
Microstrip Patch Antenna

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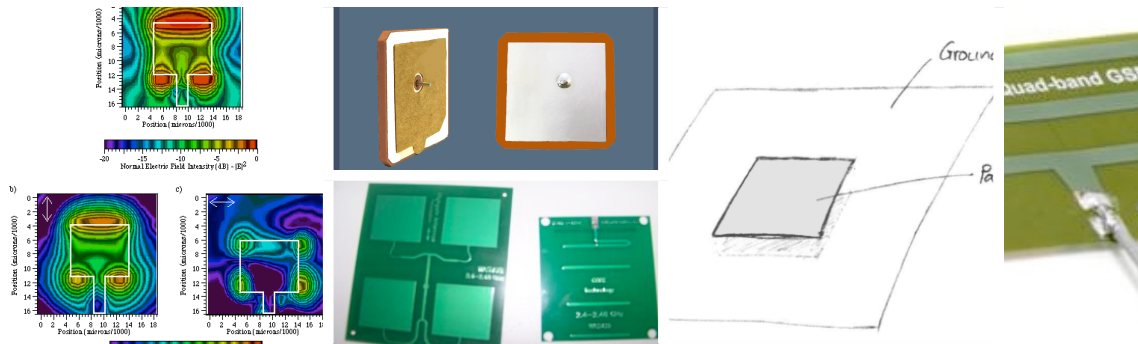
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Introduction



Different types of microstrip patch antenna's

Background

Although I grew up without knowing in which educational direction I might go, I have always been interested in gadgets and especially mobile phones. As I later became a student in Electrical Engineering I became more and more interested in wireless technology as I saw it had a big potential for the future. Growing up with fictional movies and dreams about what the future might bring, has also inspired me to think in terms of wireless technology. It has always been my plan to do something with wireless technology every since I became a student. Working for Nokia inspired me to connect the two dots together and I became interested in making a patch antenna.

I have previously worked on a pineapple antenna, which I constructed by assembling two cans after emptying the pineapple out of them. The same principle is widely known as the “Pringles Antenna” around the internet, after using a Pringles can.

Motivation

My motivation leaning about patch antennas is targeted through a future job, when I have gained more knowledge and experience besides this project.

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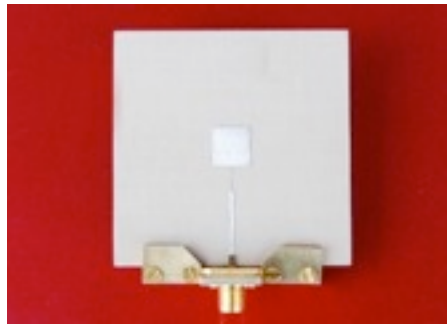
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Objectives

The objective of this project is as follows:

- To understand the basics behind antenna theory and microstrip patch antennas.
- A design with calculations of a patch antenna.
- Simulations of the design patch antenna.
- If there is time and if available a build and testing of a patch antenna design will be carried out.



Patch antenna.

A report has to be written and handed in by December 14, 2010 at 5.00 PM.

The total amount of time spent on this project has to be at least 135-140 hours as a 3 credit independent study.

1. WI-FI

As of today, Wi-Fi (also known as WLAN) has become a standard in most computers. Almost every modern mobile phone, and other gadgets are being implemented with Wi-Fi technology. Wi-Fi makes it possible for the user to connect to the internet or a LAN (Local Area Network) though a wireless connection (hence the name WLAN - Wireless Local Area Network).

Wi-Fi was introduced in the mid/late 1990's and has become very popular ever since, and it is still growing. The technology is still being developed to enhance faster speed, transfer rates and range of usage.

Wi-Fi uses the technical term "IEEE 802.11" and has standards in the names of "802.11 b/g/n". The term "802.11 b/g" is the most popular and that which is used today. "802.11 n" is becoming the new standard with faster speed, transfer rates and range of usage. Because 802.11 b/g is the mostly used today I have decided to make my patch antenna work for this standard.

1.1 Channels

802.11 b/g is in the 2.4GHz range and is divided into 14 channels, whereas only channels 1-11 is allowed for usage in the United States. Each Wi-Fi channel has a 22 MHz span. In this project my antenna will try and cover channels 1-13.

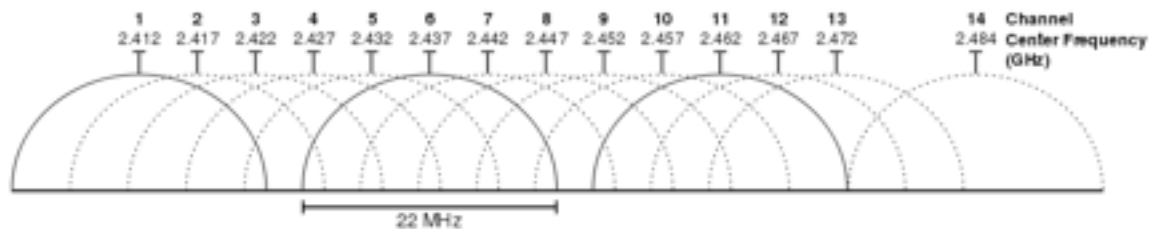


Figure 1.1: "802.11 b/g" channels.

As seen in figure 1.1 each channel is separated with 5MHz and thus create interference. Using only channels 1, 6 and 11 will avoid interference. 802.11 b/g has a bandwidth of 55 MHz. I will try and fit this bandwidth within the design parameters of my patch antenna.

2. Microstrip Antenna Theory

2.1 INTRODUCTION TO ANTENNAS

An antenna is a transducer that transmits or receives electromagnetic waves. By transmitting a signal into radio waves the antenna transforms electric current into electromagnetic waves and vice versa by receiving. Antennas are also said to radiate when transmitting. The IEEE definition of an antenna is given by the following phrase:

“That part of a transmitting or receiving system that is designed to radiate or receive electromagnetic waves”.

2.1.1 Antenna radiation

An antenna radiates by changing the flow of current inside a conduction wire. There are two ways to do this.

1. By time-varying (change of velocity, acceleration and/or de-acceleration) the current in a straight wire, the current will create a flow making the antenna radiate. If there is no motion of flow or if the flow of current is uniform, the straight wire will not radiate.
2. If we bend the wire, even with uniform velocity, the curve along the wire will create an acceleration in the current flow and the wire will therefore radiate.

If the charge is oscillating with time, the radiation will also occur even along a straight wire.

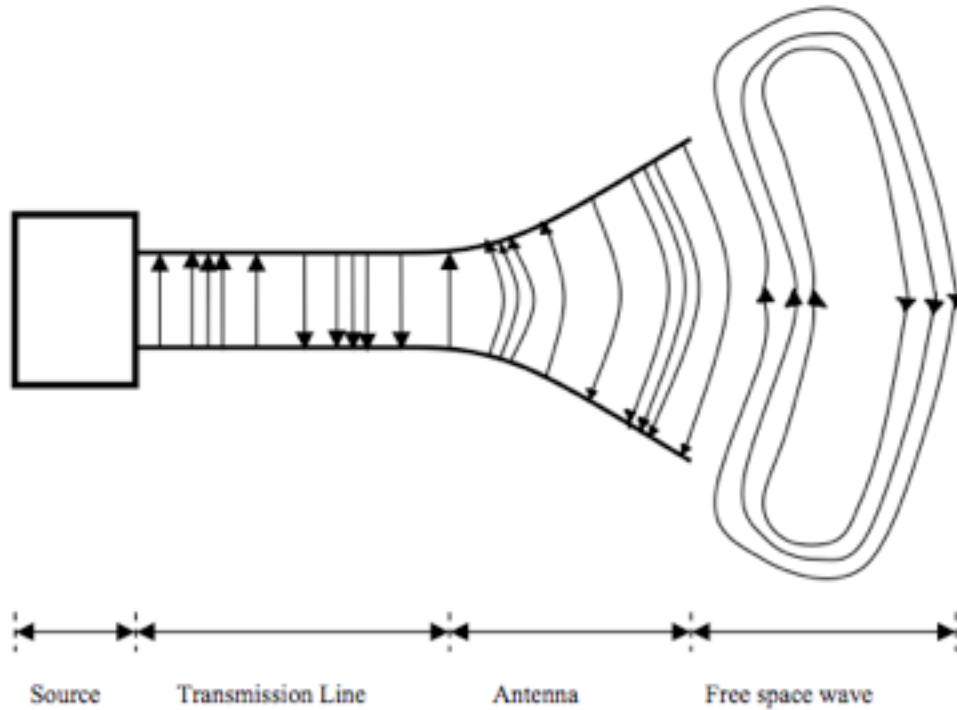


Figure 2.1: A radiating antenna.

By looking at figure 2.1, we see a radiating antenna. A sinusoidal voltage source is connected to a transmission line (transmission line is described in the next subsection), and is creating an electric field. When an electric field is created, electric force lines are created which are tangential to the electric fieldlines. The electric force lines force the free electrons to displace along the conduction wire and thereby create the flow in current. When the free electrons move, they create a magnetic field.

Because of the time varying electric and magnetic fields, electromagnetic waves are created. As these waves approach the open air, free space waves are formed, more commonly known as radio waves. When the electromagnetic waves are within the transmission line, the antenna is sustained due to the charges from the voltage source, but as soon as they enter the free space they create closed loop because of the nature of physics.

To transfer my oscillating flow of current from my voltage source to my antenna I need to connect my antenna to a transmission line. A transmission line is described next.

2.2 BASICS OF A TRANSMISSION LINE

2.2.1 Transmission Line

To connect my antenna to a certain utility (or application) I need a transmission line. In figure 2.2 a two-port transmission line is shown (earlier described as the two conducting wires).

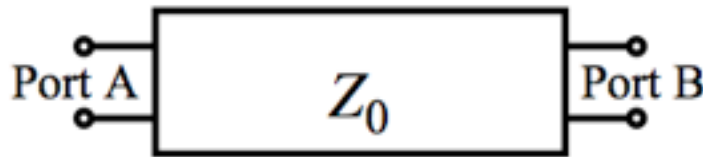


Figure 2.2: Two-port transmission Line

Transmission lines are measured in ohms and are described as Z_0 , the characteristic impedance. Z_0 varies in value depending on what transmission line is used. In a coaxial cable the transmission line is standardized to either 50Ω , which is used in most antenna applications, or 75Ω for use in satellite communication or very high frequency applications. Other transmission lines include as twisted or untwisted pair cables, standardized to 100Ω and 300Ω , for use in radio communication.

The purpose of a transmission line is to transfer power from one end (Port A) to another end (Port B) without any loss, as we want all generated power to radiate from the antenna into the free space. Z_0 is not ideal and we will therefore have a loss inside the transmission line in terms of impedance.

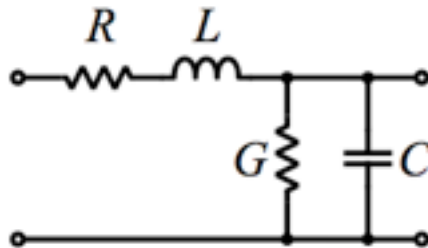


Figure 2.3: Transmission line elements

Figure 2.3 shows the elements of a transmission line, also called the “Telegrapher’s equation”, where:

R = transmission line resistance (Ohm).

L = transmission line inductance (Henry).

C = transmission line capacitance (Farad).

G = transmission line conductance (Siemens).

The transmission lines output impedance is calculated from the following formula:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$

As we can see from the equation above the characteristic impedance changes as frequency changes.

2.2.2 Matching and reflection

When sending power from the source to the load, I want as much power to be absorbed by the load (in this case the load is my antenna), and as little power to be reflected back to my source. This can be ensured by setting $Z_L = Z_0$, which says that the transmission line is matched.

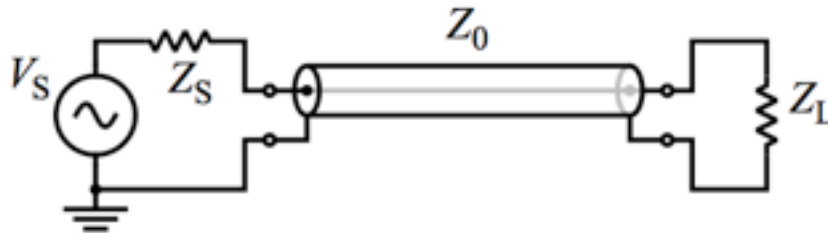


Figure 2.4: Transmission line between source and load.

As stated earlier, the characteristic impedance Z_0 changes when change in frequency occurs. We therefore need to match our load impedance to the operating frequency of the antenna. When the load impedance is not 100% matched to Z_0 a mismatch occurs. This means that some of the power in my forward wave will be reflected back to the source and will be lost in terms of heat. If

the system is totally mismatched the whole power could be reflected back and eventually end up damaging the source. It is therefore very important to match your system correctly.

To see how much is reflected back to the source I can calculate the reflection coefficient Γ .

$$\Gamma = \frac{Z_L - Z_S}{Z_L + Z_S}$$

The reflection coefficient is a complex number ($a+jb$). If the imaginary part is 0, then if:

$\Gamma = -1$, the line is short-circuited (maximum negative reflection (phase shift of 180° (or Π)).

$\Gamma = 0$, the line is perfectly matched (no reflection).

$\Gamma = 1$, the line is open-circuited (maximum positive reflection).

Because the reflection coefficient is complex, it changes with frequency. Instead of calculating the reflection coefficient a number of times a Smith chart is used for a graphical expression of the reflection coefficient.

