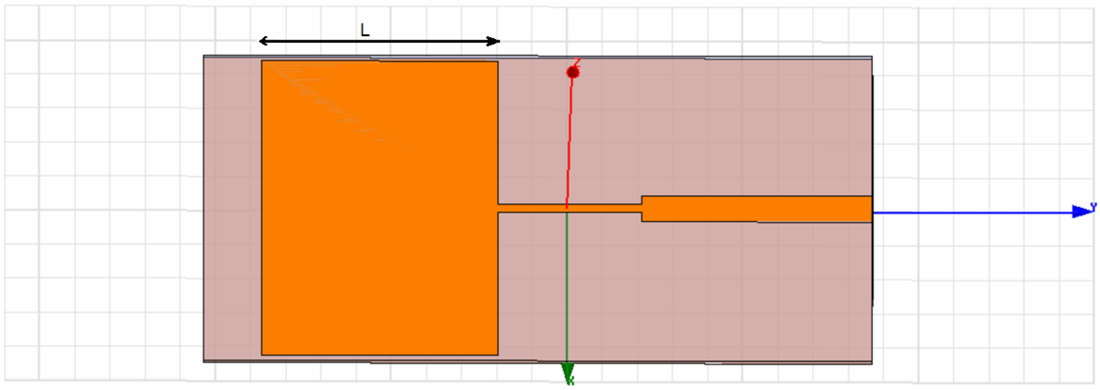

Figure (4.8): The antenna design after changing *L*.

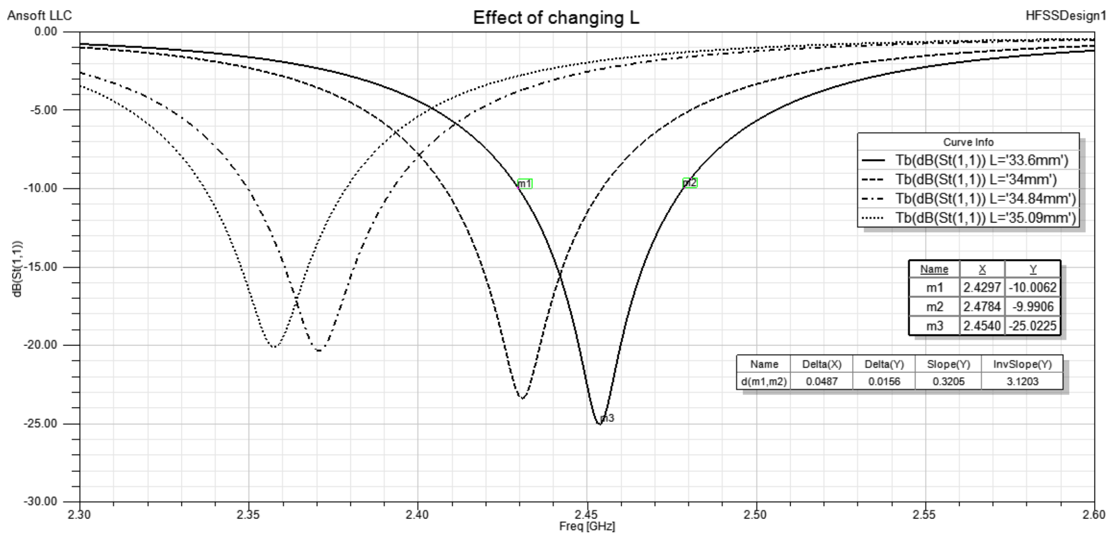

Figure (4.9): S_{11} for a rectangular microstrip antenna by changing L

Figure 4.9 shows that the center frequency is shifted from 2.358 GHz to 2.454 GHz which lies in the proposed bandwidth with S_{11} -25.02 dB and 48.7 MHz bandwidth begins in 2.43 GHz to 2.478 GHz which also lies in the proposed bandwidth.

Also the width of the quarter wave transformer may be changed but it will not affect in the location of the center frequency but it affects on matching of the antenna as shown in Figure 4.10 with L equals 33.6 mm.

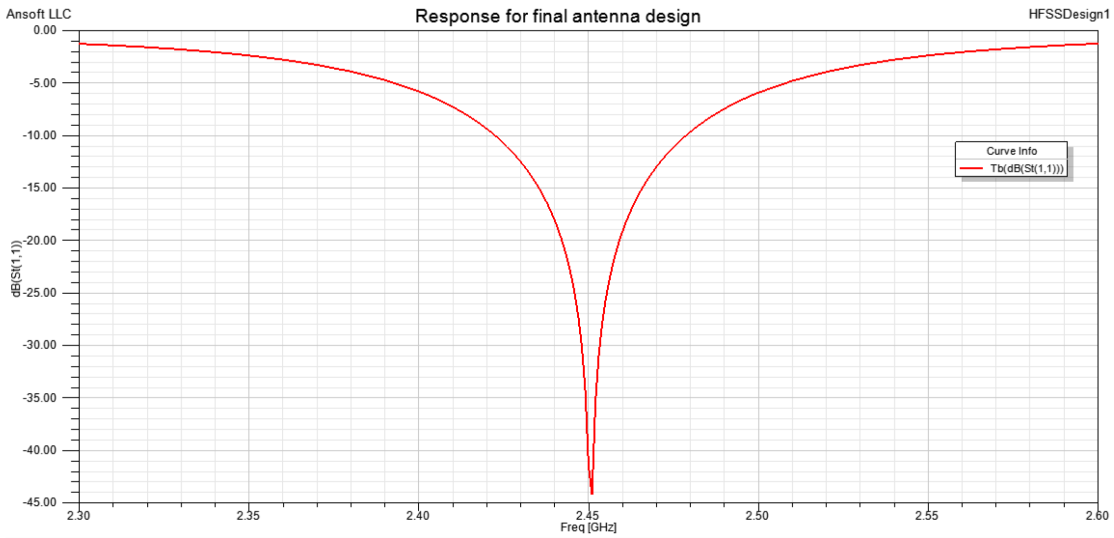

Figure (4.10): Final S_{11} design of the final design

This variable did not change the center frequency that is still located in 2.451 GHz but the resulted S_{11} at this point is decreased from -25.02 dB to -44.16 dB that is better, but the effect on the bandwidth is limited.

The final values of the antenna are: W=43.29 mm, L=33.6 mm, Wg=44.81 mm and Lg=103.541 mm.

4.3 Design of the 3rd order hairpin bandpass filter

This section will discuss the design of a 3rd order bandpass filter with hairpin shape and its proposed specifications, equations, calculations simulations and optimizing the results to get optimum return loss.

4.3.1 Filter Design Specifications

As shown in section 3.6. The first step to design any filter is to identify the proposed filter specifications. So the proposed filter has to deal with 2.4 GHz band that is so called as ISM band "industrial, scientific and medical band" that has a center frequency at 2.45 GHz and its range of frequencies from 2.4 GHz to 2.4835 GHz, so it can be designed in the range from 2.4 GHz to 2.5 GHz that makes the bandwidth 100 MHz as shown in Table 4.1

Table (4-1): Proposed operating frequency and bandwidth

Center frequency	Operating band	Bandwidth
2.45 GHz	2.4 GHz - 2.5GHz	100 MHz

Next step in filter design procedure is to find the element values of the corresponding lowpass prototype filter where its order is three which means that the number of resonators n=3.

Initially let the filter's passband ripple equal 0.1 dB, so from Equation 3.32 the passband return loss will equal -16.43 dB, and from Table 3.2 the g-element values will be $g_0=g_4=1$, $g_1=g_3=1.0316$, $g_2=1.1474$. From Equation 3.34 FBW=2/49=.0408. External quality factors can be calculated from Equations 3.43 and 3.44 that will be $Q_{e1}=Q_{e2}=25.2742$. Coupling coefficients can be calculated from Equation 3.45 that will be $K_{12}=K_{23}=0.0375$.

So the proposed filter calculations are shown in Table 4.2

Table (4-2): Proposed filter calculations

n	FBW	Q_{e1}	Q_{e2}	K_{12}	K_{23}
3	2/49	25.2742	25.2742	0.0375	0.0375

Now the first step is completed.

