DIRECTIONAL PATCH ANTENNA ARRAY DESIGN FOR DESKTOP WIRELESS INTERNET

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2010
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Directional Patch Antenna Array Design for Desktop Wireless Internet

Abstract
To improve desktop wireless internet access, two patch antenna configurations were investigated in order to find an alternative for a dipole antenna, which is commonly used for Wi-Fi access. A Linksys WMP54G pci-card wireless adapter was used to test three antennas for 802.11g communications. The standard WMP54G dipole antenna and two patch antenna configurations were tested for Wi-Fi signal strength. The antennas were also tested in Cal Poly’s anechoic chamber to obtain resonant frequency and gain measurements. A 2.437 GHz center frequency, 72 MHz bandwidth antenna consisting of a single microstrip patch was constructed. After the patch antenna was determined to operate over the frequency range, an array of four patch elements on a single substrate was constructed. Although the patch array did not match the calculated gain, beamwidth and sidelobe radiation level improvements over a single patch design were achieved. Both patch antenna configurations had greater gain over the WMP54G dipole antenna, however improved Wi-Fi signal strength was only accomplished when used in direct line of sight communications. In conditions where obstructions, such as walls, prevented line of sight, the dipole antenna achieved greater signal strength. Line of sight connections favored the directional antennas.
I. Introduction

A home wireless network is relatively simple to setup and enables wireless internet access at high data rates for an entire household. Because of these advantages, wireless networks are a popular choice over intrusive wired networks. With the close proximity of multiple wireless networks, interference from adjacent residential and commercial systems can prove problematic. The IEEE 802.11g protocol has 14 channels, 11 allowed for use in the US [1]. Each channel has 22 MHz bandwidth and 5 MHz separation between channel center frequencies, see Figure 1. The US operating frequencies range from 2.401 GHz to 2.473 GHz, a 72 MHz bandwidth. Each channel overlaps with between four and eight adjacent channels, leaving the possibility of only three simultaneous operating, non-overlapping channels. This means that an apartment complex with a high density of wireless networks will encounter difficulties in establishing connections without interference.

Figure 1 shows the center frequencies of the 802.11g channels and which channels overlap with neighboring channels [1]. Channels 1, 6, and 11, bolded in Figure 1, are the only three channels which can be operated simultaneously while not overlapping. Note that channels 12 through 14, although shown in the diagram, are not permitted for use in the US.

In addition to 802.11g channel saturation, many devices such as microwave ovens and Bluetooth adaptors cause additional interference in the 2.4 GHz range. Current wireless adapters sold for desktop computers use omnidirectional dipole antennas that are very susceptible to interference. A directional antenna will alleviate many of these problems, as relatively low-power side lobes can minimize signals received from unintentional sources and diminish interference sent to other networks. The increased main beam gain also allows improved signal reception.

In order to be used in a home or office environment, the antenna should include necessary hardware to mount on the computer, but should also be compact. To allow directional adjustments, the mounting structure may include a magnetic base with a rotating mount. A depiction of the patch antenna configuration is shown in Figure 2.

Figure 1: <en.wikipedia.org/wiki/802.11g> “2.4 GHz Wi-Fi channels (802.11b,g WLAN).svg” distributed for free use under the Creative Commons Attribution-Share Alike 3.0 Unported licence
Figure 2: Patch Antenna Configuration
II. Background

A patch antenna was chosen because of its directivity, low profile, and ease of fabrication. Patch antennas consist of a copper sheet on a grounded dielectric substrate. The parallel patch and ground plane create a transmission line within the substrate. Patches may be rectangular, circular, or one of many other configurations. A rectangular patch was chosen for this project because it provides linear polarization, which will allow for the best communication with wireless routers which use linearly polarized dipole antennas. The patch is a resonant antenna; the patch is designed so that the length is a half wavelength in the dielectric. The patch width affects characteristic impedance and is determined to ensure matching between the coax cable and the microstrip patch.

A patch antenna is more compact than a traditional half wavelength dipole antenna because the wavelength in the dielectric is less than in free space. The substrate has a greater dielectric permittivity than free space, causing electromagnetic waves to travel more slowly at a given frequency. Since the waves are traveling slower, the wavelength increases. Substrate material with a higher dielectric constant will result in a patch design of a shorter length. When choosing substrate height, there is a tradeoff between efficiency and bandwidth. A thicker substrate will be more lossy through the creation of surface waves, but will allow for a higher bandwidth than patch antennas with a thinner substrate [2].