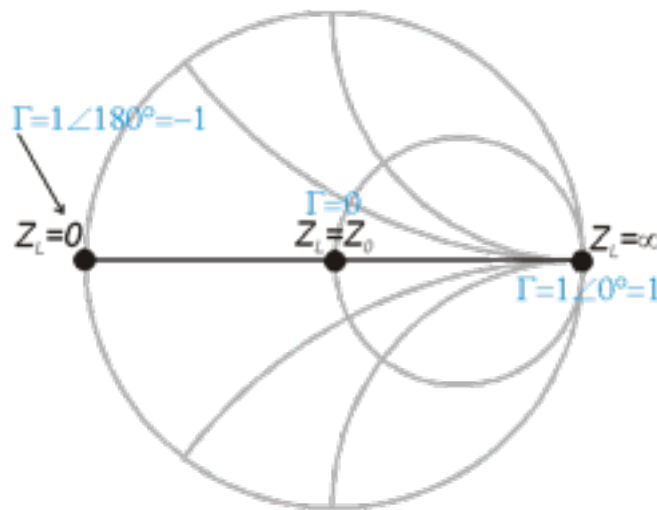


Figure 2.5: Smith Chart

Usually the impedance of the load and source do not match. To ensure a matched system an additional circuit of lumped components is required. The circuitry will not be described in this project.

2.2.3 VSWR

Another way to see how much my system is matched, VSWR (Voltage Standing Wave Ratio) can be used. VSWR is the ratio between the maximum voltage and minimum voltage in the transmission line, and can be defined as follows:

$$VSWR = \frac{1 + \rho}{1 - \rho}, \text{ where } \rho \text{ is the magnitude of } |\Gamma|.$$

When the system is matched the reflection coefficient approaches 0, while VSWR approaches to 1.

2.2.4 Return loss

Return loss is the power of the reflected signal in a transmission line. It can be calculated by the following equation and is given in dB.

$$RL_{dB} = -20 \log_{10} |\Gamma|$$

The return loss is also stated as the S_{11} of the S-parameters.

2.2.5 S-parameters

Transmission lines have S-parameters, also called “Scattering parameters” which refer to RF’s voltage out versus voltage in and are measured in dB. S-parameters are a complex number but they mostly only refer to the magnitude as you want to know how much loss or gain you get.

In figure 2.6 and table 2.7, S-parameters are described.

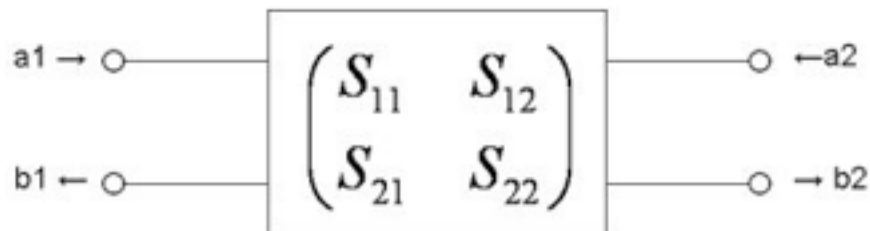


Figure 2.6: S-parameters in a two-port network.

S parameter	Description
S_{11}	The input ports voltage reflection coefficient.
S_{12}	The reverse voltage gain.
S_{21}	The forward voltage gain.
S_{22}	The output ports voltage reflection coefficient.

Table 2.7: S-parameters

When an RF signal enters a port, some fraction of that signal bounces back out of that port. Some of it “scatters” and exits other ports and might even be amplified. Some of it disappears as heat or even electromagnetic radiation. S-parameters can be either be calculated or measured on a network analyzer.

2.3 ANTENNA INPUT IMPEDANCE

Just like the transmission line, the impedance of an antenna consists of real and imaginary parts.

The following equation can be used to define the input impedance of an antenna:

$$Z_A = R_A + jX_A$$

where,

Z_A is the antenna impedance at terminals.

R_A is the antenna resistance at terminals.

X_A is the antenna reactance at terminals.

R_A of the impedance of an antenna can be divided into radiation and loss resistances.

$$R_A = R_r + R_L$$

Where,

R_r is the radiation resistance.

R_L is the loss resistance.

Figure 2.8 shows an equivalent circuit of an antenna.

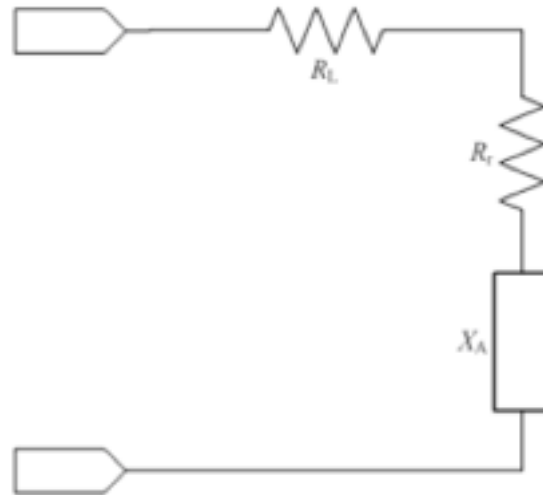


Figure 2.8: Antenna equivalent circuit.

The imaginary part represents the power stored in the near field of the antenna.

The radiation resistance is a parameter determined by the shape, size, and type of the antenna, but not by the material. The power associated with the radiation resistance is the power actually radiated by the antenna while the power dissipated in the loss resistance is lost as heat in the antenna itself due to dielectric or conducting losses. The imaginary part is causing the radiator to act similarly to lumped components, such as coils and capacitors.

2.4 FIELD REGIONS

When radiating power from the antenna, the radiated wave crosses the near field and the far field. On figure 2.9 we see an illusion of the near field and the far field.

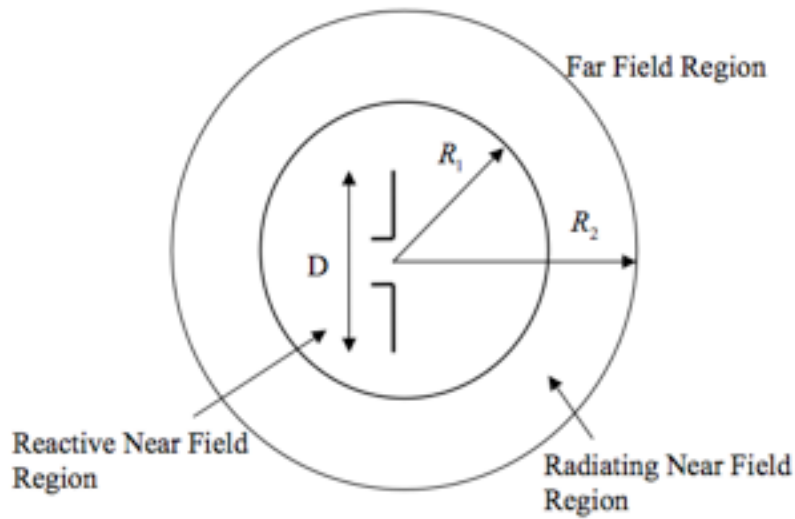


Figure 2.9: Near and Far Field regions.

The radius of the near field is said to be 1 wavelength of the radiated wave. Figure 2.10 shows a wavelength.

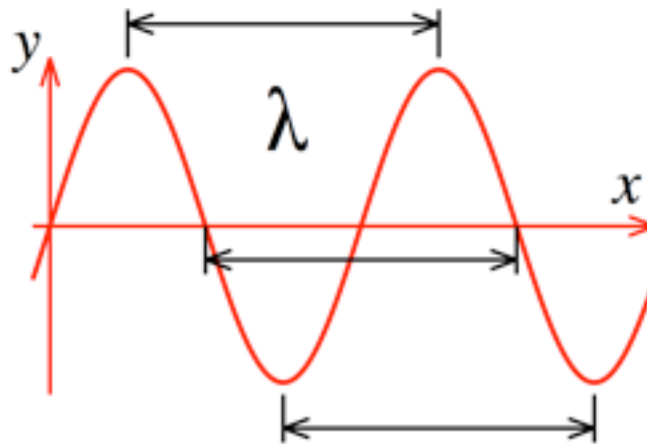


Figure 2.10: Wavelength of a sine wave.

A wavelength is measured in meters by dividing speed of light over the operating frequency.

2.4.1 Near field

The near field is divided into two fields: the reactive near field and the radiative near field. The reactive near field region covers from $R_1 = 0$ to $R_1 = \lambda / 2\pi$ or $0.159 \times \lambda$, while the radiative near field covers the remainder from $R_2 = \lambda / 2\pi$ to $R_2 = \lambda$.

To understand how far the radius is, I will calculate the radiative near field radius as I know the operating frequency of my WiFi antenna:

$$\lambda = \frac{c}{f_c} = \frac{3 \cdot 10^8 \text{ m / s}}{2.45 \text{ GHz}} = 122.449 \text{ mm} \approx 122 \text{ mm}$$

It is obvious to understand that I cannot place a computer 0.2 m away from the base station. Therefore it is important to understand the far field.

2.4.2 Far field

The far field is said to start from 2λ to infinity. What happens between λ and 2λ is called the transition zone. The transition zone has parameters of both the near field and the far field. Here follows the calculation for 2λ .

$$2\lambda = 2 \frac{c}{f_c} = 2 \frac{3 \cdot 10^8 \text{ m / s}}{2.45 \text{ GHz}} = 244.898 \text{ mm} \approx 245 \text{ mm}$$

The far field generally fall off in amplitude by $1/r$. As the sphere is proportional to r^2 the amplitude is therefore the total energy per unit area is proportional to $1 / r^2$ in free space. Free space means without interference from other signals or attenuation from the surroundings of the world.

It is worthy of note that the sphere radiation is a rough calculation as you typically have different radiation pattern, based on the type of antenna used. Therefore the sphere is often used as a “rule of thumb”.

2.5 BANDWIDTH

The bandwidth of an antenna can be defined as the range of usable frequencies within the performance of the antenna around the resonance frequency. Usually only the impedance bandwidth is specified, but there are other important bandwidth definitions as well such as: polarization bandwidth, directivity bandwidth, and effectivity bandwidth, where the last two, if often combined, are defined as gain bandwidth.

Impedance bandwidth (also referred to as return loss bandwidth) already depends on a large number of parameters such as the dielectric and size of the ground plane, both of which alter the Q factor. The type of feed structure of the antenna affects the bandwidth as well.

Because the impedance bandwidth consists of many parameters, it is too complex to make a design calculation from a desired bandwidth point of view. Therefore these are often measured within the return loss of -6 dB.

A rough equation for free space (without dielectric) can be made by the following equation:

$$\frac{\delta f}{f_{res}} = \frac{Z_0}{2R_{rad}} \frac{d}{W}$$

Where,

d = The height of the patch above the ground plane.

W = The width of the patch (typically a half – wavelength).

Z_0 = The impedance of free space ($\approx 377\Omega$).

R_{rad} = The radiation resistance of the antenna.

As discussed earlier in this chapter, one method of judging how efficiently an antenna is operating over the required range of frequencies is by measuring its VSWR. This can then be compared with the return loss to find the bandwidth of the antenna.

A desired bandwidth can be calculated by the following equation.

$$BW_{Broadband} = \frac{f_H}{f_L}$$
$$BW_{narrowband}(\%) = \left(\frac{f_H - f_L}{f_C} \right) \cdot 100$$

Where,

f_H = The Upper Frequency,

f_L = The Lower Frequency,

f_C = The Center Frequency.

An antenna is said to be broadband if it is 2 or higher.

As I already know my operating frequencies I can therefore calculate my desired bandwidth and compare it the measured result to see if the bandwidth in the antenna is being obtained.

$$BW_{Broadband} = \frac{f_H}{f_L} = \frac{2.472GHz}{2.412GHz} \approx 1.02$$

By calculating the broadband equation I can already assume that the bandwidth for my antenna is a narrowband as the broadband bandwidth is less than 2.

$$BW_{narrowband}(\%) = \left(\frac{f_H - f_L}{f_C} \right) \cdot 100 = \left(\frac{2.472GHz - 2.412GHz}{2.45GHz} \right) \cdot 100\% \approx 2.45\%$$

The antenna bandwidth can also be determined by using a definition of the antenna beamwidth.

2.6 RADIATION PATTERN

2.6.1 Beamwidth

The beamwidth is the angle between the half-power (-3dB) of the peak effective radiated power. It is usually expressed in degrees and is shown in the horizontal plane by being displayed in a polar diagram. Figure 2.11 shows a beamwidth in a polar diagram.

By a simpler description it can be said that the received signal can be reached within a beamwidth-margin of the pointed direction. It can be calculated by the following equation.

$$\theta_{3dB} = \cos^{-1} \left(\sin(\theta_0) - 0.443 \frac{\lambda_0}{l} \right) - \cos^{-1} \left(\sin(\theta_0) + 0.443 \frac{\lambda_0}{l} \right)$$

Where,

θ_0 = The pointing direction.

λ_0 = The free space wavelength.

l = The total length of the patch.

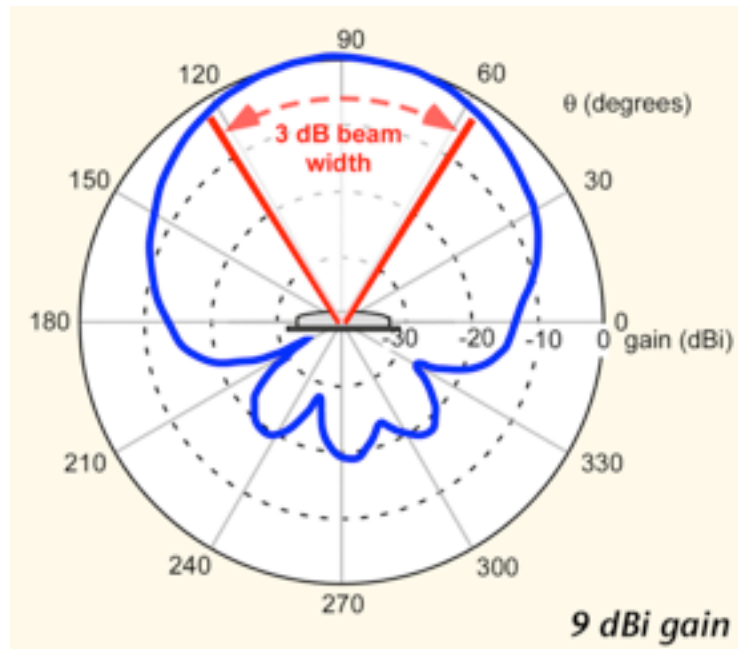


Figure 2.11: A polar diagram showing beamwidth.