4.3.2 The Single Resonator Design

The next step as shown in section 3.6 is to choose the shape of the resonator that is hairpin and apply this shape in HFSS simulator in Eigen mode to find the dimension for the proposed center frequency that in 2.45 GHz approximately with the same material as the antenna. The design is shown in Figure 4.11.

The parameters are *resW*, *resG*, *resY1* and *resX1*, where *resY1=2resW+resG* and *resG=2resW*. This means that the parameters that will be changed are *resX1* and *resW*.

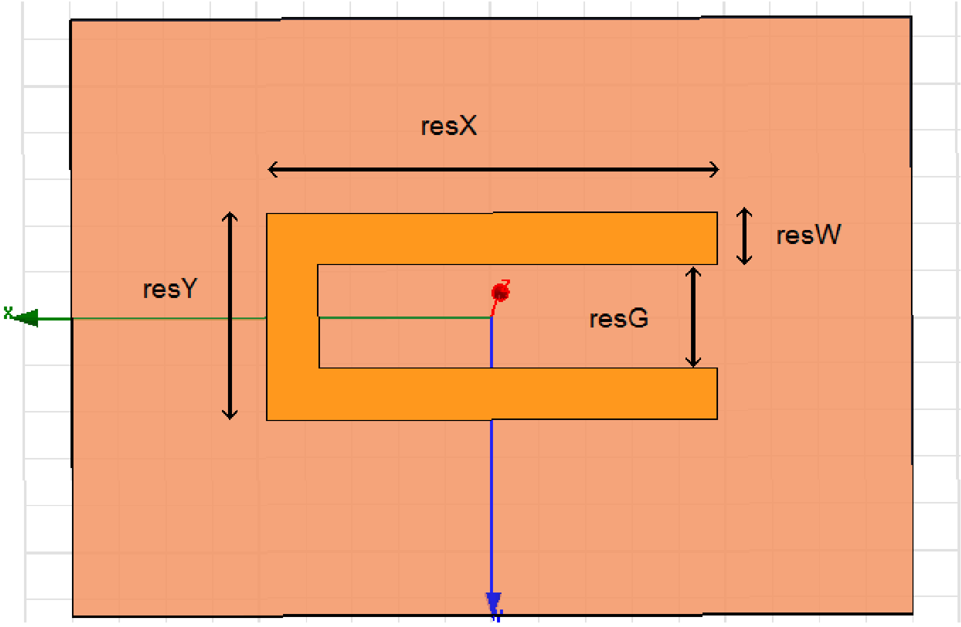

Figure (4.11): Structure of a single resonator in Eigen mode

by changing the values of *resX1* and *resW* it is appears that at *resX1*=19.3 mm and *resW*= 2.22mm the first mode will be at 2.505 GHz as shown in Table 4.3.

Table (4-3): Resonators Dimensions in Eigen mode for Figure 4.12

resX1	*resW*	f_0
19.33 mm	2.22 mm	2.505 GHz

Then it is needed to find the feeder type and location to achieve the proposed external quality factor Q_e = 25.2742.

There are two major types of feeders: center tapped and coupled feeder. First apply the center tapped feeder as shown in Figure 4.12 as the second port is weakly coupled.

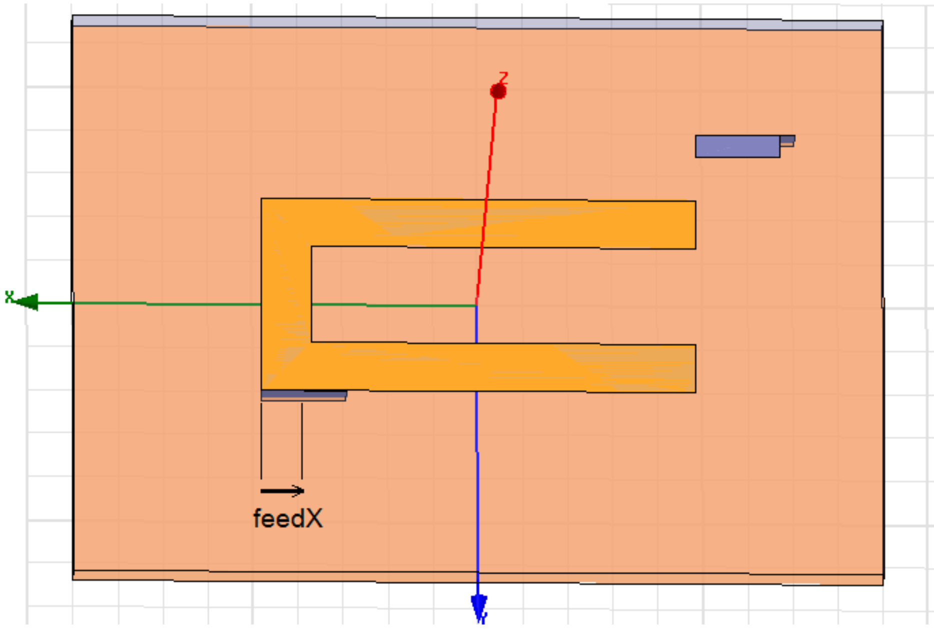

Figure (4.12): Structure of single resonator with port 2 is weakly coupled and port 1 is coupled with a center tapped feeder

By changing *feedX* value. The resulted S_{12} is shown in Figure 4.13 and the resulted external quality factor is shown in Table 4.4.

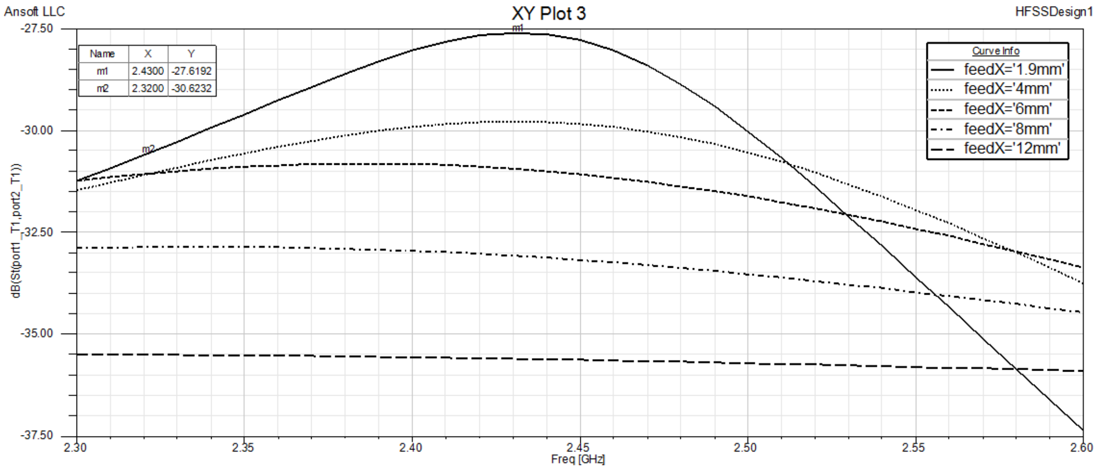

Figure (4.13): S_{12} as the tapped feeder position feedX is changed.

Table (4-4): External quality factor for different locations of feeder

feedX in mm	f_0 in GHz	BW_{3dB} in MHz	Q_e
1.9	2.43	180	13.5
2	2.43	190	12.78947
2.1	2.43	200	12.15
2.25	2.43	210	11.57143
2.5	2.43	225	10.8

It appears that the bandwidth is very large that makes Q_e very low and far from our target. A gap *d* between the port and the resonator can be made to reduce the coupling as shown in Figure 4.14.