FR-4 was chosen as the substrate for this project. Available FR-4 boards have a dielectric constant of 4.4 and are 62 mils thick, whereas the available Duroid has a dielectric constant of 2.33 and a thickness of 31 mils. The choice of FR-4 over Duroid allows for a greater bandwidth but a reduced efficiency. FR-4 is also advantageous due to lower cost. The dielectric constant of FR-4 material is not as closely monitored as Duroid. FR-4 is produced for circuit board substrates and is not solely intended for microstrip circuits. Duroid is produced specifically for microstrip circuits and its dielectric constant is maintained to ensure microstrip circuits fabricated with a Duroid substrate operate as designed. The choice of FR-4 as a substrate, early in the design process, overlooked the dielectric constant inaccuracies, causing design difficulties. This is discussed in the “Design Difficulties” section of Chapter VI.
III. Design

System Overview

Both a single patch antenna and patch antenna array were fabricated using the milling machine in Cal Poly’s chamber lab. The FR-4 dielectric board as well as the soldered SMA connectors were supplied by the chamber lab. The final constructed antennas are shown in Figures 3 and 4. The antennas are supported by a rotating mount structure. Both antennas are fed by coax cable connected to the WMP54G pci-card. The dielectric substrate is supported in each corner by plastic stand-offs, keeping the ground plane from contacting the mounting hardware. The single patch substrate length and width dimensions are 4” by 4”. A Plexiglas base is used to support the antenna and mounting hardware. For further improvement, magnets may be added to the base, further increasing stability.

![Patch Antenna in use for Wi-Fi communications](image)

Figure 3: Patch Antenna in use for Wi-Fi communications
Refer to Appendix A for the single patch antenna design procedure.

Patch Antenna Array

In order to increase main beam gain, reduce side lobe radiation, and increase directivity, the patch antenna design was expanded to a four element array. The design layout is shown in Figure 5. Four elements are used, separated by $\lambda/2$. The patch length and width for each element is the same as the single patch antenna described above. The probe position was optimized in HFSS to ensure a 50 Ohm match including adjacent patch coupling. A rectangular distribution of antenna elements was chosen to obtain identical E-plane and H-plane array factor patterns. Four identical antenna elements were used to allow array factor application to the measured patch radiation pattern for array predictions. The $xyz$ coordinates are defined to the left of the figure. The array substrate is in the $xy$-plane. The $z$-direction is perpendicular to the substrate.
The design values shown in Figure 5 are outlined in Table 1. Note that the element spacing refers to the distance between probe feed locations or between corresponding edges of adjacent patch edges, not to the separation distance. Separation distance refers to the distance between adjacent patch edges.

**Table 1: Summary of values chosen for Patch Array Dimensions**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dimension</th>
</tr>
</thead>
<tbody>
<tr>
<td>Patch Width</td>
<td>W = 37.5759 mm</td>
</tr>
<tr>
<td>Patch Length</td>
<td>L = 28.2084 mm</td>
</tr>
<tr>
<td>Element Spacing</td>
<td>λ/2 = 61.551 mm</td>
</tr>
<tr>
<td>E-Plane Separation Distance</td>
<td>33.35 mm</td>
</tr>
<tr>
<td>H-Plane Separation Distance</td>
<td>23.97 mm</td>
</tr>
<tr>
<td>Probe feed location</td>
<td>6.95 mm</td>
</tr>
</tbody>
</table>
The patches are fed by a corporate feed network composed of three cascaded Wilkinson dividers: see Figure 6 for diagram. Each of the four elements receives an equal amplitude, in phase signal. 50 Ohm transmission lines are used to ensure matching between the antennas and feed network. Vias are drilled in corresponding locations in the power divider and antenna array boards allowing the traces to be soldered to the patch elements. Wilkinson dividers were used over Tee section dividers due to inter-port isolation provided by Wilkinson Dividers. In the diagram, the patch antenna locations, shown with dotted lines, are on the opposite side of the substrate and are connected through the vias.