2.6 MICROSTRIP ANTENNA

A microstrip antenna consists of a metallic pattern on one side of a dielectric substrate and ground plane on the other side of the substrate. In this project I have focused on making a microstrip patch antenna. Figure 2.12 shows a microstrip patch on a dielectric substrate.

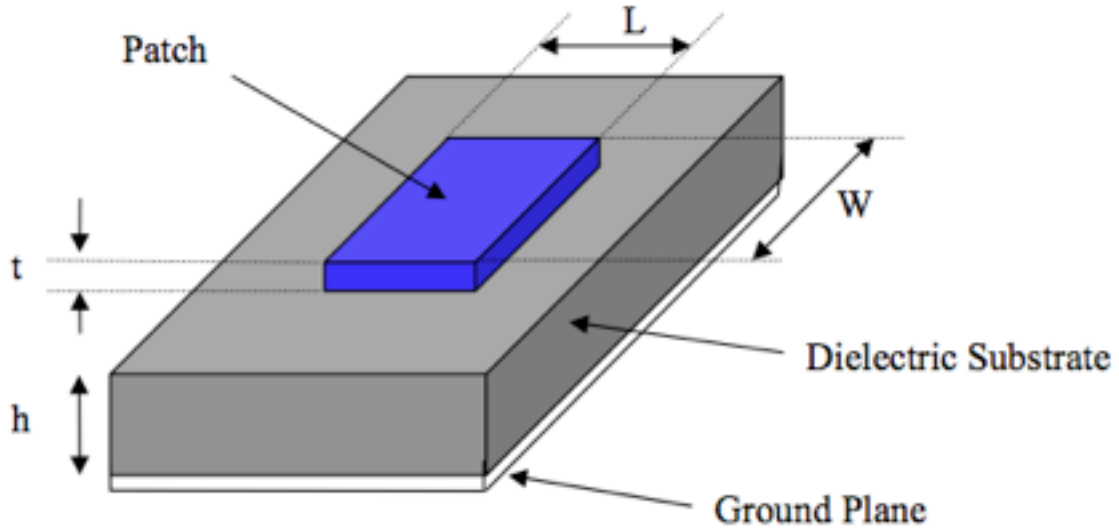


Figure 2.12: Structure of a microstrip patch antenna.

The antenna patch can have different shapes, but is most likely rectangular. In order to make performance predictions the rectangular patch antenna has the following parameters, where λ_0 is the wavelength in vacuum also called the free-space wavelength.

$$\text{Length}(L) : 0.3333\lambda_0 < L < 0.5\lambda_0$$

$$\text{Height}(h) : 0.003\lambda_0 \leq h \leq 0.05\lambda_0$$

$$\text{Thickness}(t) : t \ll \lambda_0$$

$$\text{Dielectric constant}(\epsilon_r) : 2.2 \leq \epsilon_r \leq 12$$

In electromagnetic radiation λ is often given instead of λ_0 as the speed of light in vacuum is very close to the speed of light in air.

As described later in this chapter the length of the patch is very important when it comes to the radiation. Looking at the parameters of the length, the length is slightly less than $\lambda/2$. That is because the microstrip patch antenna is constructed on the theory based on one-half wavelength.

2.6.1 Advantages and disadvantages

Microstrip antennas are becoming more and more popular every day. And with a more modern world where the internet and WiFi are delivered in many stores, more and more gadgets are using microstrip antennas. Some of the advantages are:

- Light weight.
- Low volume.
- Easy integration with Microwave Integrated Circuit's (MIC).

On the other hand, microstrip antennas also features some disadvantages compared to conventional antennas:

- Narrow bandwidth
- Low efficiency
- Low gain
- Extra radiation from feeds and junction
- Surface waves
- Low power handling capacity.

2.6.2 Q factor (Quality factor)

Microstrip antennas have a very high antenna Q (quality factor). Q represents the losses in the antenna, where a large Q leads to narrow bandwidth and low efficiency. Q can be reduced in the antenna by increasing the dielectric substrate thickness, but this will cause less power delivered from the source because of power loss in the dielectric substrate and making surface waves, as the power is scattered by the dielectric bends.

2.7 ANTENNA FEED

There are different ways to feed the antenna. It is obvious that one cannot have merely a patch and transmit power through it without actually delivering the power to the patch and vice versa by receiving.

I have been looking at two different feed methods for the patch antenna. These follow in figure 2.13 and figure 2.14.

2.7.1 Microstrip feed line

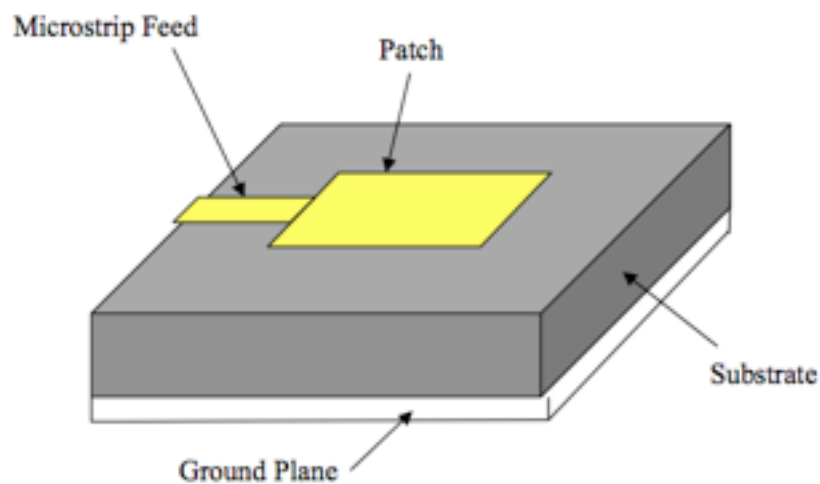


Figure 2.13: Microstrip feed line to the antenna.

Figure 2.13 shows the feed line method to the antenna. The feed line to the patch antenna is in its origin a transmission line and is therefore often referred to as the transmission line feed.

The feed line width is smaller than the patch and is etched directly to the edge of the patch so that power is transferred from the source through a coaxial cable, into the feed line and then to the patch. The purpose of the feed line is to match the impedance from the patch without any additional matching component, however because the feed line is a patch itself it can cause radiation interfering with the patch which will decrease the bandwidth of the antenna.

When designing the feed line, this must be along the side of the length, as the current flow is along the direction of the feed wire and at the length is where the maximum radiation of the patch is created.

2.7.2 Coaxial feed

Another way to feed the antenna is to directly connect it to a coaxial cable to avoid radiation from the feed line. Figure 2.14 shows a coaxial connected to the patch.

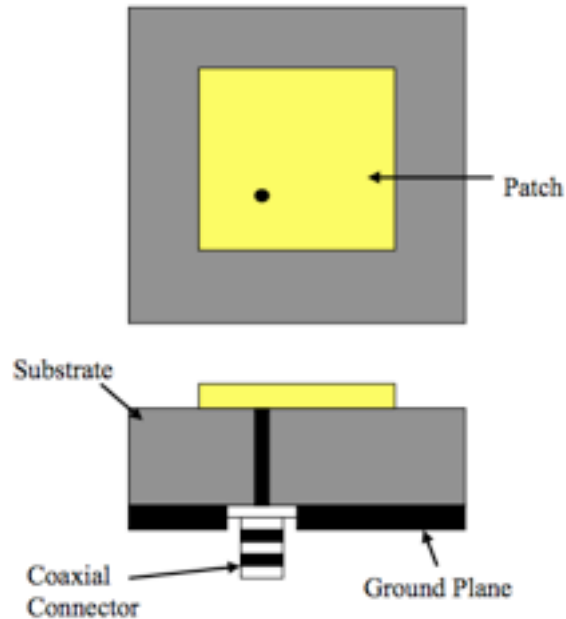


Figure 2.14: Coaxial feed to the antenna.

The coaxial feed (also referred to as a probe feed) is a very common way to feed the antenna. As seen from figure 2.14 the coaxial connector is inserted at the ground plane where the inner conductor extends through the dielectric substrate and is soldered on to the patch.

The advantages of this kind of feed are it's lower radiation and the fact that it can be placed anywhere to match the impedance of the patch (though a calculation of the x,y plan of the patch). The disadvantage is that it provides a narrow bandwidth and is it is difficult to drill though the substrate. With a fragile dielectric substrate you might end up damaging the substrate or the inner connector might not perfectly fit the drilled hole and thereby create power loss to air conductivity. Another problem is that with a thick substrate the inner coaxial conductor has to be longer leading to higher input inductivity which creates matching problems.

I chose to use the feed line method as this is the most popular and as this seems more straightforward when seen from a simulation perspective. Next will follow a description on how to define the size of the patch and the line feed dimensions.

2.8 TRANSMISSION LINE MODEL (MICROSTRIP FEED LINE)

Figure 2.15 shows a microstrip patch antenna with a microstrip feed line connection on a dielectric substrate.

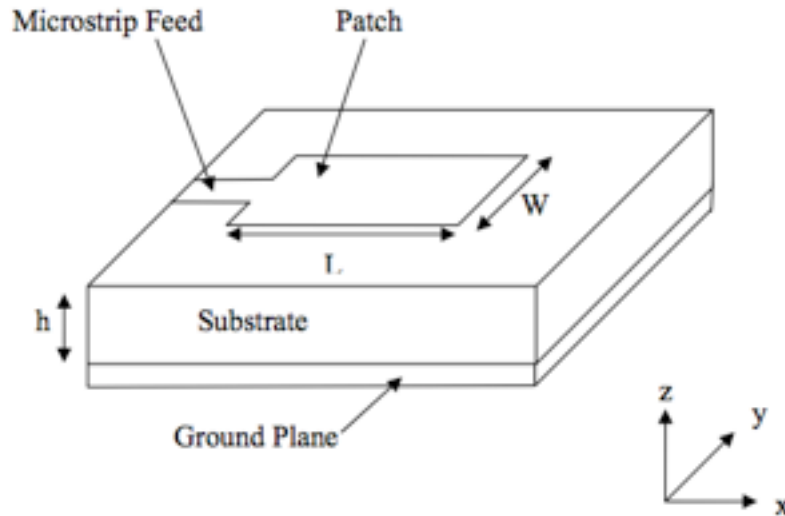


Figure 2.15: Microstrip line feed model.

In order for the antenna to operate under the fundamental TM_{10} mode, the length of the patch must be slightly less than $\lambda/2$ where λ is defined as:

$$\lambda = \frac{\lambda_0}{\sqrt{\epsilon_{\text{reff}}}}$$

“A transverse mode of a beam of electromagnetic radiation is a particular electromagnetic field pattern of radiation measured in a plane perpendicular (i.e. transverse) to the propagation direction of the beam.”¹

TM_{10} mode implies that the field varies one $\lambda/2$ cycle along the length, and there is no variation along the width of the patch. TM stands for Transverse Magnetic which means it has no magnetic field in the direction of propagation. There are different modes, where '10' or '1,0' is the dominant mode.

¹ http://en.wikipedia.org/wiki/Transverse_mode

As seen from figure 2.12 the conducting transmission line has a height 't'. Because of this the transmission line cannot support transverse-electric-magnetic (TEM). TEM means a direct transfer of the electric fieldlines to the dielectric. As seen on figure 2.16, this cannot be supported because some of the electric fieldlines are going into the air before entering the dielectric substrate.

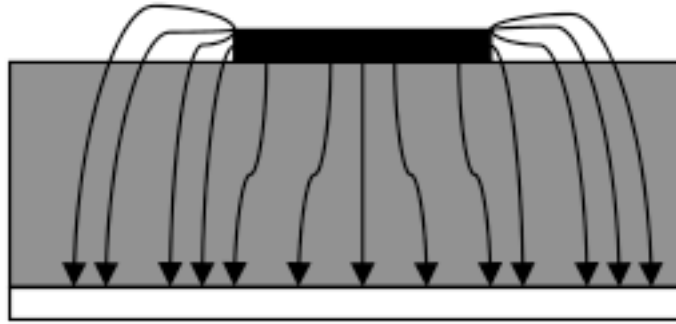


Figure 2.16: Electric field lines in a transmission line.

Because the electric field lines are moving into the air before entering the dielectric substrate the ϵ_r will be replaced by ϵ_{eff} which is slightly less than ϵ_r .

The equation of ϵ_{eff} is given by:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \cdot \left[1 + 12 \cdot \frac{h}{W} \right]^{-\frac{1}{2}}$$

Where,

ϵ_r = The dielectric constant of the substrate.

h = The height of the the dielectric substrate.

W = The width of the patch.

Because the electric field line move through the air, the length of the patch is extended on both sides. Figure 2.17 (a) shows the two radiating slots along the length of the patch.

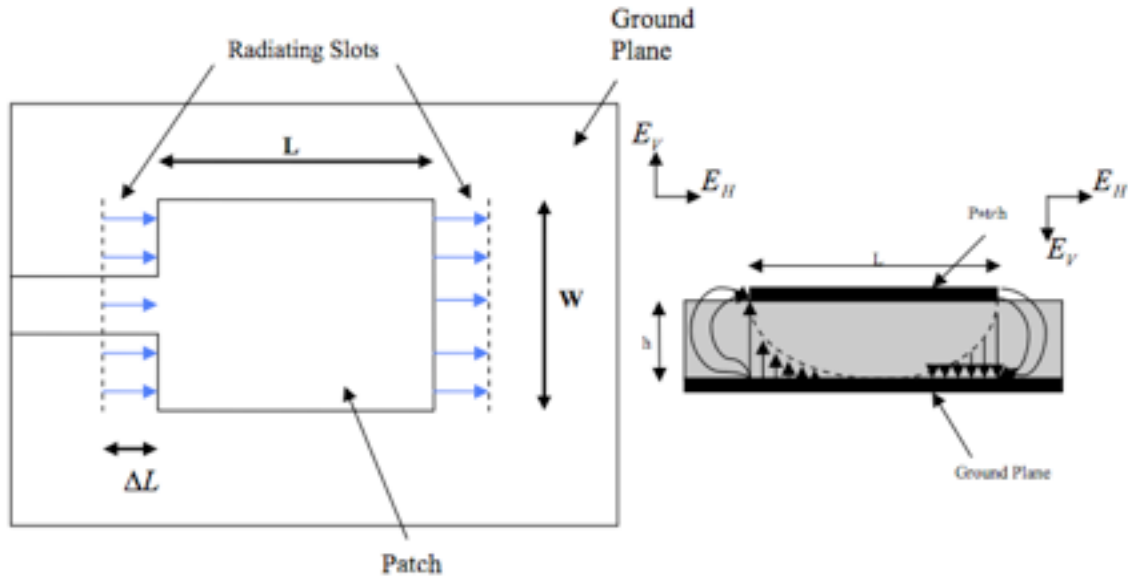


Figure 2.17: (a) Top view of the antenna. (b) Side view of the antenna.