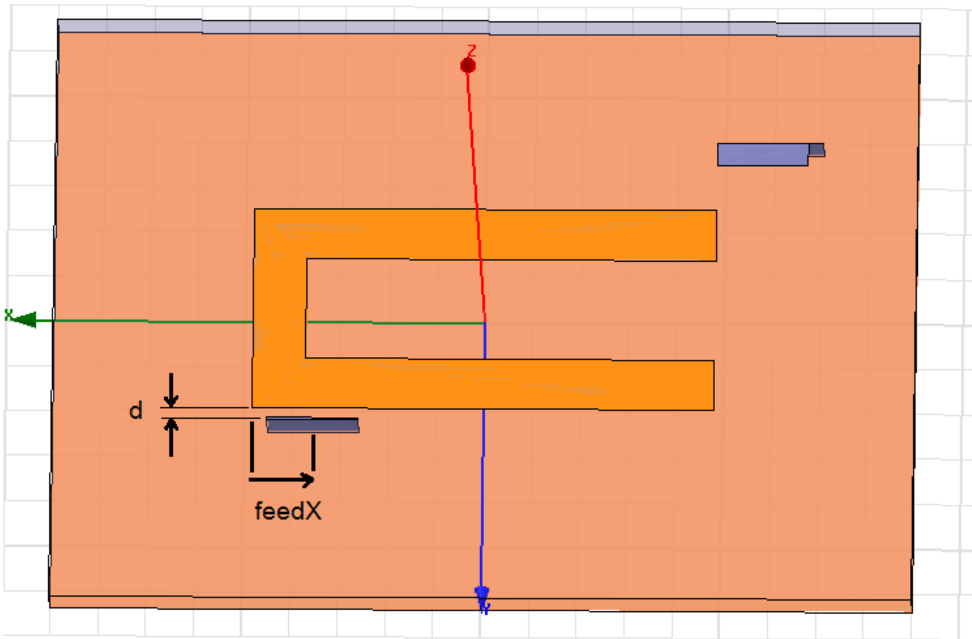

Figure (4.14): Structure of single resonator with port 2 is weakly coupled and port 1 is coupled with a gap between resonator and feeder

Now the bandwidth is much smaller than the proposed bandwidth because the coupling is too weak as shown in Figure 4.15. The resulted external quality factor is shown in Table 4.5.

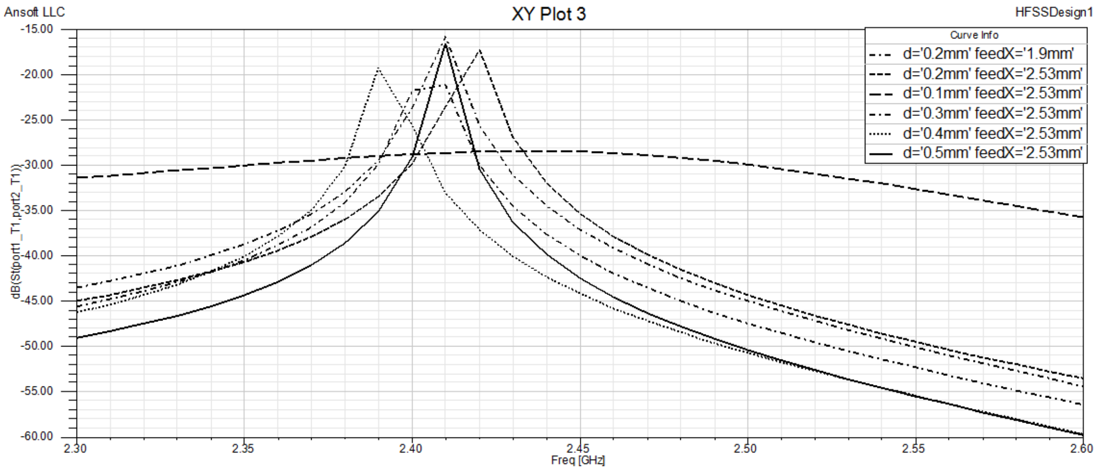

Figure (4.15): S_{12} as the feeder gap d and position feedX are changed.

Table (4-5): External quality factor for coupled feeder with different spacing for Figure 4.15

d in mm	*feedX* in mm	f_0 in GHz	BW_{3dB} in MHz	*Qe*
0.1	2.53	2.435	230	10.59
0.2	1.9	2.41	8	301.25
0.2	2.53	2.42	9	268.89
0.4	2.53	2.39	7	341.43
0.5	2.53	2.41	4	602.50

It appears that center tapped feeder fails to achieve the proposed external quality factor. The coupled feeder may be used.

The coupled feeder is shown in Figure 4.16. The port will be connected to a line that is gapped from the resonator by distance *d* and is parallel to the resonator leg. The increase of length of the line will increase the port coupling to achieve the required bandwidth and proposed external quality factor.

By changing the parameters *feedX* and *d*. It appears that when *feedX* is fixed and *d* is increasing f_o is increasing, and in fixed d and increasing *feedX* the bandwidth is increasing as shown in Table 4.6.

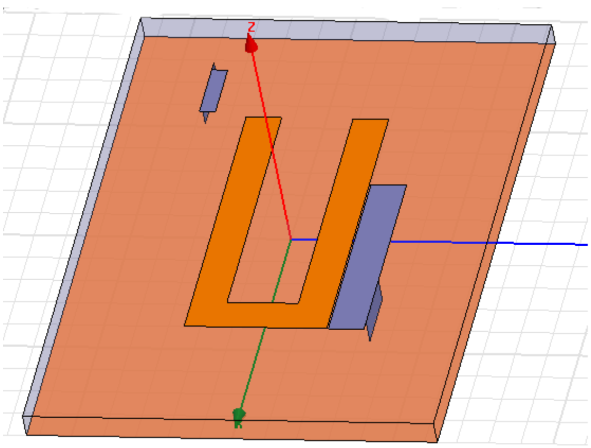

Figure (4.16): Structure of single resonator with port 2 is weakly coupled and port 1 is coupled

Table (4-6): External quality factor for coupled feeder with different spacing for Figure 4.17

d in mm	*feedX* in mm	f_o in GHz	BW_{3dB} in MHz	Q_e
0.5	9	2.45	86.9	28.19
0.5	9.5	2.45	93.3	26.26
0.5	10	2.45	100.3	24.43
0.6	9.5	2.46	93.9	26.2
0.6	10	2.46	>100	<24.6

The chosen values of parameters are d=0.5mm and *feedX*=9.5 that result Q_e=26.26

4.3.3 Coupled resonator design

After designing the single resonator and its feeder. The next step is to find the inter resonator coupling between two resonators to build the initial design of the filter.

In this step it is needed to make both ports weakly coupled to calculate the coupling between two resonators by equation 3.47 using the resulted insertion loss. The design is shown in Figure 4.17.

In this case the effected parameters are *d* which is the distance between the resonators, *feedX* that is feeder position along the leg of the resonator, *dc* that is the small cut from the leg of the resonator "for matching purposes" and gap1 that is the gap that makes the ports weakly coupled. These parameters will result two frequencies f_1 and f_2 as shown in Figure 4.18 that are substituted in equation 3.47 to get the resulted coupling k to find the proposed *kc* as shown in Table 4.7

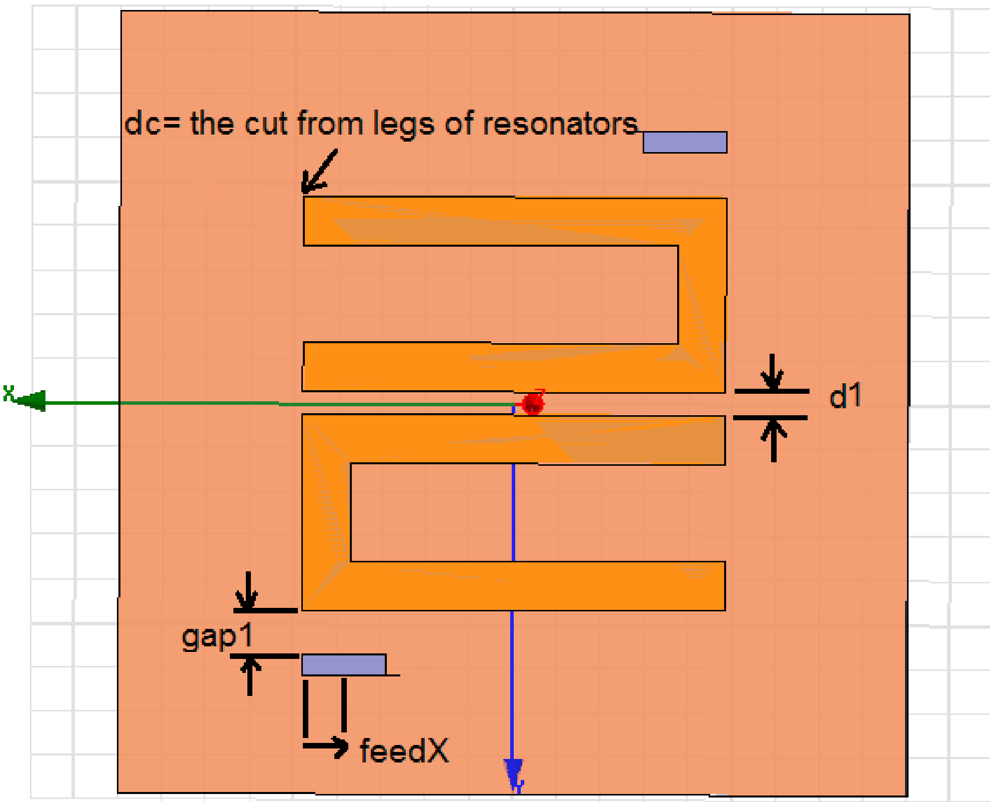

Figure (4.17): Structure of two resonators design with ports are weakly coupled and the resonators are coupled