Figure 6: Power Divider Board Layout

Figure 7 shows a typical Wilkinson divider layout with design values for this project. The input and output ports are 50 Ohm microstrip traces. The two traces, connecting the input and output ports, have impedance $Z_0\sqrt{2} = 70.71$ Ohms and length $\lambda/4 = 30.78$ mm. A 100 Ohm resistor connects the input and output ports. The power divider is modeled using ADS to determine trace widths.
A Wilkinson divider with curved $\lambda/4$ sections was designed. This design is advantageous because a surface mount resistor is used instead of a traditional resistor. At the operating frequency of the antenna, a traditional resistor has electrically long, longer than $\lambda/10$, leads. Using long leads will create reflections, decreasing the efficiency of the divider. Using the ADS “DesignGuide” tool, the Wilkinson divider in Figure 8 was developed.
The DesignGuide parameters include center frequency, bandwidth, characteristic impedance, power ratio, and resistor gap. A space is between the $\lambda/4$ transformers is specified as 30 mils, allowing for a 5 mil overlap on each side of the 40 mil length chip resistor. ADS optimized the initial design to obtain the 2.44GHz center frequency. The $\lambda/4$ traces were extended by 2.388 mm. The optimized Wilkinson divider S-parameters are shown in Figure 9.

![Optimized Wilkinson Divider Characteristics](image)

The Wilkinson divider $|S_{21}|$ and $|S_{31}|$ are about -3dB for all frequencies across the operating range. $|S_{11}|$ and $|S_{22}|$ exhibit values less than -35 dB. Isolation between output ports, $|S_{23}|$, is less than -30 dB from 2 to 3GHz. Input and output port analysis meets the -10 dB return loss required for efficient operation.

The feed network was created by cascading three identical Wilkinson dividers. The feed network layout is shown in Figure 10. The microstrip line lengths connecting the Wilkinson Dividers were determined to provide $\lambda/2$ separation between probe feed locations. The line lengths at the input and output ports do not affect the S-parameter magnitudes. The input port will be connected to a 50 Ohm coax cable and the output ports will connect to the patch antennas. Line length will affect the phase of the received signal, however this will not affect operation because the traces to each antenna are equal lengths. The phase must be equal between ports for the array to operate correctly. A phase difference between ports will direct the main lobe direction reduce gain. Ports with equal phase were designed to provide maximum gain at broadside.
ADS simulation of the feed network resulted in the S-parameters shown in Figure 11.

![Figure 10: ADS Layout of Power Divider Circuit (colors inverted)](image)

![Figure 11: Feed network S-Parameter analysis](image)
The feed network simulation shows return loss below than the -10 dB minimum for matching at input and all output ports. The Wilkinson divider and feed network S-parameter values are summarized in Table 2.

Table 2: Measured and ideal power divider S-Parameters

<table>
<thead>
<tr>
<th></th>
<th>$S_{11}$ (dB)</th>
<th>$S_{22}$ (dB)</th>
<th>$S_{21}$ (dB)</th>
<th>$S_{23}$ (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ideal 2-way Divider</td>
<td>$-\infty$</td>
<td>$-\infty$</td>
<td>-3.0</td>
<td>$-\infty$</td>
</tr>
<tr>
<td>Single Divider</td>
<td>-34.627</td>
<td>-34.894</td>
<td>-3.0122</td>
<td>-34.952</td>
</tr>
<tr>
<td>Ideal 4-way Divider</td>
<td>$-\infty$</td>
<td>$-\infty$</td>
<td>-6.0</td>
<td>$-\infty$</td>
</tr>
<tr>
<td>Divider Board</td>
<td>-30.017</td>
<td>-26.154</td>
<td>-6.0256</td>
<td>-38.551</td>
</tr>
</tbody>
</table>

The feed network and antenna array substrates are combined into a single unit by connecting the substrate ground planes by using a conductive epoxy. Vias in the feed network and array substrates allow for a conducting pin to connect the feed traces to the patch antennas through a hole in the ground plane. Figure 12 shows a cross section of this configuration.
IV. Antenna Testing

The dipole radiation patterns shown in Figures 13 and 14 were taken in the anechoic chamber. A wood and plastic test fixture was constructed to minimize reflections and input impedance effects. Although most fasteners were nylon, a two metal nuts and two washers were used. Although any metal in the chamber is not ideal, metal hardware was minimized, giving the most accurate test possible with the given materials. Metal hardware was used because the test fixture could not be created without the rigidity provided by metal hardware. No metal was contacting the antenna during testing. The test fixture used is shown in Figure 17. The most accurate readings were taken as the swept angle was between -90° and 90°. At angles beyond ±90°, the test fixture was in-between the send and receive antennas. This is shown in Figure 13 where |S21| decreases near ±180 degrees. The dipole radiation pattern is expected to be uniform for all angles in the H-Plane.