From figure 2.17 (a) it can be seen that the patch antenna is represented by two slots (one in each end), separated by a length L . Both ends are open circuited. Along the width the voltage is at a maximum and the current is at a minimum due to the open ends.

Looking at figure 2.17 (b) the electric field lines at the two edges of the width are in opposite directions (E_V). They are out of phase and thereby cancel each other out. The two components which are in phase (E_H), give the maximum radiated field by combining the resulting fields. It is said that the radiation is produced by these two radiation slots.

The extended length ΔL can now be calculated by the following equation:

$$\Delta L = 0.412 \cdot h \cdot \frac{(\epsilon_{reff} + 0.3) \left(\frac{W}{h} + 0.264 \right)}{(\epsilon_{reff} - 0.258) \left(\frac{W}{h} + 0.8 \right)}$$

Thus the width is extended too, there are no reason in calculating the extension because the electric fields cancel each other out.

2.8.1 Width

The width W can be calculated by the following equation:

$$W = \frac{c}{2 \cdot f_c \cdot \sqrt{\frac{\epsilon_r + 1}{2}}}$$

Where,

c = The speed of light.

f_c = the resonance frequency

ϵ_r = the dielectric constant of the substrate.

2.8.2 Length

The effective length L_{eff} can be calculated by the following equation:

$$L_{eff} = \frac{c}{2 \cdot f_c \cdot \sqrt{\epsilon_{reff}}}$$

I can then calculate the actual length of my patch by the following equation:

$$L = L_{eff} - 2 \cdot \Delta L$$

2.8.3 Ground planes

Essentially the transmission line model is applicable to an infinite ground plane only. However, it has been shown that a finite ground plane can be used for if the ground plane is 6 times larger than the height of the dielectric substrate plus the used length or width. The ground plane can now be calculated as:

$$W_g = 6 \cdot h + W$$

$$L_g = 6 \cdot h + L$$

3. PCB substrate

3.1 INTRODUCTION

In today's market there are a lot of different PCB substrate products. Unfortunately there is not only one product that can cover all applications. It all depends on the application itself. Even though your application is simple it is still difficult to meet all requirements. Depending on your application you need to consider your selection of substrate. Properties to be considered includes dielectric constant, loss tangent, and their variation with temperature, frequency, dimensions, stability, thickness, resistance to chemicals, flexibility etc.

“RF applications are characterized by the need for low dielectric losses, low leakage, a need for a low and uniform dielectric constant accompanied by a low layer count. Further, since this type of PCB tends to be small, cost of the dielectric material has less effect on overall product cost than other cost components. As a result, using more expensive materials to meet performance goals is acceptable. Choosing a material based on its dielectric constant characteristics and losses usually dominates over other considerations.”²

In this chapter I will describe an overview of the most important properties I have looked for throughout this project. Furthermore will I describe a short introduction to the most common types of substrate and choose my selected substrate for this project.

As the project focuses on designing a WIFI antenna (close to High Frequency) I will mainly focus on substrate properties based for this project.

² <http://www.speedingedge.com/PDF-Files/tutorial.pdf>

3.2 SUBSTRATE PROPERTIES

3.2.1 Dielectric constant

The dielectric constant (also called “ ϵ_r ”, “DK” or “relative static permittivity”) is the ratio between the stored amount of electrical energy in a material and to that stored by a vacuum (which is by definition 1). It is also a measure of the degree to which an electromagnetic wave is slowed down as it travels through the insulating material. Dielectrics are i.e. used in capacitors to store more electrical charge than vacuum.

The lower the dielectric constant is, the better the material works as an insulator. The better an insulator, the better it resists electrons from being absorbed in the dielectric material, creating less loss. In table 3.1 I have made a selection of ϵ_r values in some materials.

Material	ϵ_r
Vacuum	1
Air	1.00059
PTFE (teflon)	2.1
Paper	3.5
Rubber	7
Silicon	11.68
Water	80.1 (20 °C)
Metal	∞ (Infinite)

Table 3.1: room temperature @ < 1 kHz³.

It is important to know that the dielectric constant varies with both temperature and frequency (usually goes down as frequency goes up) depending on the material. It is therefore a good idea to look in the datasheet of the selected substrate.

³ http://en.wikipedia.org/wiki/Relative_permittivity

3.2.2 Loss tangent

The loss tangent (also called “ $\tan \delta$ ”, “DF”, “low loss”, “dissipation factor”) is a measure of how much of the electromagnetic field traveling through a dielectric is absorbed or lost in the dielectric, usually through heat.

As the dielectric in a substrate is similar to the dielectric in a capacitor, the loss tangent can best be described as the loss through an equivalent series resistor (ESR) inside a capacitor. A small ESR describes a good capacitor with low loss.

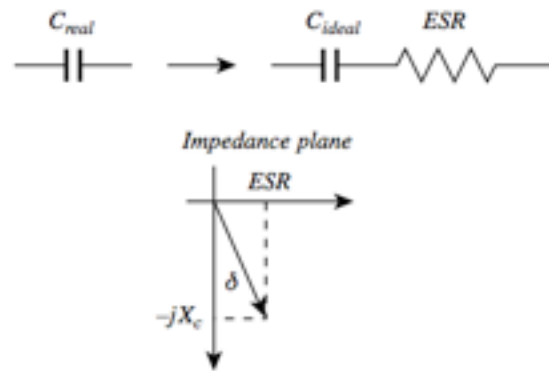


Figure 3.2: Loss tangent.

The loss tangent is often thought of as power loss and might have a low number at first glance (like 0.01 @ 1 GHz for FR-4 substrate). This might not look like much but if you power out 1000W on an antenna and the loss tangent is 0.01, then you roughly have a power loss of 10 W. The loss tangent also varies with frequency and temperature depending on which material is used. The higher loss tangent, the higher loss you will have in the dielectric, which leads to reduced antenna efficiency.

The loss tangent is mostly of a concern in high speed digital signal design (i.e. Ethernet link @ 2.4 Gigabit per second usage). As velocity of signals increases as the frequency goes up, the loss will be higher, because of higher frequencies.

3.2.3 Thickness

The thinner the substrate is, the less loss, but the less power you can send through it, because the transmission line has to be thinner to keep the same impedance.

With a thicker substrate you need a wider strip line to keep the same impedance. This will give a higher Q in the copper = more power through it, while the disadvantage will be more weight and higher radiated power from the transmission line which we want as little as possible.

As an alternative you could use a substrate with a lower dielectric constant, because that way you can increase the microstrip line (transmission line) = higher Q, without increasing the thickness of the substrate.

The thickness also depends on your application. For instance: a designer wants to design a thin mobile phone because that attracts certain customers. This means that there is use for a thinner substrate and you therefore have to cater to it.

3.2.3.1 Short Summary

Let's say I want the most optimal substrate for my microstrip patch antenna @ 2.45GHz and my dielectric constant is fixed, then:

- Increasing track width, decreases Z_0 .
- Increasing substrate height, increases Z_0 .
- Increasing copper clad, decreases Z_0 (although it has small effect on Z_0).

It is important to know that you should select a substrate out from your application, as you can adapt your needs by selecting a different substrate. There isn't any super substrate that works perfectly with all applications as they all have different properties.

3.2.5 Dielectric strength

The dielectric strength (also called “Electric strength”) is how much potential (voltage) the dielectric material can resist before it makes a dielectric breakdown. A dielectric breakdown is when the dielectric material is damaged and could mean that the material doesn’t work as an insulator anymore, which will lead to short circuiting. This property will only be useful in high power/high voltage applications. Here follows some critical properties which should be considered for power/high voltage applications:

- The dielectric strength increases if the material thickness increases.
- The dielectric strength decreases if the frequency increases.
- The dielectric strength decreases if the temperature increases.
- The dielectric strength decreases if the humidity increases.

Dielectric voltage breakdown is not really an issue as most laminates can withstand high power voltage such as 20 kV/mm.

3.3 SOME COMMON SUBSTRATES

Material	T _g	ϵ_r^*	Tan (δ)	DBV (V/mil)	WA, %
Standard FR-4 Epoxy Glass	125C	4.1	0.02	1100	0.14
Multifunctional FR-4	145C	4.1	0.022	1050	0.13
Tetra Functional FR-4	150C	4.1	0.022	1050	0.13
Nelco N4000-6	170C	4	0.012	1300	0.10
GETEK	180C	3.9	0.008	1100	0.12
BT Epoxy Glass	185C	4.1	0.023	1350	0.20
Cyanate Ester	245C	3.8	0.005	800	0.70
Polyimide Glass	285C	4.1	0.015	1200	0.43
Teflon	N/A	2.2	0.0002	450	0.01
		* Measured with a TDR using velocity method. Resin content 55%			

T_g = glass transition temperature
 ϵ_r = relative dielectric constant
 Tan (δ) = loss tangent

DBV = dielectric breakdown voltage
 WA = water absorption

All materials with woven glass reinforcement except teflon.

Table 3.3: Values for some common substrates.

3.4 SUBSTRATE MATERIALS

3.4.1 Ceramic substrate

The ceramic substrate is mainly used in small size applications with frequencies below 1 GHz. It has low loss tangent and has good chemical resistance, but is also very expensive. Besides that, ceramic is very hard to produce and handle. For instance it is very hard to drill holes in the substrate without breaking it. Some ceramic material has a high dielectric constant which is used where you need an important size reduction.

3.4.2 Synthetic substrate

Synthetic substrate is commonly made out of organic material like PTFE (also known as Teflon). These materials possess low loss tangent and low ϵ_r . The only problem is that this material is very soft and can therefore easily change the characteristics of a microstrip antenna if it is not handled well enough.

3.4.3 Composite material substrate

Composite material is made out of mixed chemicals between fiberglass, ceramic or quartz and synthetic material. There is a wide variety of composite material on the market which has been modified so they fit both to antenna fabrication and standard PCB design.

3.4.4 Low-cost low-loss substrate

Ceramic-, synthetic and Composite material substrate is usually used where other applications are needed or a microstrip antenna needs perfection and also it is too expensive to use in consumer electronics such as TV's, mobile phones, etc.

3.4.5 FR-4

FR-4 substrate is a very common and by far the most used substrate in consumer electronics market as it has a good quality-to-price ratio. It is mostly used where cost is more efficient than performance.

FR-4 is a standard with many different distributors making many different FR-4 quality and property boards. It is made of woven fiberglass with an epoxy resin binder (binds the copper clad to the dielectric substrate) that is flame resistant. The dielectric constant goes down the more the FR-4 PCB is reinforced with epoxy resin instead of fiberglass as this is not determined as a standardized parameter. 100% epoxy resin boards has a dielectric constant of 3.4 @ 1MHz.



Figure 3.4: Copper clad FR-4 PCB substrate.

The FR-4 changes its dielectric constant along its area which makes it too unstable to mass produce precise antennas on it. Also, the FR-4 has a higher loss at frequencies over 3GHz, because of the sensitivity of the cheap substrate. Other products are therefore recommended to perform better than FR-4 in RF applications. A highly recommended distributor is Rogers, who is a little more expensive but performs much better in RF applications.

In the cellphone industry, companies use higher quality FR-4 substrate because it is more cost efficient, but from only one manufacture so they can be sure of the quality and properties when mass producing. The performance is typically around -13 dBm.

3.4.6 Rogers

Rogers Corporation is a company that specializes in high frequency PCB's and make some of the best on the market in the low-cost low-loss substrate area. They match the loss tangent to other high end (not low-loss low-cost) substrate material products. They usually have a slightly higher dielectric constant but also cost around 3-4 times less than the other substrate products.

Rogers also makes low cost types with higher dielectric constants so they can be used for other applications such as high power.

In table 3.5 we see the datasheet for Rogers RO4003 substrate. The dielectric constant is 3.38 @ 10 GHz. This means that the dielectric constant will slightly be higher around 2.4 GHz which is the frequency domain for this project. The loss tangent is 0.0021 @ 2.5GHz which is good.