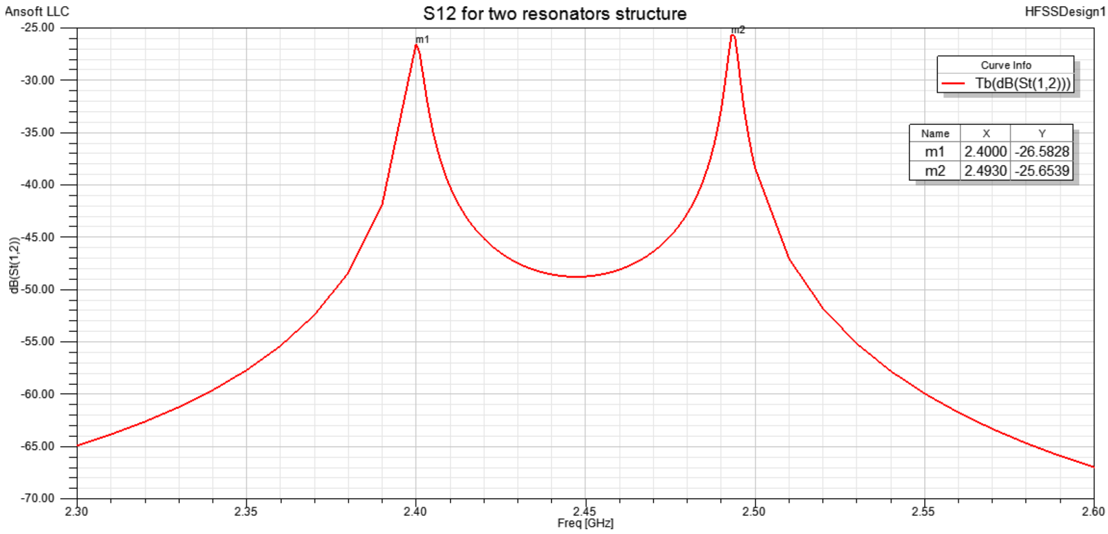

Figure (4.18): S_{12} for structure in Figure 4.18

Table (4-7): Coupling Coefficient for various parameters for two resonators design

dc	*d1*	*feedX*	*gap1*	f_1	f_2	*f2-f1*	*k*
0	2	9	3	2.34	2.47	130	0.0540
0	2.3	9	3	2.31	2.42	110	0.0465
0	2.5	9	3	2.34	2.44	100	0.0418
0	2.6	9	3	2.33	2.43	100	0.0420
0	2.8	9	3	2.35	2.44	90	0.0376
0	2.2	9.5	3	2.35	2.46	110	0.0457
0	2.5	9.5	3	2.36	2.46	100	0.0415
0	3	9.5	3	2.34	2.42	80	0.0336
0	2.6	9	4	2.34	2.44	100	0.0418
0.4	3.5	9.5	3	2.42	2.48	60	0.0245
0.4	3.5	9.5	4	2.41	2.47	60	0.0246
0.4	3.4	9.5	4	2.411	2.475	64	0.0262
0.4	3.3	9.5	4	2.402	2.475	73	0.0299
0.4	3	9.5	4	2.41	2.488	78	0.0318
0.5	3.5	9.5	4	2.418	2.48	62	0.0253
0.5	3.4	9.5	4	2.406	2.471	65	0.0267
0.5	3.3	9.5	4	2.37	2.456	86	0.0356
0.5	3.2	9.5	4	2.409	2.483	74	0.0302
0.5	3.1	9.5	4	2.407	2.483	76	0.0311
0.5	3	9.5	4	2.419	2.496	77	0.0313
0.5	2.9	9.5	4	2.416	2.499	83	0.0338
0.5	2.8	9.5	4	2.4	2.486	86	0.0352
0.5	2.7	9.5	4	2.4	2.493	93	0.0380
0.5	2.6	9.5	4	2.4	2.494	94	0.0384

The chosen values of the parameters are *dc*= 0.5mm, *d1*= 2.7mm, *feedX*=9.5mm and *gap1*=4mm.

4.3.4 Final third order filter design

4.3.4.1 Initial design

After completing steps in previous subsections, it is needed to build the initial design by connecting the three resonators together as shown in Figure 4.19 and run the analyzer to find the resulted return loss S_{11} and S_{12} for this initial design as shown in Figure 4.20