**Wi-Fi Dipole Measured H-Plane Co-Pol Radiation Pattern (Phi = 90; Swept Theta)**

![Figure 13: Measured H-Plane Dipole Radiation Pattern](image-url)
Figure 14 shows the E-pane and H-plane configurations that were used to obtain the measurements in Figures 13 and 14.

Figure 15 shows the E-pane and H-plane configurations that were used to obtain the measurements in Figures 13 and 14.
Shown in Figure 16 and Figure 17, test fixtures build out of wood and plastic materials were used to support the antennas while testing in the anechoic chamber. The use of non-metallic materials reduces reflections, helping to simulate a free space environment.

Figure 16: Patch antenna on test fixture in anechoic chamber

Figure 17: WMP54G standard dipole antenna on test fixture
Figure 18 shows a return loss measurement of the WMP54G dipole antenna. As shown in Table 3, the dipole antenna is matched below -15 dB return loss for the 802.11g operating frequency range. The antenna has a 152 MHz -10 dB return loss bandwidth.

| Freq (GHz) | $|S_{11}|$ (dB) | Operating Point   |
|------------|----------------|-------------------|
| 2.366      | -10.117        | Min Operating Freq|
| 2.401      | -14.888        | 802.11g Min       |
| 2.438      | -24.608        | 802.11g Center    |
| 2.446      | -25.332        | Best Match        |
| 2.473      | -16.859        | 802.11g Max       |
| 2.518      | -10.201        | Max Operating Freq|
V. System Results

Table 4 summarizes the testing results and expected values of gain and beamwidth for the antenna configurations tested.

Table 4: Gain and Beamwidth calculation for each antenna configuration

<table>
<thead>
<tr>
<th>Antenna Configuration</th>
<th>Max Gain (dB)</th>
<th>H-plane -3dB Beamwidth (deg)</th>
<th>E-plane -3dB Beamwidth (deg)</th>
</tr>
</thead>
<tbody>
<tr>
<td>WMP54G Dipole</td>
<td>4.57</td>
<td>360</td>
<td>30</td>
</tr>
<tr>
<td>Single Patch Antenna</td>
<td>5.66</td>
<td>70</td>
<td>80</td>
</tr>
<tr>
<td>Measured Patch Array</td>
<td>7.55</td>
<td>48</td>
<td>52</td>
</tr>
<tr>
<td>Expected Patch Array</td>
<td>11.66</td>
<td>45</td>
<td>47</td>
</tr>
</tbody>
</table>

Antenna gain values were calculated from the Friis transmission formula, using anechoic chamber measurements. The separation distance, transmitted power, operating frequency, and cable losses must be measured in order to accurately calculate the gain of an antenna under test. Equations 1 through 4 show the required calculations to derive antenna gain.

\[ P_r(dB) = P_t(dB) + G_r(dB) + G_t(dB) - SL(dB) \]  \hspace{1cm} (1)

\[ SL(dB) = 10 \cdot \log_{10} \left( \frac{4\pi R}{\lambda} \right)^2 = 20 \cdot \log_{10} \left( \frac{4\pi R}{\lambda} \right) \]  \hspace{1cm} (2)

\[ S_{21}(dB) = P_r(dB) - P_t(dB) - CL(dB) \]  \hspace{1cm} (3)

\[ G_r(dB) = S_{21}(dB) - G_t(dB) + CL(dB) + SL(dB) \]  \hspace{1cm} (4)