Property	Typical Value		Direction	Units	Condition	Test Method
	RO4003™	RO4350B™				
Dielectric Constant, ϵ_r (Process specification)	3.38 ± 0.05	3.48 ± 0.05	Z	--	10 GHz/23°C	IPC-TM-650 2.5.5.5 *Clamped Stripline
⁽¹⁾ Dielectric Constant, ϵ_r (Recommended for use in circuit design)	3.55	3.66	Z	--	FSR/23°C	IPC-TM-650 2.5.5.6 Full Sheet Resonance
Dissipation Factor tan, δ	0.0027 0.0021	0.0037 0.0031	Z	--	10 GHz/23°C 2.5 GHz/23°C	IPC-TM-650 2.5.5.5
Thermal Coefficient of ϵ_r	+40	+50	Z	ppm/°C	-100°C to 250°C	IPC-TM-650 2.5.5.5
Volume Resistivity	1.7 X 10 ¹⁰	1.2 X 10 ¹⁰		MΩ•cm	COND A	IPC-TM-650 2.5.17.1
Surface Resistivity	4.2 X 10 ⁹	5.7 X 10 ⁹		MΩ	COND A	IPC-TM-650 2.5.17.1
Electrical Strength	31.2 (780)	31.2 (780)	Z	KV/mm (V/mil)	0.51mm (0.020")	IPC-TM-650 2.5.6.2
Tensile Modulus	26,889 (3900)	11,473 (1664)	Y	MPa (kpsi)	RT	ASTM D638
Tensile Strength	141 (20.4)	175 (25.4)	Y	MPa (kpsi)	RT	ASTM D638
Flexural Strength	276 (40)	255 (37)		MPa (kpsi)		IPC-TM-650 2.4.4
Dimensional Stability	<0.3	<0.5	X,Y	mm/m (mil/inch)	after etch +E2/150°C	IPC-TM-650 2.4.39A
Coefficient of Thermal Expansion	11	14	X	ppm/°C	-55 to 288°C	IPC-TM-650 2.1.41
	14	16	Y			
	46	35	Z			

Table 3.5: Datasheet values of Rogers RO4003 and RO4350 PCB

3.7 MY SELECTION

I have decided to make two different designs on which to perform the simulation: one design on a FR-4 substrate and one on a Rogers substrate. This is to see the actual differences in the two substrates. As a pre-cast of the performance, there shouldn't be a big difference between them as I don't have any limitations in terms of dimensions, operating temperature, humidity etc. as this project is for learning purpose. As the two substrates have different properties I will adapt the design to fit the substrate to the 2.4GHz range.

I have selected the Rogers 3003 as my type of Rogers substrate. I have selected this because it has good characteristics for my application. Rogers or other low-cost products might have better properties but might not be as plain all over the whole laminated area as the Rogers 4003. I.e. the Rogers 3003 PCB has a lower dielectric constant, but isn't as plain over the whole laminate area as Rogers 4003, because this is produced for other application purposes like getting rid of heat through the substrate.

3.8 SUBSTRATE LOSSES & ISSUES

3.8.1 Why dielectric loss occur

When an electromagnetic field is applied over a dielectric material, some of the energy within the electromagnetic field will be stored inside the dielectric material. This will cause the atoms inside the dielectric material to realign with the electromagnetic fieldlines, and thereby create friction which will dissipate loss in terms of heat. The loss is determined by the dielectric material, the frequency and the strength of the electromagnetic field, but the physics is more complex than so, and will not be described in this project.

3.8.2 Impedance loss

Losses from reflections are traceable to variations in impedance. These stem from variations in laminate thickness, variations in dielectric constant of the laminate and variations in final etched trace width. The first two of these are traceable to characteristics of the laminate itself and the latter to process uniformity at the fabricator.

3.8.3 Moisture absorption

Substrate materials such as high resin containing PCB's are likely to absorb moisture (a problem in very humid areas). Moisture absorption will increase the dielectric constant, create uneven surface thickness (both will result in a mismatched impedance) and, in the worst case, cause current leakage or short circuit. An extra coating can prevent this, but will also be more affordable.

3.8.4 Temperature expansion

Some laminate materials will expand because of high temperatures. For low cost FR-4 laminate, temperature expansion will happen around 125 °C. This is a problem as soldering tin melts at 185 °C. Common expansions are 2-5 % in thickness @ 300 °C, depending on the material. Expansion can have bad consequences under production as boards can fail even before they hit the market.

4. Design

4.1 ROGERS 3003 DESIGN

The datasheet for Rogers 3003 shows the different standard thickness of the substrate. Figure 4.1 shows an excerpt of the datasheet, where the standard thickness of the substrate is stated.

STANDARD THICKNESS:		STANDARD PANEL SIZE:	STANDARD COPPER CLADDING:
RO3003: 0.005" (0.13 mm) 0.010" (0.25 mm) 0.020" (0.50 mm) 0.030" (0.75 mm) 0.060" (1.52 mm)	RO3006/3010: 0.005" (0.13 mm) 0.010" (0.25 mm) 0.025" (0.64 mm) 0.050" (1.28 mm)	RO3003: 12" X 18" (305 X 457mm) 24" X 18" (610 X 457mm) 24" X 36" (610 X 915mm) RO3006/3010: 18" X 12" (457 X 305mm) 18" X 24" (457 X 610mm) 18" X 36" (457 X 915mm) 18" X 48" (457 X 1.224m)	½ oz. (17µm), 1 oz. (35µm), 2 oz. (70µm) electrodeposited copper foil.

Figure 4.1: Standard thickness of Rogers 3003 substrate.

As this is for educational purposes I made a choice of making the substrate thickness of 1.52 mm. Normally for RF applications one would choose around 0.8 mm, but the thicker the substrate is the wider the track has to be, which will increase my precision during simulations.

I can then calculate the following:

4.1.1 Width

$$W = \frac{c}{2 \cdot f_c \cdot \sqrt{\frac{\epsilon_r + 1}{2}}} = \frac{3 \cdot 10^8}{2 \cdot 2.45 \text{GHz} \cdot \sqrt{\frac{3.0 + 1}{2}}} \approx 43.29 \text{mm}$$

4.1.2 Effective dielectric constant

$$\epsilon_{\text{reff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \cdot \left[1 + 12 \cdot \frac{h}{W} \right]^{-\frac{1}{2}} = \frac{3.0 + 1}{2} + \frac{3.0 - 1}{2} \cdot \left[1 + 12 \cdot \frac{1.52}{43.29} \right]^{-\frac{1}{2}} = 2.8388$$

I can then calculate the effective length.

4.1.3 Effective length

$$L_{\text{eff}} = \frac{c}{2 \cdot f_c \cdot \sqrt{\epsilon_{\text{reff}}}} = \frac{3.8 \cdot 10^8}{2 \cdot 2.45 \text{GHz} \cdot \sqrt{2.8388}} \approx 36.34 \text{mm}$$

4.1.4 Delta length

To calculate delta length ΔL , I calculate the following:

$$\Delta L = 0.412 \cdot h \cdot \frac{(\epsilon_{\text{reff}} + 0.3) \left(\frac{W}{h} + 0.264 \right)}{(\epsilon_{\text{reff}} - 0.258) \left(\frac{W}{h} + 0.8 \right)} = 0.412 \cdot 1.52 \cdot \frac{(2.8388 + 0.3) \left(\frac{43.29 \text{mm}}{1.52 \text{mm}} + 0.264 \right)}{(2.8388 - 0.258) \left(\frac{43.29 \text{mm}}{1.52 \text{mm}} + 0.8 \right)} = 747.7 \mu\text{m}$$

4.1.5 Actual length

With the effective length and delta length I can calculate the actual length of my patch:

$$L = L_{\text{eff}} - 2 \cdot \Delta L = 36.34 \text{mm} - 2 \cdot 747.6993 \mu\text{m} \approx 34.84 \text{mm}$$

4.1.6 Ground plane

The width and length ground plane can be calculated by the following equations:

$$W_g = 6 \cdot h + W = 6 \cdot (1.52) + 43.29 \text{mm} \approx 52.41 \text{mm}$$

$$L_g = 6 \cdot h + L = 6 \cdot (1.52) + 34.84 \text{mm} \approx 43.96 \text{mm}$$

4.2 FR-4 DESIGN

Same procedure was done with the FR-4 substrate. As there are a lot of different FR-4 substrates on the market I choose to use a standard thickness of 1.6 mm. When making my simulation model a standard dielectric for FR-4 was defined as 4.4, so I choose to make my design from this.

I can then calculate the following:

4.2.1 Width

$$W = \frac{c}{2 \cdot f_c \cdot \sqrt{\frac{\epsilon_r + 1}{2}}} = \frac{3 \cdot 10^8}{2 \cdot 2.45 \text{GHz} \cdot \sqrt{\frac{4.4 + 1}{2}}} \approx 37.26 \text{mm}$$

4.2.2 Effective dielectric constant

$$\epsilon_{\text{reff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r + 1}{2} \cdot \left[1 + 12 \cdot \frac{h}{W} \right]^{-\frac{1}{2}} = \frac{3.0 + 1}{2} + \frac{3.0 + 1}{2} \cdot \left[1 + 12 \cdot \frac{1.6}{37.26} \right]^{-\frac{1}{2}} = 4.081$$

I can then calculate the effective length.

4.2.3 Effective length

$$L_{\text{eff}} = \frac{c}{2 \cdot f_c \cdot \sqrt{\epsilon_{\text{reff}}}} = \frac{3.8 \cdot 10^8}{2 \cdot 2.45 \text{GHz} \cdot \sqrt{4.081}} \approx 30.31 \text{mm}$$

4.2.4 Delta length

To calculate delta length ΔL , I calculate the following:

$$\Delta L = 0.412 \cdot h \cdot \frac{(\epsilon_{\text{reff}} + 0.3) \left(\frac{W}{h} + 0.264 \right)}{(\epsilon_{\text{reff}} - 0.258) \left(\frac{W}{h} + 0.8 \right)} = 0.412 \cdot 1.6 \cdot \frac{(4.081 + 0.3) \left(\frac{37.26 \text{mm}}{1.6 \text{mm}} + 0.264 \right)}{(4.081 - 0.258) \left(\frac{37.26 \text{mm}}{1.6 \text{mm}} + 0.8 \right)} = 738.61 \mu\text{m}$$

4.2.5 Actual length

With the effective length and delta length I can calculate the actual length of my patch:

$$L = L_{eff} - 2 \cdot \Delta L = 30.31mm - 2 \cdot 738.61\mu m \approx 28.83mm$$

4.2.6 Ground plane

The width and length ground plane can be calculated by the following equations:

$$W_g = 6 \cdot h + W = 6 \cdot (1.6) + 37.26mm \approx 46.86mm$$

$$L_g = 6 \cdot h + L = 6 \cdot (1.6) + 28.83mm \approx 38.43mm$$

By looking at the calculation for Rogers 3003 and FR-4 substrate the conclusion can already be made as by increasing the dielectric constant the patch dimensions will be smaller.

5. Simulations

5.1 APPLICATION SOFTWARE

For my simulation I used an application from Ansoft called HFSS. After a lot of research I found out that Ansoft HFSS is the most precise antenna simulation software on the market. Figure 5.1 shows the typical layout for Ansoft HFSS.

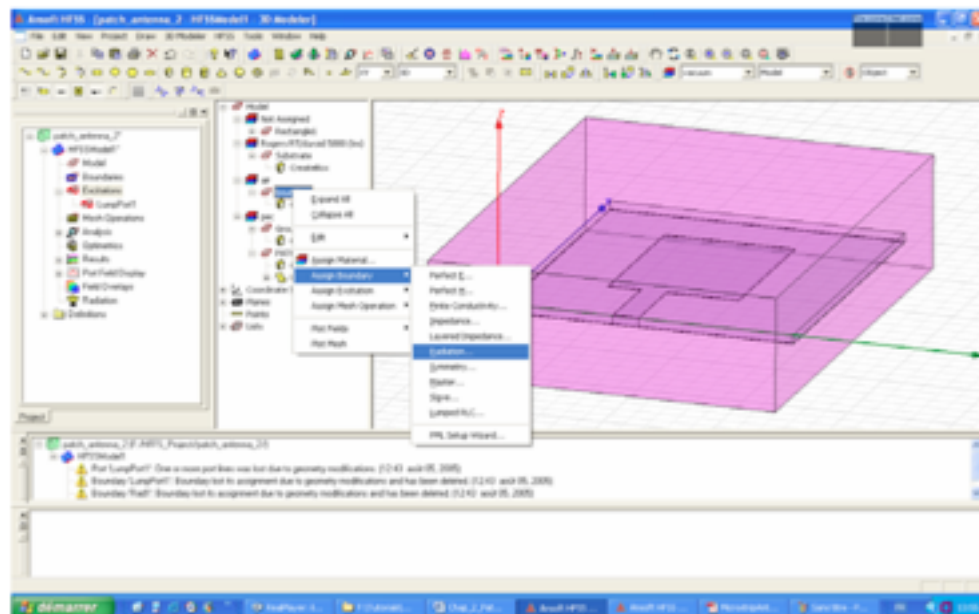


Figure 5.1: Layout of Ansoft HFSS.

I was looking into other antenna software such as Agilent ADS, AWR Microwave Office and I even created a model in Ansoft Designer SV2 as well. The Ansoft Designer model was used as a fast model for my antenna patch, but missing details made me switch to Ansoft HFSS.

5.2 ROGERS 3003 SIMULATION

5.2.1 1. Design

By using Ansoft HFSS simulation software I inputted the calculations from section 4.1 into a simulation model. Figure 5.1 shows my simulation model my patch antenna.

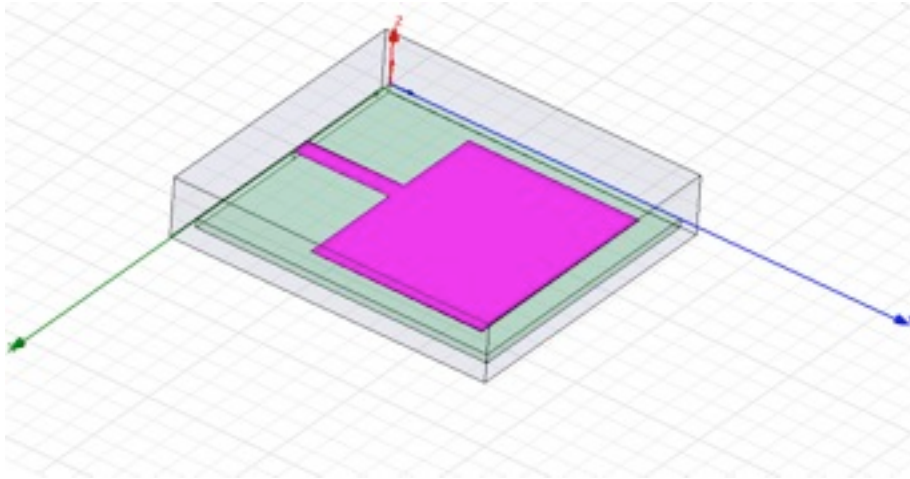


Figure 5.1: Simulation model for my patch antenna with Rogers 3003.