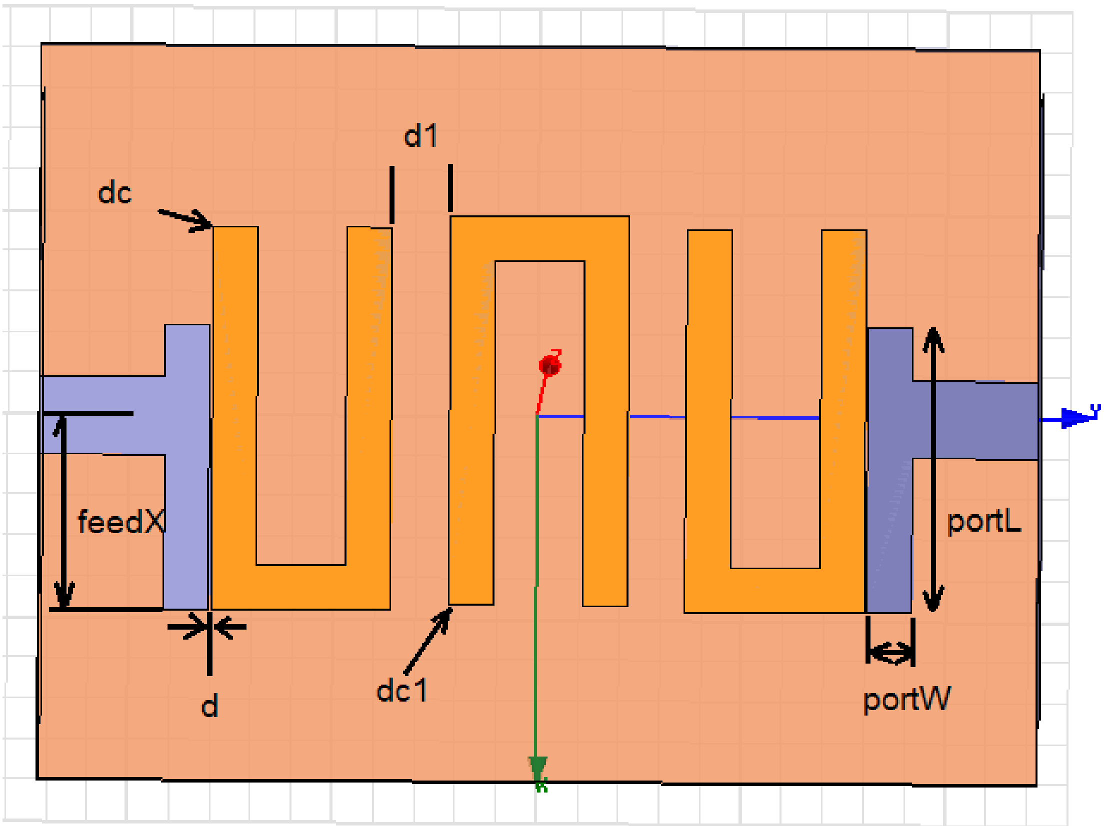

Figure (4.19): Structure of the microstrip 3rd order filter

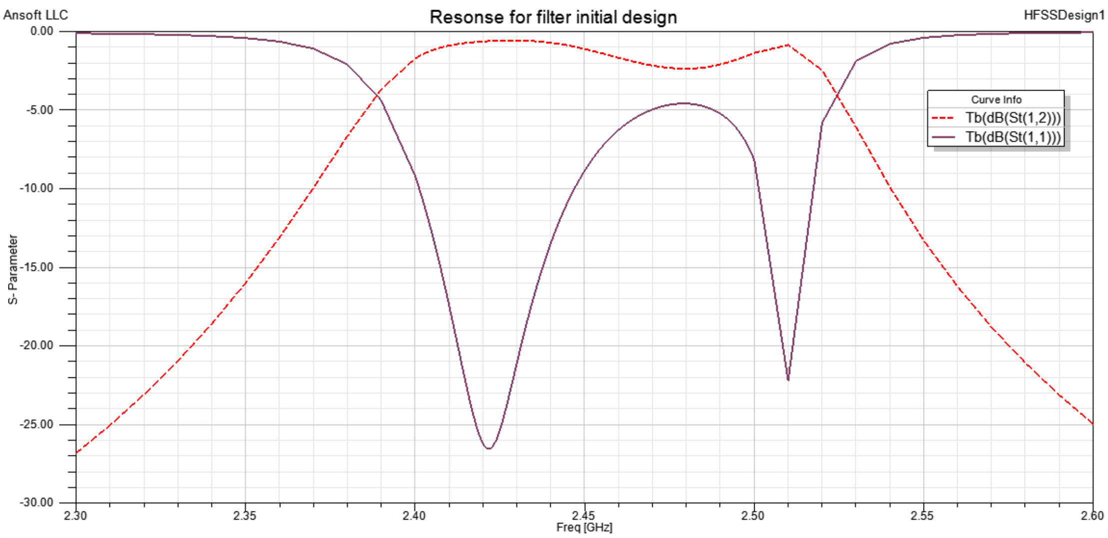

Figure (4.20): Initial response of the filter

4.3.4.2 Modifications in filter design

To reach the final design, it is needed to change the values of some parameters to achieve the system specifications as shown in section 4.3.1. These parameters are *d* that is the gap between the feeder and the resonator, *d1* that is the gap between two resonators, *dc* that is the cut from the legs of the first and last resonators and *dc1* is the cut from the legs of the middle resonator.

The first values of the parameters that achieve the requirements are d=0.15mm, d1=3mm, dc=0.5mm and dc1=0.25mm. The resulted S_{11} and S_{12} are shown in Figure 4.21.

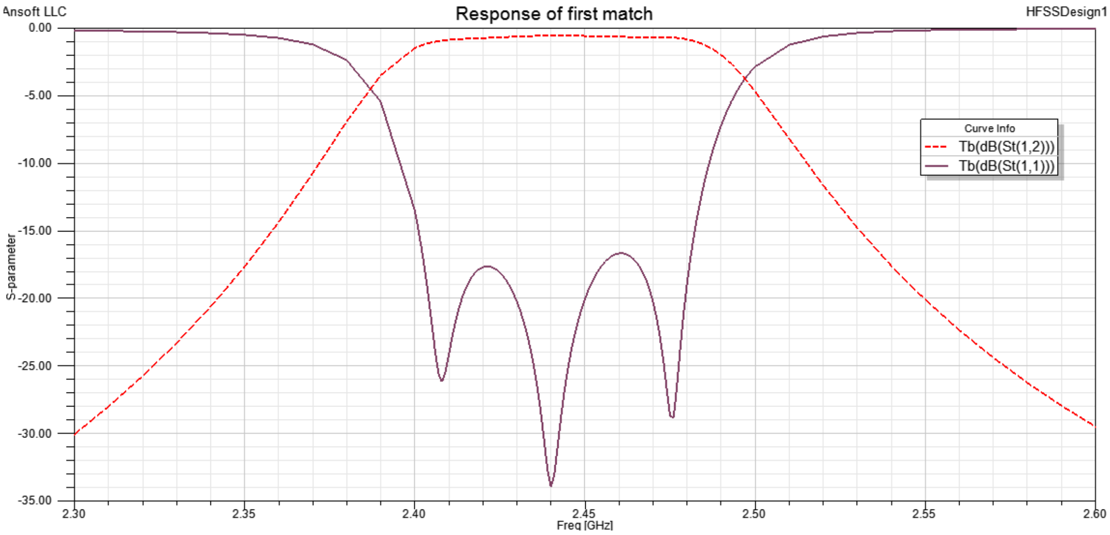

Figure (4.21): Response for first values that match the proposed filter

By optimizing the parameters again, the resulted S_{11} and S_{12} are shown in Figure 4.22 that is the final response of the filter. The Final values of the parameters are d=0.1mm, d1=2.8mm, dc=0.6mm and dc1=0.3mm.

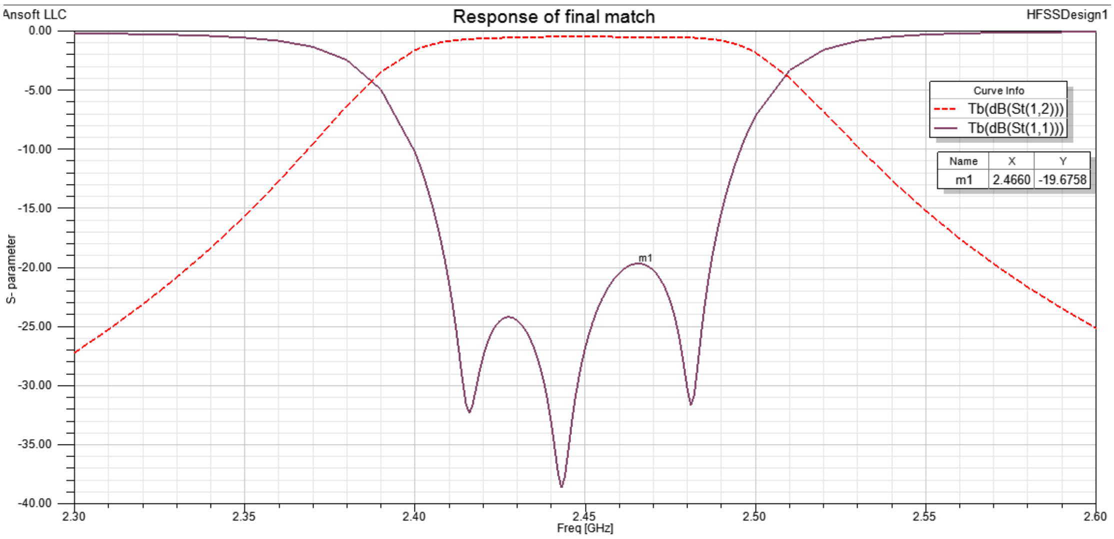

Figure (4.22): Final response of filter

4.4 First model of Filtering Antenna Design:

In section 4.2 the antenna is successfully designed that is the first step in filtering antenna design. Filter is successfully designed in section 4.3 that is the second step. In this section, filtering antenna will be built by combining both designs in one design that will be defined as the first model of filtering antenna design.

4.4.1 Initial Design

Initial design of the filtering antenna can be done by connecting the final design of the antenna in section 4.2 to the final design of the filter from section 4.3 to build the initial design of the filtering antenna as shown in Figure 4.23 and run the analyzer.