In the above equations, \( P \) and \( G \) are power and gain, with subscripts \( r \) and \( t \) indicating receive or transmit. \( SL \) and \( CL \) are the spreading loss and cable loss on both transmit and receive sides. Spreading loss results from the electromagnetic waves dispersing away from the send antenna. Spreading loss increases exponentially as a function of the separation distance. \( R \) is the separation distance between the apertures of the send and receive antennas. Table 5 shows a sample antenna gain calculation.
Table 5: Example gain calculation using Patch Array data

<table>
<thead>
<tr>
<th>Measured Data:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Test Frequency (GHz)</td>
</tr>
<tr>
<td>Separation Distance (m)</td>
</tr>
<tr>
<td>Measured S&lt;sub&gt;21&lt;/sub&gt; (dB)</td>
</tr>
<tr>
<td>Cable Loss (dB)</td>
</tr>
<tr>
<td>Transmit Antenna Gain (dB)</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Calculated Values:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lambda (m)</td>
</tr>
<tr>
<td>SL (dB)</td>
</tr>
<tr>
<td>Receive Antenna Gain (dB)</td>
</tr>
</tbody>
</table>

**Single Patch Antenna**

A patch antenna model was created in Ansoft HFSS to calculate expected gain and radiation patterns. Values calculated in the design procedure were used to create the HFSS model. The Design Procedure section in Appendix A includes equations and the design development procedure. The resulting design parameters are shown in Table 6.

Table 6: Patch Antenna Design Variables

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
<th>Term</th>
</tr>
</thead>
<tbody>
<tr>
<td>F</td>
<td>2.437 GHz</td>
<td>Center Frequency</td>
</tr>
<tr>
<td>ε&lt;sub&gt;ε, eff&lt;/sub&gt;</td>
<td>4.093</td>
<td>Effective Dielectric Constant</td>
</tr>
<tr>
<td>W</td>
<td>37.4217 mm</td>
<td>Patch Width</td>
</tr>
<tr>
<td>ΔL</td>
<td>691.96 um</td>
<td>Fringing Length</td>
</tr>
<tr>
<td>L</td>
<td>29.0246 mm</td>
<td>Patch Length</td>
</tr>
<tr>
<td>λ&lt;sub&gt;o&lt;/sub&gt;</td>
<td>123.1 mm</td>
<td>Wavelength</td>
</tr>
<tr>
<td>Yo</td>
<td>9.64 mm</td>
<td>Inset feed distance</td>
</tr>
</tbody>
</table>

HFSS was used to optimize the patch dimensions and probe feed position. The optimization process was used to achieve the correct resonant frequency and to ensure a 50 Ohm match to minimize loss. The design values before and after optimization are shown in Table 7.

Table 7: Initial and Optimized Patch Dimensions

<table>
<thead>
<tr>
<th>Variable</th>
<th>Designed Value</th>
<th>Optimized Value</th>
<th>% Change</th>
<th>Term</th>
</tr>
</thead>
<tbody>
<tr>
<td>W</td>
<td>37.4217 mm</td>
<td>37.5789 mm</td>
<td>0.42 %</td>
<td>Patch Width</td>
</tr>
<tr>
<td>L</td>
<td>29.0246 mm</td>
<td>28.2084 mm</td>
<td>2.81 %</td>
<td>Patch Length</td>
</tr>
<tr>
<td>Probe</td>
<td>9.64 mm</td>
<td>7.5658 mm</td>
<td>21.52 %</td>
<td>Probe feed location</td>
</tr>
</tbody>
</table>
The HFSS 3D model used in this design is shown in Figure 19. Patch impedance characteristics were most sensitive. Small changes in probe location result in large percent changes in input impedance. The patch width changed less than ½% during optimization. Length was optimized to ensure correct resonant frequency at 2.437 GHz.

The optimized design was fabricated using the milling machine in Cal Poly’s chamber lab. A comparison between the HFSS analysis and the measured radiation patterns are shown in Figures 20 and 21.
In Figures 20 and 21, radiation patterns from HFSS simulations and the modified patch antenna are compared. The measured $|S_{21}|$ values were converted to gain using the Friis transmission formula for comparison to simulation results. In the E-plane co-pol plot, $\theta$ is set to 0 degrees and $\phi$ is swept from 0 to 360 degrees. In the H-plane co-pol plot, $\theta$ is set to 90 degrees and $\phi$ is swept from 0 to 360 degrees. In both plots, a $\phi$ value of zero degrees specifies the patch antenna broadside directed at the transmit antenna. The broadside direction is perpendicular to the plane of the substrate.