To calculate the feed width I used a calculator provided by AppCad. AppCad is a very well known calculator used by engineers who design microstrip antennas. Figure 5.2 shows the AppCad calculator with the calculated values.

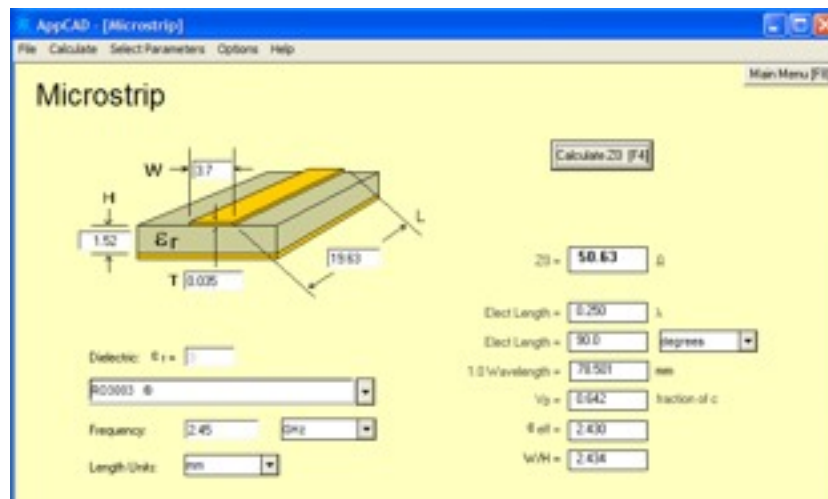


Figure 5.2: AppCad calculator.

The feed line is calculated as a 50Ω transmission line. As seen from figure 5.2 the width of the feed line was calculated to 3.7 mm. The length of the feed line does not matter as it is only the loss in the copper length that counts. By using the calculator the length is designed from the fact that the phase has been rotated 90 degrees giving me a length of 19.63 mm.

As I have to insert the feed line, the length of the substrate and ground will be longer in my model. The calculated ground planes from section 4.1.6 shows the minimum length, so by extending the ground and substrate length should not have any effect.

Figure 5.3 shows the return loss for my antenna.

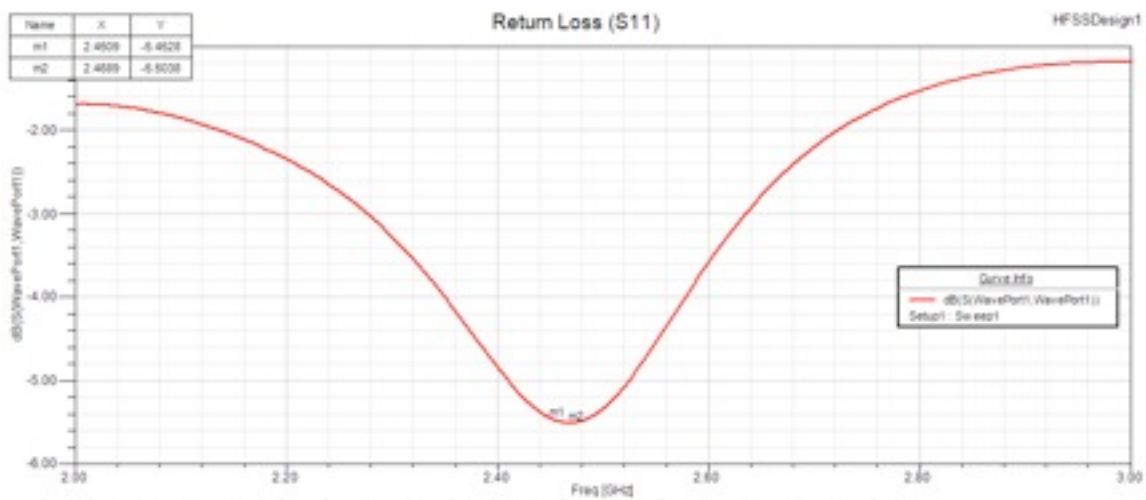


Figure 5.3: Return loss of patch antenna.

By looking at figure 5.3 the antenna gives me a return loss of -5.5dB @ 2.45 GHz, which is not very optimal.

Figure 5.4 shows the smith chart of my antenna.

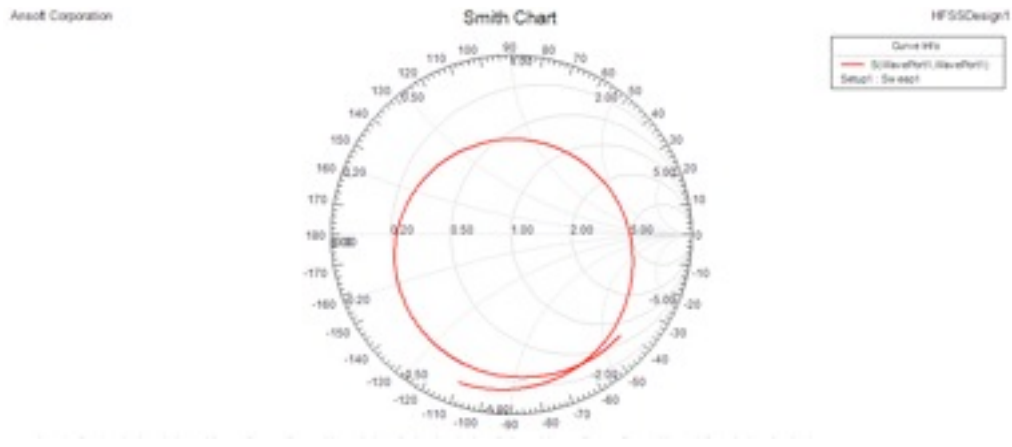


Figure 5.4: Smith chart of patch antenna.

Looking at figure 5.4 there is no doubt that the system is not matched. By using an online calculator⁴ I was able to verify that my calculations for the length and width were correct. I noticed on the calculator that the impedance of the antenna was 180Ω . It was obvious that connecting a 50Ω feed line would result in a mismatch.

⁴ <http://www.emtalk.com/mpacalc.php>

5.2.2 2. Design

By determining the outcome of the first design I need to alter my design to match the antenna impedance with the load impedance of 50Ω . By doing some research I ended up using a $\lambda/4$ -transformer in which I was able to transform my patch antenna down close to 50Ω . As my original width of my feed was 3.7, I needed another feed line in between the patch and the original feed line with a width of $3.7/4 = 0.925$ mm. The longer the first feed line is, the longer the patch has to be to keep same frequency.

It is also important to make sure that the length of the feed line doesn't have the length so standing waves within the feed line appear. If this happens the waves magnitude will cancel out each other and you will thereby have a lower return loss.

Figure 5.4 shows my new design for rogers 3003 substrate.

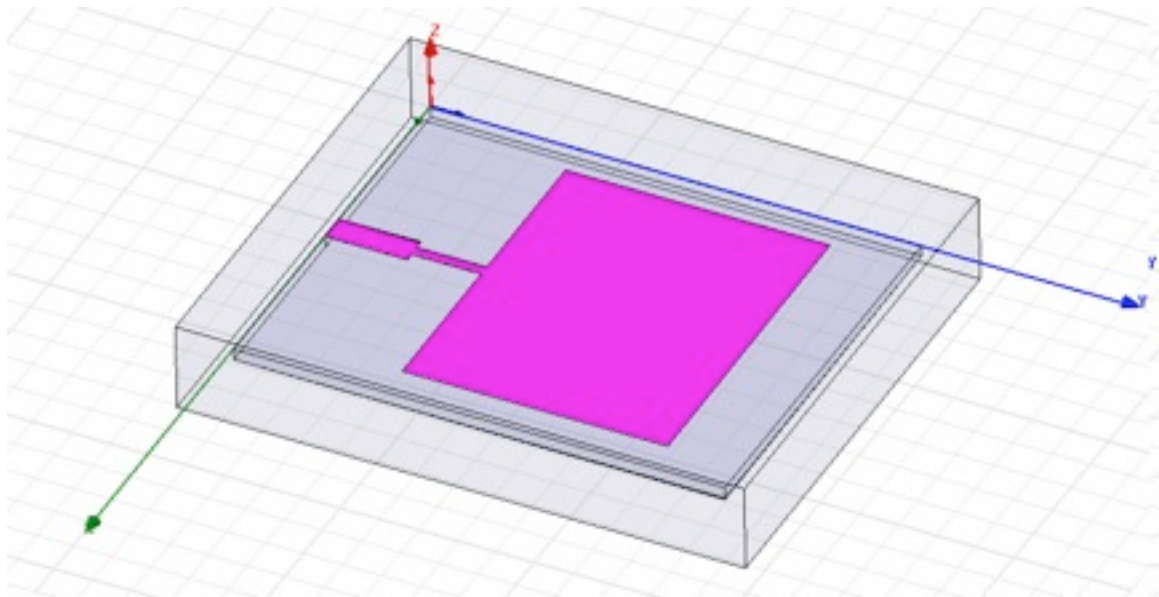


Figure 5.4: Simulation model of the final design with Rogers 3003.

After obtaining the new design, it took some tweaking of the length and width to be able to get a better result. The tweaking was minimalistic and patch has to be very precise. A difference in the length of just 0.5 mm could alter the center frequency radically and make a difference at 2.45 GHz return loss going up from -20dB to around -10dB.

Figure 5.5 shows the return loss for the final result.

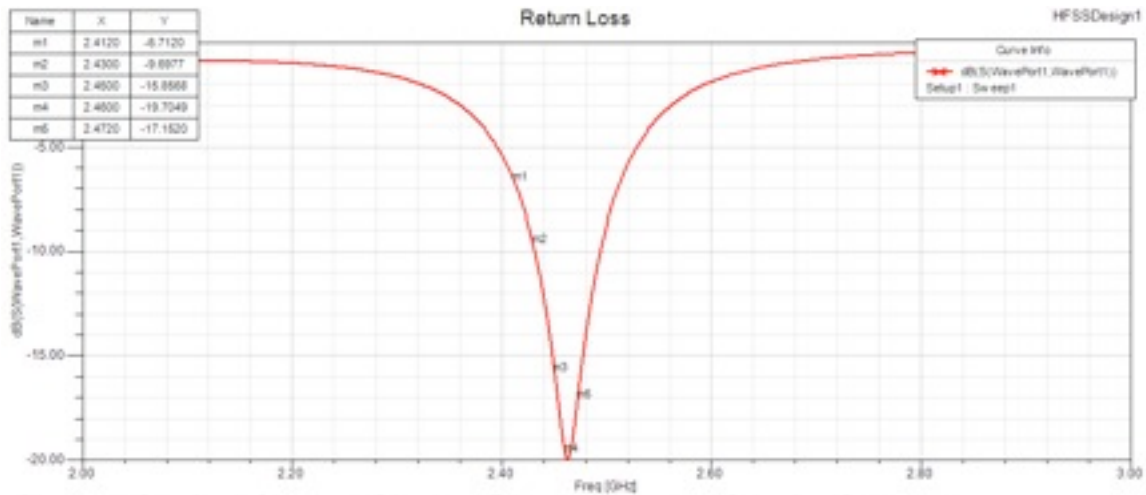


Figure 5.5: Return loss for the final result.

Looking at figure 5.5 the antenna gives a very narrow bandwidth. I knew that it was sure to be narrow, but I didn't think it would be quite as narrow as the result. Table 5.6 shows the marker values.

Marker	Frequency	Return Loss
M ₁	2.4120 GHz	-6.71 dB
M ₂	2.4300GHz	-9.70 dB
M ₃	2.4500 GHz	-15.86 dB
M ₄	2.600 GHz	-19.70 dB
M ₅	2.4720 GHz	-17.15 dB

Table 5.6: Return loss trace measurements.

Looking at table 5.6 it can be seen that the antenna covers all Wi-Fi channels, although a return loss between -6.71 dB to -9.70 of some of the first channels is not that impressive and is doubtful that these will be a good choice. Having a broadband of under -15 dB from 2.44 GHz to 2.48 GHz is acceptable. Using channel 11 @ 2.462 GHz with a return loss of around -18 dB would be an ideal choice.

Figure 5.7 shows the smith chart of the final result.

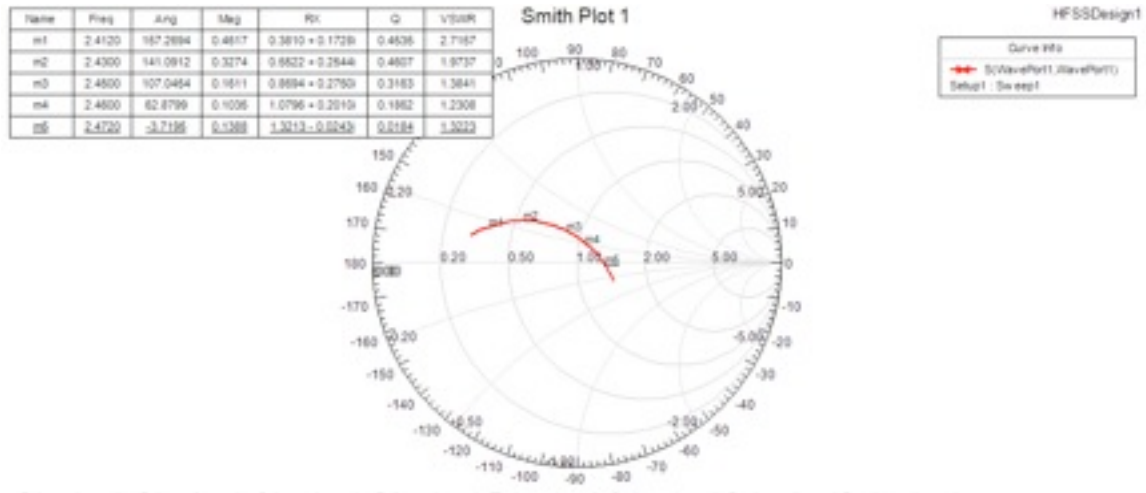


Figure 5.7: Smith chart of final result.