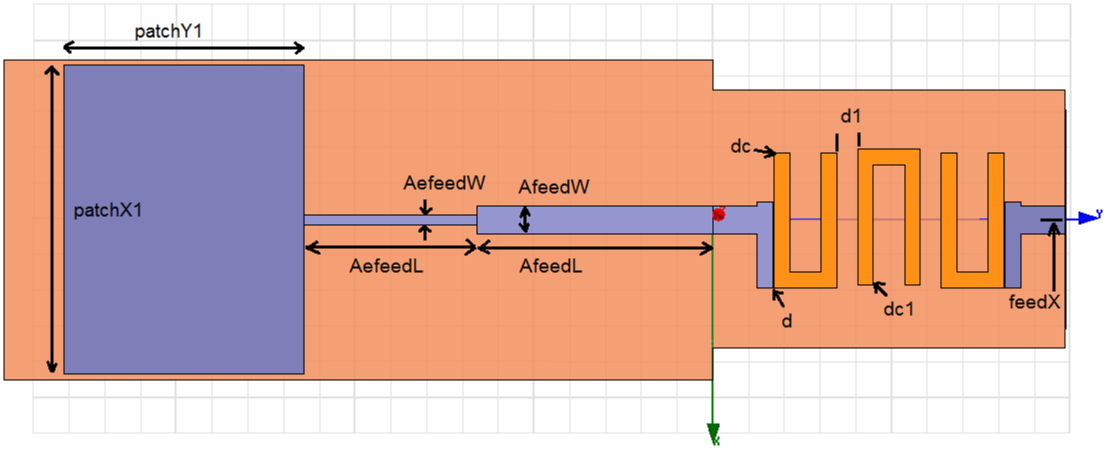

Figure (4.23): First model of Microstrip filtering antenna

The initial return loss is shown in Figure 4.24 that still does not reach the system specifications. This model still needs to be modified to achieve the ISM system specifications.

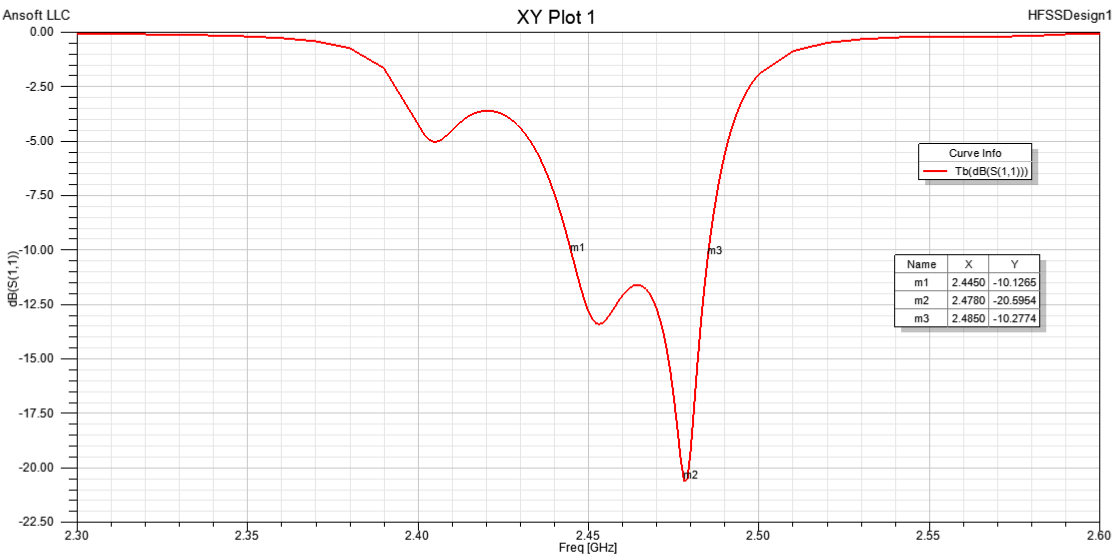

Figure (4.24): Initial response of the first model of filtering antenna.

4.4.2 Modifications in Filtering Antenna Design

After simulating the initial design. The bandwidth does not cover the ISM band. The design needs to be modified to optimize the bandwidth to reach to ISM system requirements.

This can be done by enhancing the matching between the two parts of the filtering antenna (antenna and filter). The final parameters are shown in Table 4.8 and the final return loss and gain are shown in Figure 4.25 and Figure 4.26 respectively.

It appears that the bandwidth of the design is enhanced. The antenna bandwidth of the antenna increased from 57 MHz to 83.5 MHz that is 146.5% from the original bandwidth.

Table (4-8): Parameters of the first model of filtering antenna

Parameter	Description	Value (mm)
patchX1	Rectangular antenna width	43.29
patchY1	Rectangular antenna length	33.6
d1	Spacing between two resonators	2.9
d	The gap between the port and the resonator	0.15
dc	Cut from legs of resonators 1 and 3	0.5
dc1	Cut from legs of resonators 2	0.4
portL	Port length	12.015
AfeedL	Antenna feeder length	33
AfeedW	Antenna feeder width	3.82
AefeedL	Antenna edge feeder length	24.2
AefeedW	Antenna edge feeder width	1.4
feedX	Feeder port position	9.5
feedW	Feeder port width	3.82
portW	Port width	2.22

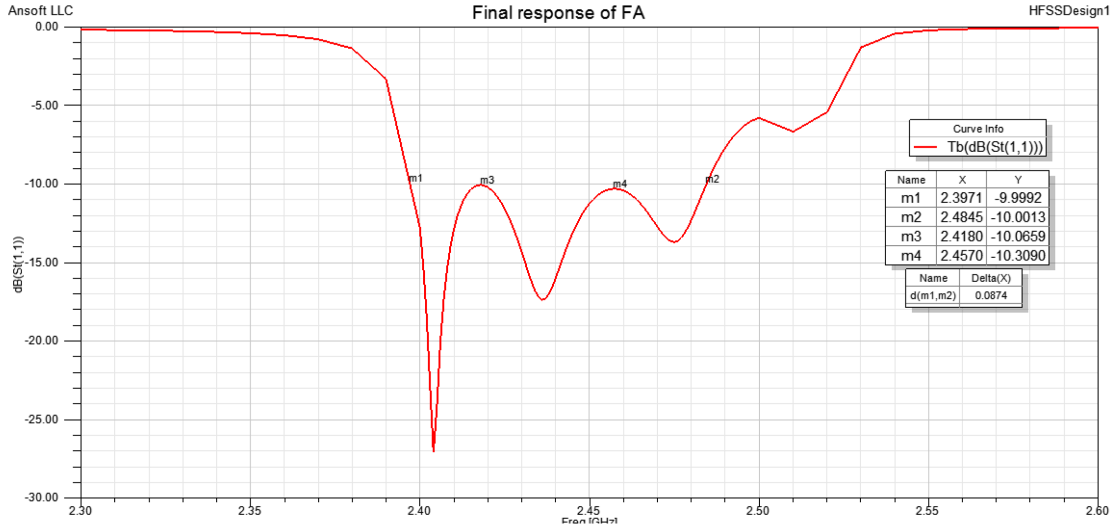

Figure (4.25): Final response of the first model of filtering antenna after optimization.

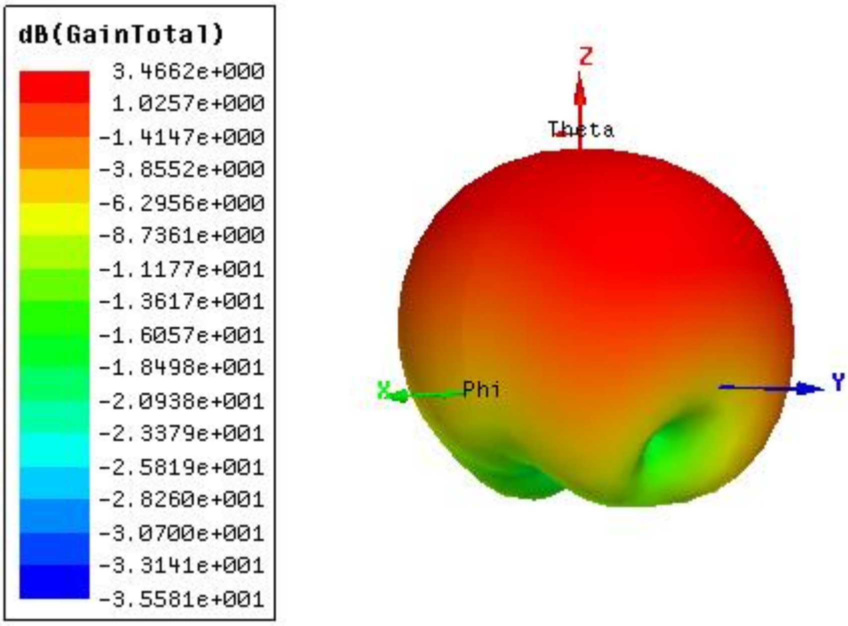

Figure (4.26): Gain of the first model of filtering antenna after optimization.