![Patch Antenna H-Plane Co Pol Radiation Patterns Gain (dB) , f = 2.437 GHz](image)

Figure 21: Comparison of HFSS model and Measured Patch Antenna Radiation Pattern

A significant amount of error is introduced from the testing apparatus behind the antenna. Interference from the testing apparatus behind the antenna contributes to the difference between the expected and measured data. No metal was used in the test fixture for this antenna, minimizing interference as much as possible. For testing hardware, see Figure 16 in Chapter IV.

Figure 22 shows the patch antenna return loss measurement and how it compares to the HFSS model. HFSS simulations predict a 54 MHz -10 dB return loss bandwidth with 2.437 GHz center frequency. The fabricated patch operates at a 2.44 GHz center frequency and has a 63 MHz -10 dB return loss bandwidth. This is a 14.2% improvement over the HFSS bandwidth prediction. Although there was bandwidth improvement, matching at the center frequency suffered. Return loss is about 30 dB higher than expected at the center frequency. Matching is very sensitive to probe feed location. Inaccuracies in drilling the probe feed hole may have caused this discrepancy.
Figure 22: Comparison of Measured and Modeled Patch Antenna Return Loss

The patch antenna measured bandwidth, center frequency and error calculations are shown in Table 8.

Table 8: Comparison of Patch Antenna Design and Measured Values

<table>
<thead>
<tr>
<th></th>
<th>Design Center Frequency</th>
<th>2.437 GHz</th>
<th>Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measured Center Frequency</td>
<td>2.4406 GHz</td>
<td></td>
<td>3.6 MHz (0.15%)</td>
</tr>
<tr>
<td>802.11g Bandwidth</td>
<td>2.95%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Measured Bandwidth (Return Loss = -10 dB)</td>
<td>2.29%</td>
<td>22.37% error</td>
<td></td>
</tr>
<tr>
<td>Measured Bandwidth (VSWR=2)</td>
<td>2.43%</td>
<td>17.63% error</td>
<td></td>
</tr>
</tbody>
</table>

The fabricated patch has a -10 dB return loss bandwidth of 62 MHz. The antenna operating range is from 2.406 GHz to 2.468 GHz. To operate on the full 802.11g frequency range, the antenna must operate from 2.401 to 2.473 GHz, a 72 MHz bandwidth. However, the patch antenna will operate on channels 2 through 10. Only channels 1 and 11 fall outside this antenna’s operating range, leaving channels 2 through 10 available for use.
**Patch Antenna Array**

The antenna array was tested in the anechoic chamber to verify the resonant frequency. The input port has a return loss less than -35 dB across the 802.11g operating range. The array center frequency was measured as 2.442 GHz: see Figure 23 below. Therefore, the adjusted array will operate within the 802.11g operating range of 2.401GHz to 2.473GHz. The adjusted array has a resonant frequency of 2.437 GHz plus a 0.205% error. The array does not require tuning.

Array radiation gain patterns were taken in the anechoic chamber for gain and directivity verification. As discussed in Chapter V, Table 4 quantifies the patch array improvement in gain and beamwidth. Figure 24 shows the measured and expected array radiation patterns. The plots in Figure 24 are normalized to the maximum gain for each measurement so that shape can be considered independently of the magnitude. This shows the similar beamwidth stated above, as well as a similar front to back ratio. The E-plane measurement shows the expected nulls at endfire, but the H-plane does not. The endfire direction is perpendicular to the direction of maximum gain and is in the plane of the antenna substrate, shown in the diagram at ± 90°.
Figure 25 shows that the patch array improved beamwidth and gain over the single patch antenna. The array’s gain at endfire is about 10 dB less than the patch endfire gain, for both H-plane and E-plane. As expected from the array factor calculations, the array emits more radiation behind the ground plane than the single patch. The array factor predicts maximum gain in both directions perpendicular to the substrate.
Figure 25: Measured Radiation Patterns of Patch Array and Single Patch Antenna
VI. Conclusion

Design Difficulties

The first prototype of both the single patch antenna and the array exhibited a resonant frequency higher than the design specification. The prototype patch, with length 29.02 mm, had a resonant frequency of 2.52 GHz. The target antenna operating range is 2.401 GHz to 2.473 GHz. The antenna center frequency was outside of the operating range, causing it to be inoperable for 802.11g communications. The patch length was adjusted using copper tape, tuning the patch resonant frequency. Figure 26 shows the tuned patch antenna. After a return loss measurement verified the 2.437 GHz resonant frequency, a second patch was fabricated with the length of the tuned patch, all other dimensions the same. The second patch, fabricated with the new length, is the final patch design for this project. Its operation is discussed throughout this report.