Table 5.8 gives a clarification of the markers.

Marker	Frequency	VSWR
M1	2.4120 GHz	2.7157
M2	2.4300 GHz	1.9737
M3	2.4500 GHz	1.3841
M4	2.4600 GHz	1.2308
M5	2.4720 GHz	1.3223

Table 5.8: Smith chart trace measurements.

From the VSWR values in table 5.8 marker M₄ @ 2.46 GHz gives me the best result. M₃ and M₄ is within an acceptable range of, while M₁ and M₂ is not very ideal as the loss will be too big which the return loss from figure 5.5 would also state.

Figure 5.9 shows the total radiation of the antenna.

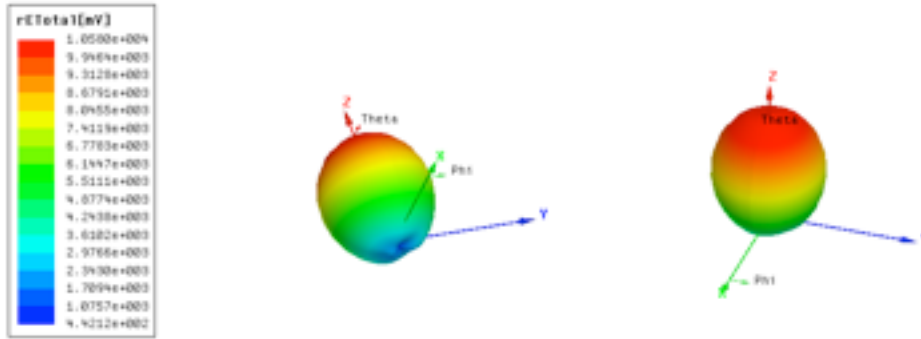


Figure 5.9: Total radiation of the final result.

Looking at figure 5.9 it can be seen that most radiation is going in the Z -direction. Unfortunately it doesn't tell us the distance. At the top red field it gives a total radiation of 10.58 mV.

I was trying to get a beamwidth radiation pattern, but my inexperienced expertise in Ansoft HFSS made it difficult.

5.3 FR-4 SIMULATION

5.3.1 1. Design

By using same approach as section 5.2.1 I inputted the calculations from section 4.2 into the simulation model. Figure 5.10 shows my simulation model my patch antenna.

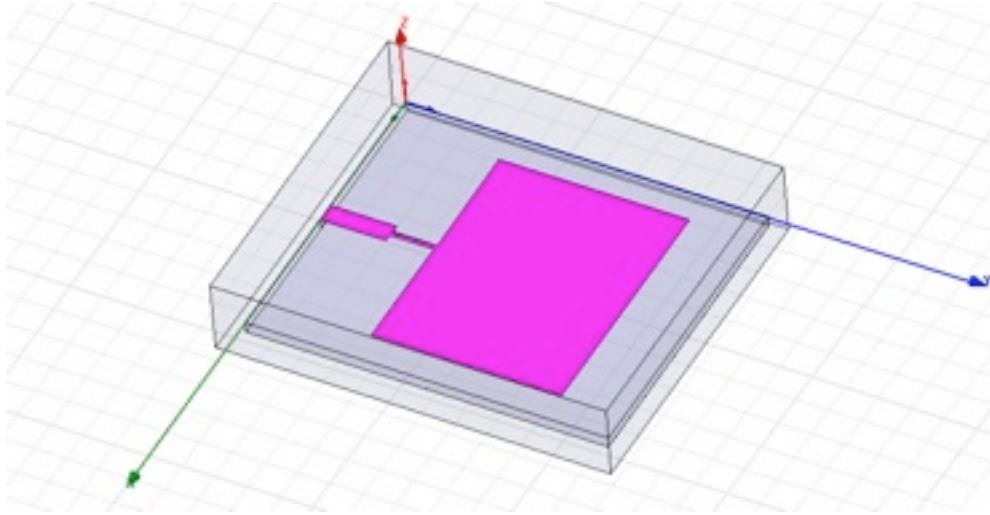


Figure 5.10: Simulation model for my patch antenna with FR-4 substrate.

I already included my $\lambda/4$ feed line transformer in the design. Figure 5.11 shows the return loss for my FR-4 antenna.

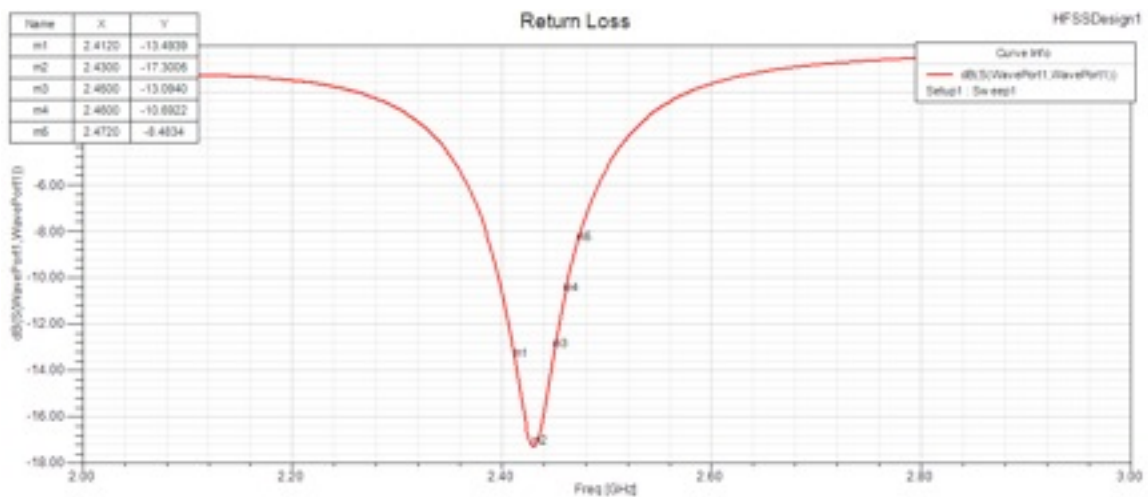


Figure 5.11: Return loss of the patch antenna with FR-4 substrate.

Table 5.12 clarifies the markers.

Marker	Frequency	Return Loss
M1	2.4120 GHz	-13.49 dB
M2	2.4300GHz	-17.3 dB
M3	2.4500 GHz	-13.09 dB
M4	2.600 GHz	-10.69 dB
M5	2.4720 GHz	-8.48 dB

Table 5.12: Return loss trace measurements.

Looking at figure 5.11 and table 5.12 it appears that the patch antenna with FR-4 substrate have a wider broadband where more frequencies within the Wi-Fi spectrum are covered with an acceptable return loss.

Figure 5.13 shows the smith chart of the 1. design.

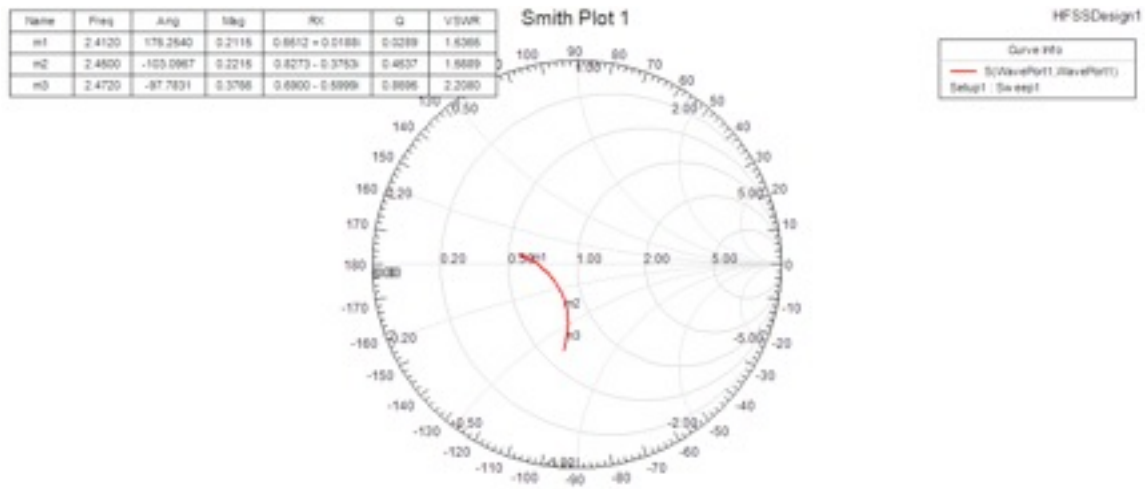


Figure 5.13: Smith chart of the path antenna with FR-4 substrate.

Table 5.14 gives a clarification of the markers.

Marker	Frequency	VSWR
M ₁	2.4120 GHz	1.5365
M ₂	2.4500 GHz	1.5689
M ₃	2.4720 GHz	2.2080

Table 5.14: Smith chart trace measurements.

From the VSWR values in table 5.11 marker M₁ and M₂ is doubtful acceptable. Although the frequencies gives me a good return loss it is not very perfectly matched. Marker M₃ would not be an acceptable VSWR value.

To try and create a better result I once again tweaked the dimensions of my patch antenna.

5.3.2 2. Design

Figure 5.15 shows my 2. design with FR-4 substrate.

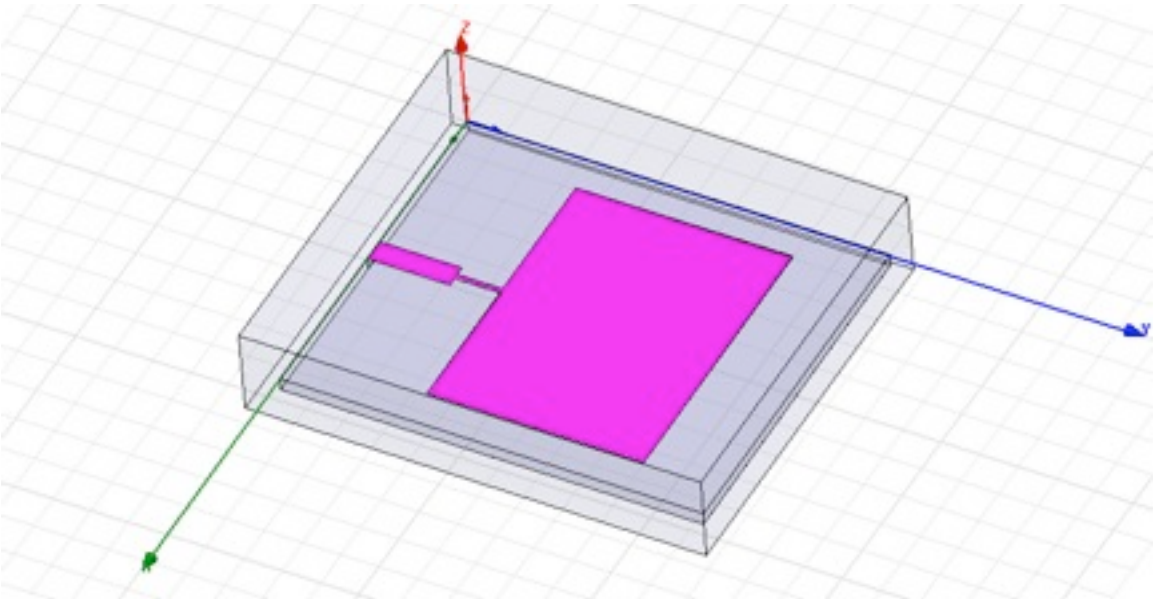


Figure 5.15: Simulation model of the 2. design with FR-4 substrate.

Figure 5.16 shows the return loss for the 2. design.

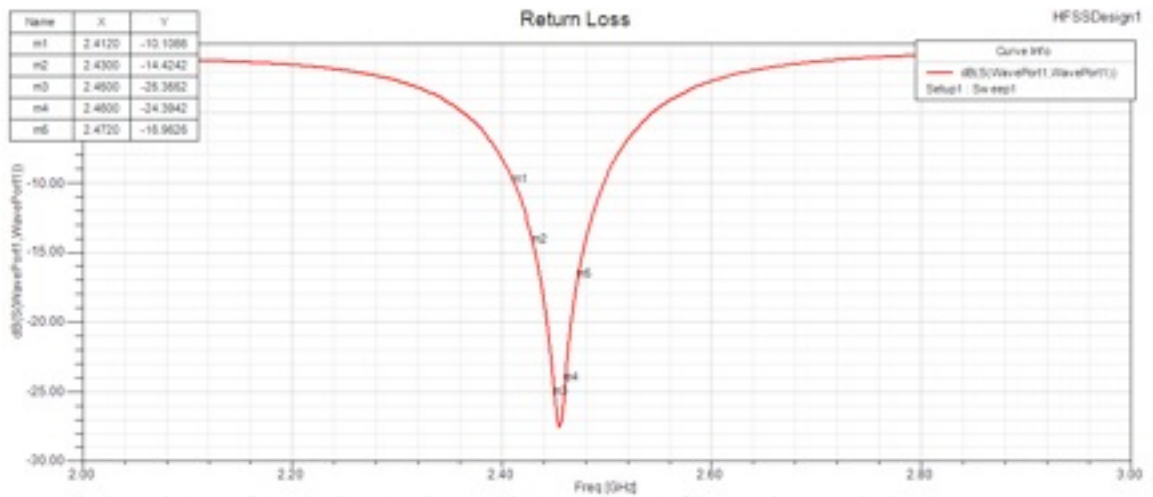


Figure 5.16: Return loss for the 2. design with FR-4 substrate.

Table 5.17 gives a clarification of the markers.

Marker	Frequency	Return Loss
M1	2.4120 GHz	-10.11 dB
M2	2.4300GHz	-14.42 dB
M3	2.4500 GHz	-25.36 dB
M4	2.600 GHz	-24.39 dB
M5	2.4720 GHz	-16.96 dB

Table 5.17: Return loss trace measurements.