4.5 Developing Filtering Antenna Design:

The next sections present different designs for filtering antenna to reduce the size. The second model is done by removing the quarter-wave transformer and adding a transmission line with variable length and width. Moreover, adding an open shunt stub is also used to achieve matching in the third model.

4.5.1 Matching using Antenna Feed Line Length

For the structure that is shown in Figure 4.26. Antenna feeder line length *AfeedL* is changed, to achieve matching and enhance the return loss.

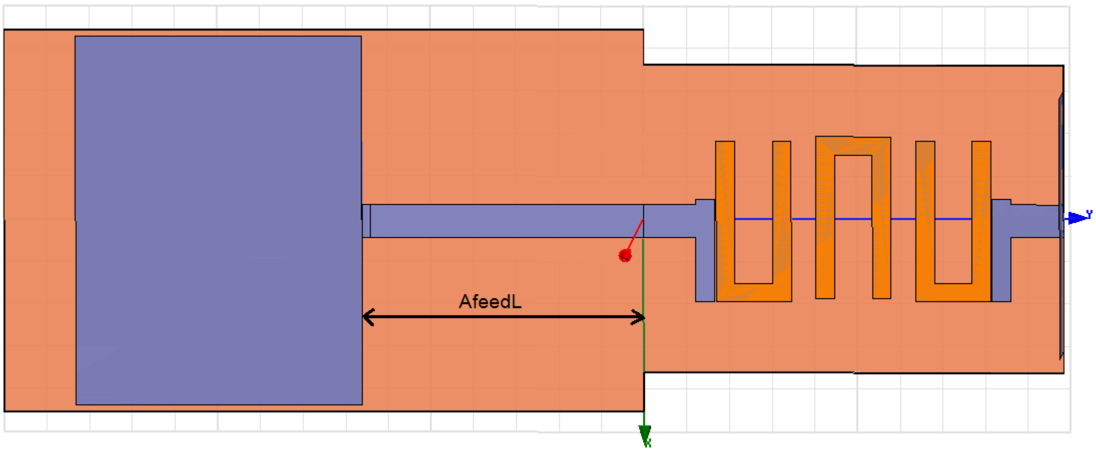

Figure (4.27): Second model of filtering antenna structure.

The initial value of *AfeedL* is 33mm as in the first model. The return loss is shown in Figure 4.26 for values of *AfeedL*: 27mm, 30mm, 33mm and 36mm. The return loss is shown in Figure 4.27. From Figure 4.27 as *AfeedL*=27mm, the fourth return zero appeared at 2.355GHz but further improvement needs to be done by adding more parameters such as the width of the line *AfeedW* as will be shown in the next section.

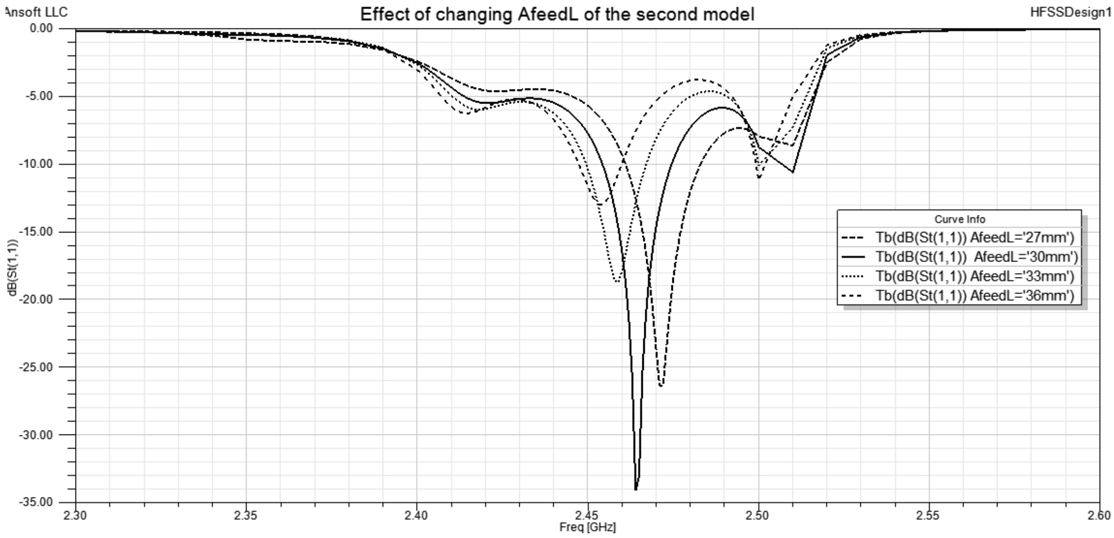

Figure (4.28): Response of structure in Figure 4.26 and the effect of *AfeedL*.

4.5.2 Matching using Antenna Feed Line Length and Width

From the structure in Figure 4.26, the antenna feed line width *AfeedW* is changed to 0.42mm to study its effects on the matching and also *AfeedL* is changed. The response is shown in Figure 4.28.

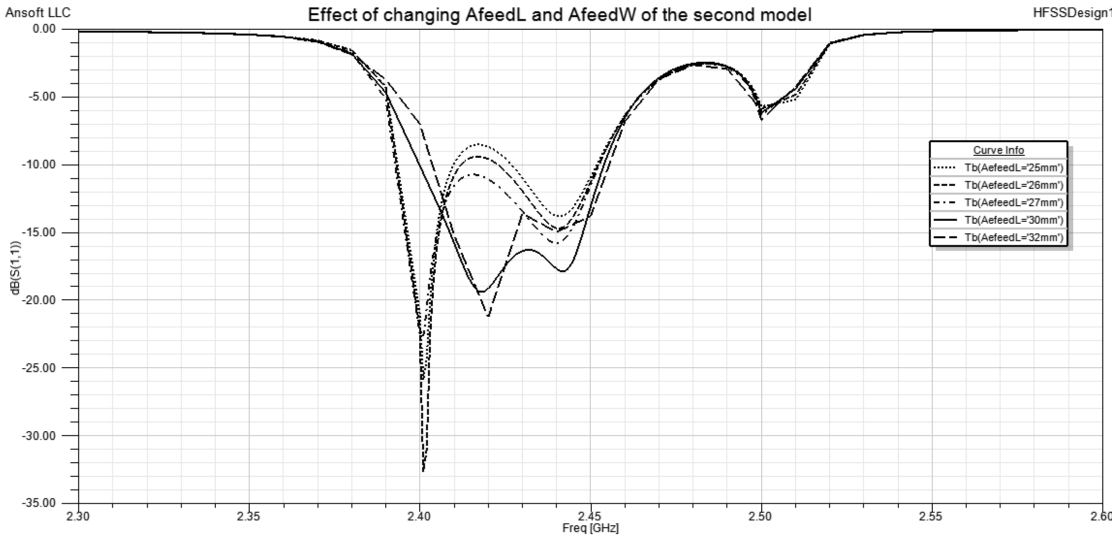

Figure (4.29): Response of structure in Figure 4.27 and the effect of *AfeedL* and *AfeedW*.

The resulted response is promising. The active band (within -10 dB) is increased to 55MHz that begins from 2.4GHz to 2.455GHz for *AfeedW*=0.42mm and *AfeedL*=31mm. These results still needs further work in the future.

4.5.3 Matching using Antenna Feed Line Length and Stubs.

From the structure in Figure 4.26, open shunt stubs will be added to the design as in Figure 4.29. This design will be defined as the third model of the filtering antenna. The parameters of this model will be *AfeedL*, *stubL*, *stubL1* and *stubW* as shown in Figure 4.29.

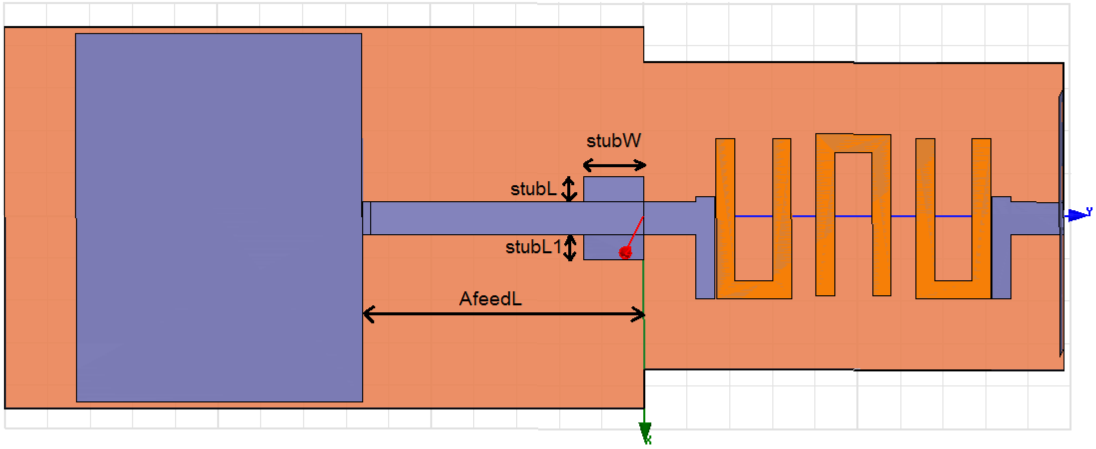

Figure (4.30): Third model of filtering antenna structure with stubs.

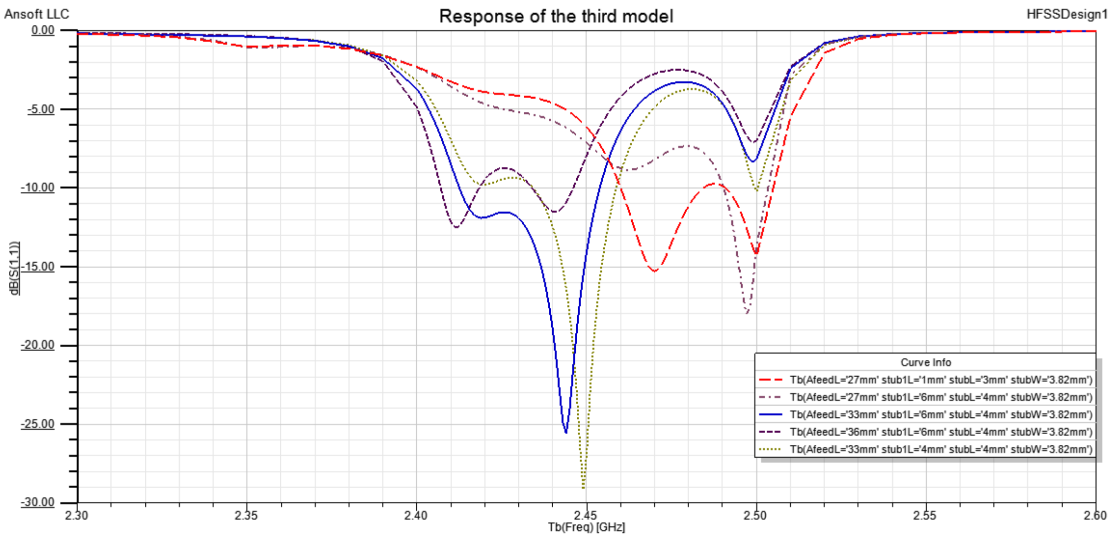

Figure (4.31): Response of the forth model of the filtering antenna.

The resulted return loss graphs are presented in Figure 4.30. They are promising. The bandwidth is 40MHz for *AfeedL*=33mm, *stubL*=4mm, *stub1L*=6mm and *stubW*=3.82mm, but the model still needs further improvement. In Future, stub widths may be changed separately and antenna feed line width also may be changed.

4.6 Comparison with previous researches:

Table 4.9 shows a summary of some of the previous researches on the filtering antenna presented in literature

Table (4-9): Previous researches of the filtering antenna

Paper №	Paper Title	Antenna type	Bandwidth	Return loss at 2.45GHz	Gain	№ of resonators	Type of the resonators
[4]	*New Printed Filtering Antenna with Selectivity Enhancement*	Printed meander line	~400MHz 2.3-2.7 GHz	15 dB	2.78 dB	1	Quarter wave resonator
[5]	*A New Compact Filtering Antenna Using Defected Ground Resonator*	Γ- shape	~400MHz 2.3-2.7 GHz	12 dB	---	1	Defect ground resonator
[6]	*A Compact Simple Structured Filtering Antenna Utilizing Filter Synthesis Technique*	Inverted L	~350MHz 2.3-2.7 GHz	12 dB	---	1	Quarter wave transmission line
[7]	*Synthesis and Design of a New Printed Filtering Antenna*	Inverted L	~150MHz At 2.4GHz band	10 dB	---	2	Parallel coupled microstrip line sections
[8]	*A Compact Printed Filtering Antenna Using a Ground-Intruded Coupled Line Resonator*	Γ- shape	~400MHz 2.3-2.7 GHz	15dB	1.2dB	1	Ground intruded coupled line resonator
[9]	*A Compact Filtering Microstrip Antenna With Quasi-Elliptic Broadside Antenna Gain Response*	U- shape	~200MHz at 5 GHz band	---	---	1	T- shape resonator
[10]	*Microstrip Patch Antenna Integration on a Bandpass Filter Topology*	Rectangular patch microstrip antenna	~50MHz at 2.4 GHz band	15dB	---	3	Hairpin resonator

4.7 Conclusion

In this Chapter a rectangular microstrip antenna is successfully designed then a hairpin microstrip filter is also successfully designed. Finally a filtering antenna is successfully designed and it achieves the focus of this thesis that is enhancing the antenna bandwidth using filtering antenna technology and designing a filtering antenna for ISM band.

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[9] C. K. Lin and S. J. Chung, "*A Compact Filtering Microstrip Antenna With Quasi-Elliptic Broadside Antenna Gain Response* ", IEEE Antennas and Wireless Propagation Letters, Vol. 10, pp 381-384, 2011.

[10] J. Verdu, J. P-Carrier, C. Collado, J.Mateu and A. Hueltes, *"Microstrip Patch Antenna Integration on a Bandpass Filter Topology"* Proceed of Mediterranean Microwave symposium, 2-5 sept. 2012, Istanbul, Turkey.

Chapter 5 : Conclusion and Future Work

This chapter will conclude this work and then recommend for future work.

5.1 Conclusion

In this work, filtering antenna has been successfully designed in three models. Filtering antenna is a combination between an antenna and a filter to work as a unit. The filter becomes a part of the antenna and the antenna will act as a new resonator in the filter. This enhances the bandwidth of the antenna and improves its frequency selectivity.

First, the antenna and the filter are designed separately. The antenna is a rectangular microstrip antenna that is fed with quarter-wave transformer and the filter is third order Chebyshev hairpin filter.

The antenna is designed first, by using the designing formulas, then the effects of the length and width of the substrate and the ground are studied to enhance the antenna performance itself. The minimum return loss of the antenna is -44.16dB and bandwidth is 56MHz.

The filter is designed next. The minimum return loss of the filter is 19.7dB and the bandwidth at the minimum return loss is 78MHz.

The filter and the antenna are then integrated to each other to build the first model which is the basic target of this work. This model reaches the target that increases the antenna bandwidth to about 145% and the antenna acts as a resonator in the filter that appears clearly from the forth return zero in the response.

Other models are built to make the filtering antenna more compact in size. They are basically built from the first model by removing the quarter-wave transformer and adding a direct transmission line connecting the antenna to the filter. Matching can then be achieved by adjusting the length and width of the transmission line or adding an open shunt stub. These models gave a promising results that need further work in the future.

5.2 Future Work

Based on the conclusions drawn and the limitations of the work presented, the following research aspects and issues would provide potential and natural progression to the accomplished works in this thesis:

- Fabricate and test the first model of filtering antenna and compare its results with the simulation results.
- Develop the other models of filtering antenna by doing further simulations to improve return loss.
- Work on other filtering antenna models to further reduce the size.