Looking at figure 5.16 and table 5.17 the antenna gives a narrower bandwidth from M3 @ 2.45 GHz with better results. Figure 5.18 shows the smith chart of the 2. design.

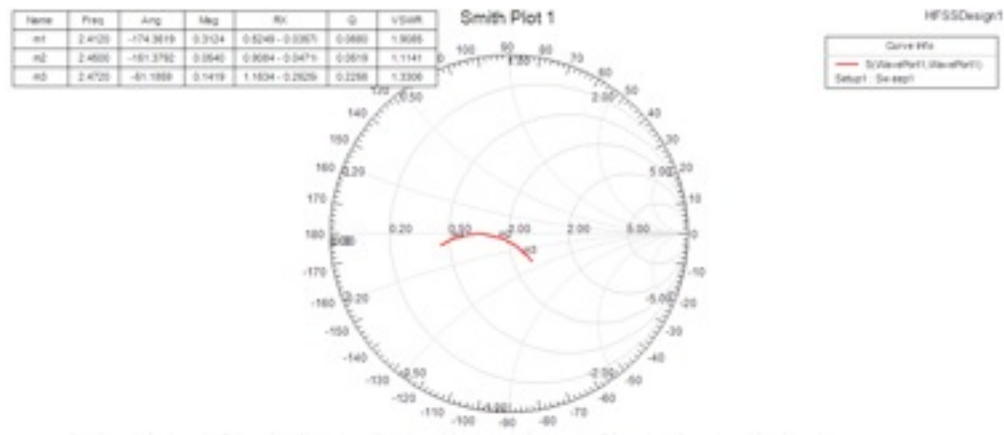


Figure 5.18: Smith chart of final result.

Table 5.19 gives a clarification of the markers.

Marker	Frequency	VSWR
M1	2.4120 GHz	1.9085
M2	2.4500 GHz	1.1141
M3	2.4720 GHz	1.3306

Table 5.19: Smith chart trace measurements.

Looking at the VSWR values from table 5.19 the middle and top frequencies in the Wi-Fi channels is covered with good matching and the best VSWR so far. Marker M1 @ 2.412 GHz is not acceptable.

Figure 5.20 shows the total radiation of the antenna.

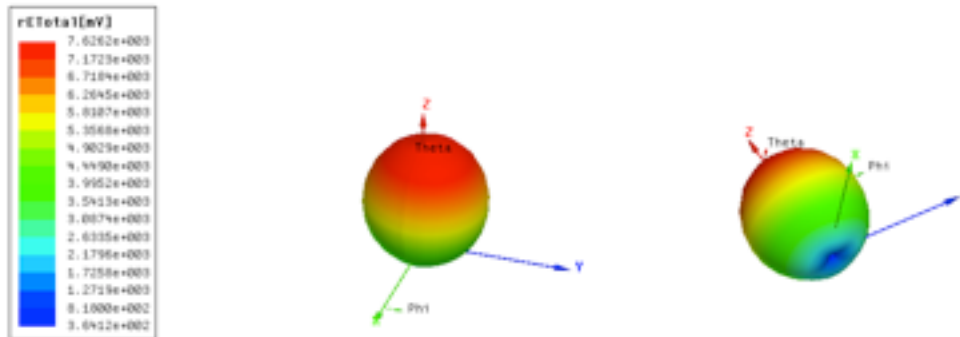


Figure 5.20: Total radiation of the final result.

Looking at figure 5.20 it can be seen that most radiation is going in the Z-direction. Unfortunately it doesn't tell us the distance. At the top red field it gives a total radiation of 7.62 mV.

The 2. design gave me first of all some better results in terms of return loss and VSWR. Especially a narrower band was used where VSWR gave me the best results so far, but didn't cover all the Wi-Fi channels. In a 3. design I will try and make the patch antenna more broadband so it will cover the whole Wi-Fi frequency spectrum with a max VSWR of around 1.3.

5.3.3 3. Design

By trying to make the 3. design more broadband I got awful return loss and smith chart results. Instead I made the broadband even narrower, trying to get a better return loss at the center frequency and thereby drag the frequencies close to the center frequency down the return loss 'whole' to give me a better return loss for all the Wi-Fi frequencies.

Figure 5.21 shows my 3. design for FR-4 substrate.

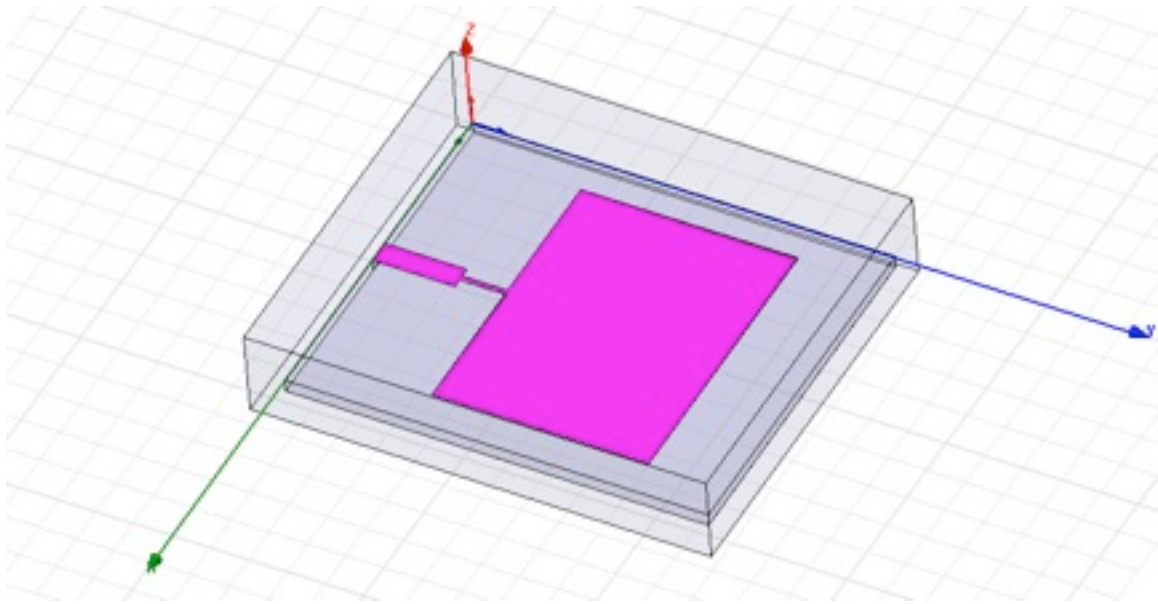


Figure 5.21: Simulation model of the 3. design with FR-4 substrate.

Figure 5.22 shows the return loss for the 3. design.

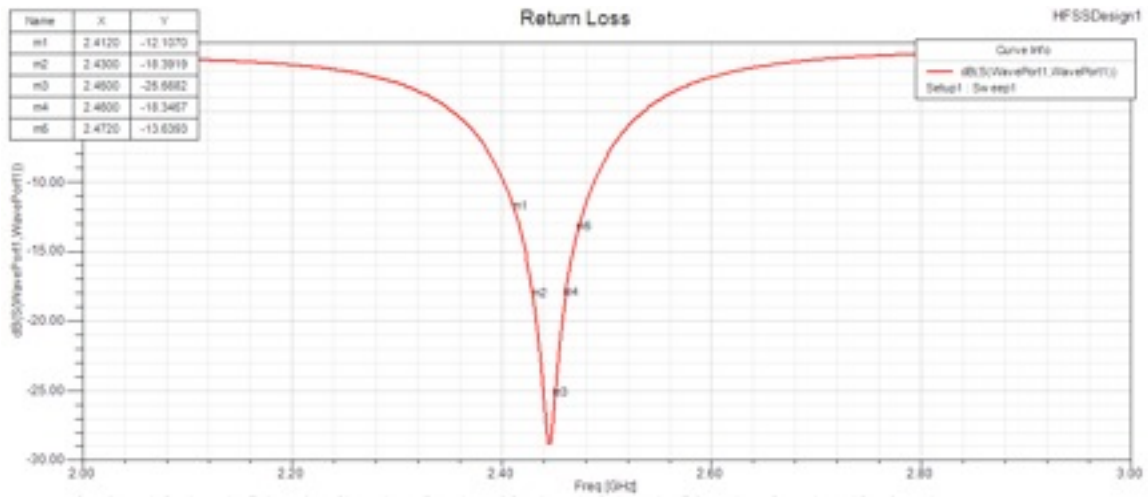


Figure 5.22: Return loss for the 3. design with FR-4 substrate.

Table 5.23 gives a clarification of the markers.

Marker	Frequency	Return Loss
M1	2.4120 GHz	--12.11 dB
M2	2.4300GHz	--18.39 dB
M3	2.4500 GHz	--25.57 dB
M4	2.600 GHz	-18.35 dB
M5	2.4720 GHz	--13.64 dB

Table 5.23: Return loss trace measurements.

Looking at figure 5.22 and table 5.23 all the Wi-Fi frequencies is covered with an acceptable return loss.

Figure 5.24 shows the smith chart of the 3. design.



Figure 5.24: Smith chart of final result.

Table 5.25 gives a clarification of the markers.

Marker	Frequency	VSWR
M1	2.4120 GHz	1.6600
M2	2.4500 GHz	1.1112
M3	2.4720 GHz	1.5252

Table 5.25: Smith chart trace measurements.

Looking at the VSWR values from table 5.19 the antenna has better VSWR values to cover all the Wi-Fi channels. The antenna might be usable for the whole area but Marker M2 @ 2.45 GHz would be the recommended frequency to turn your channel in on (Wi-Fi channel 8,9 or 10 regardless).

Figure 5.26 shows the total radiation of the antenna.

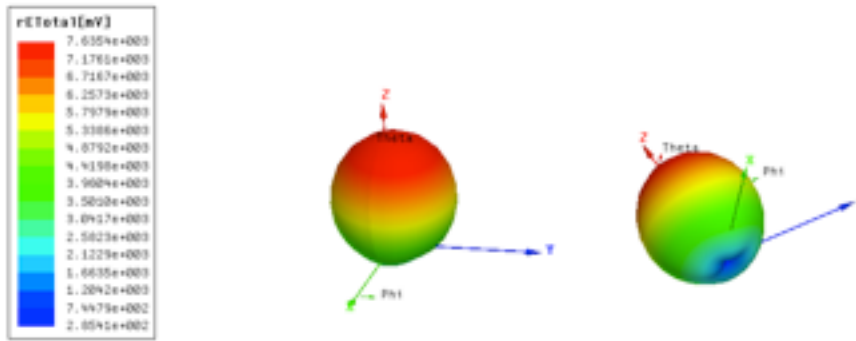


Figure 5.20: Total radiation of the final result.

Looking at figure 5.20 it can be seen that most radiation is going in the Z -direction. Unfortunately it doesn't tell us the distance. At the top red field it gives a total radiation of 7.64 mV.

6. Conclusion

The project has been a success. I was able to calculate and design two microstrip patch antennas from theoretical equations with two different substrates. From the substrate theory I was able to see that the larger the dielectric constant is, the smaller the patch antenna dimensions must be. I was able to go from knowing nothing about antenna simulation tools or simulation applications to actually building a full simulation model in Ansoft HFSS which I chose by recommendations as the best antenna simulations software on the market and which is used by professionals.

Although a simpler software such as Ansoft Designer could have been used as a simulation software, I choose to go with Ansoft HFSS to become acquainted with the software in usage in a future job.

I was able to discover that the feed line is influenced by the patch antenna and its impedance as stated in theory section 2.7.1. Also the patch antenna impedance was not included in my calculations, but I was able to determine the impedance of my antenna to sort out my feed line issues as stated in section 5.2.1 and 5.2.2.

Creating a broadband patch antenna has been more difficult than expected. From theory in section 2.5 I knew that the patch antenna was going to be a narrowband antenna, but I did not expect it to be as narrow as my simulation results point out. This could be because of the interference from the feed line; because the feed line is a patch itself, it can cause radiation that interferes with the patch. This will decrease the bandwidth of the antenna.

In the simulation results in section 5.3.2 I was able to create an antenna that would cover the higher Wi-Fi channel frequencies with a good VSWR, while the lower channel frequencies would suffer. By trying to make the patch antenna more broadband the VSWR would suffer except from the center frequency, as can be seen in section 5.3.3. From this it would be a good solution to connect 2 or 4 patch antennas with different dimensions together in one simulation model to cover a wider spectrum, so VSWR and broadband requirements would be fulfilled. Researching pictures of patch antennas showed that it was not unusual to find these types of designs.

Another way to increase the simulation results would be to apply a different type of feed line such as the probe feed. A transmission feed line is just one out of many types of feed lines that can be applied to a patch antenna. Unfortunately I didn't have enough time to further investigate other feed line methods.

It would have been a big advantage to know the Ansoft HFSS simulation software beforehand as a lot of measurements could have been applied; deeper understanding of radiation and efficiency of the antenna would have come in handy when comparing the two substrates against each other. Because of my inexperience with the software I was unfortunately not able to make a beamwidth radiation. I was able to illustrate a 3D polar plot to make a visualization of the antenna radiation.

The project has been much more theoretical than I expected it to be. Throughout the project I learned that it is impossible to know how to construct an antenna without actually understanding the theory behind it.

Although I spent a lot of time during the semester working on this project, the patch antenna is just a simple part of microstrip antenna theory. Microstrip antennas and their theory get much more complex as you want to create more efficiency and wider bandwidth. It shows that a patch antenna of this size cannot be implemented in any cellphone or other small gadgets. This can be done by bending several microstrip transmission lines.

I have learned a lot from this project and my mind has been opened to a whole new world of antennas. My knowledge about antennas has been increased but as I got deeper, I found out that there is much more to know than just a simple patch antenna. The deadline was a really critical aspect of the project. There is so much more I want to apply to my design to try and increase my antenna with better results. Also there was not enough time to actually create the antenna so it could be tested in real life.

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