![Prototype Patch Antenna Showing Copper Tape Modification](image)

The new patch length was determined using the inverse relationship between patch length and resonant frequency. Equation 5 shows the relationship used for patch length modification.

The patch achieves resonance when the length is a half wavelength giving

$$L = \frac{\lambda}{2} = \frac{c}{2f} \therefore \frac{L_1}{L_2} = \frac{f_2}{f_1}$$

Length adjustments are necessary due to dielectric permittivity ($\varepsilon_r$) inaccuracies. The antenna was designed using an $\varepsilon_r$ value of 4.4 for the FR-4 substrate. FR-4 dielectric constant is subject to variations among boards and may also be inhomogeneous in a single board. A dielectric constant measurement, before design begins, will reduce errors in the fabricated design. Using the resonant frequency and length
of the prototype patch, the dielectric constant of the FR-4 was calculated. For comparison, a three inch length, 50 Ohm microstrip thru-line was fabricated to test the FR-4 board dielectric constant. Table 9 shows the FR4 dielectric permittivity calculations.

Table 9: $\varepsilon_r$ Calculations

<table>
<thead>
<tr>
<th></th>
<th>Thru Line</th>
<th>Patch</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant freq (GHz)</td>
<td>1.04</td>
<td>2.532</td>
</tr>
<tr>
<td>L (mm)</td>
<td>76.2</td>
<td>28.117</td>
</tr>
<tr>
<td>$\varepsilon_{r_eff}$</td>
<td>3.5827</td>
<td>4.441</td>
</tr>
<tr>
<td>$\varepsilon_r$</td>
<td>4.75</td>
<td>4.79</td>
</tr>
</tbody>
</table>

Agilent ADS “LineCalc” was used to calculate the dielectric permittivity from the thru-line $\varepsilon_{r\_eff}$ measurements. The substrate height, thru-line width, and operating frequency was entered into LineCalc and the value of “Keff” was calculated. The FR-4 $\varepsilon_r$ value was determined by changing the LineCalc $\varepsilon_r$ value until the calculated Keff equaled the measured $\varepsilon_{r\_eff}$. Also, using measured from the patch antenna $\varepsilon_{r\_eff}$, the FR-4 $\varepsilon_r$ was calculated using Equation 10 in the Design Procedure section of Appendix A.

A microstrip line achieves resonance when the element length is a multiple of a half wavelength in the dielectric

$$L = n \frac{\lambda}{2} = n \frac{c}{2 f_n \sqrt{\varepsilon_{r\_eff}}}$$  \hspace{1cm} (6)

Solving Equation 6 for $f_n$ leads to

$$f_n = n \frac{c}{2 L \sqrt{\varepsilon_{r\_eff}}}$$ \hspace{1cm} (7)

Thus,

$$\varepsilon_{r\_n} = \left( n \frac{c}{2 L f_n} \right)^2$$ \hspace{1cm} (8)

The return loss measurement in Figure 27 was used to calculate the through line resonant frequency in Table 9. Since resonant frequencies $f_n$ repeat in multiples of $f_1$, the lowest resonant frequency ($f_1$) can be extracted from adjacent higher order modes. This relation is shown in Equation 7. The resonant frequencies on the 3 inch through line repeat at intervals of 1.04 GHz. See Figure 27.
Only one FR-4 PCB was tested. Several FR-4 samples should be tested to determine dielectric constant variations among boards. Using both the microstrip thru-line and the patch $\varepsilon_r$ calculations, the dielectric permittivity resulted in only a 0.84% difference. Since multiple calculation methods have resulted in similar results, the relative dielectric constant of the FR-4 board can be assumed to be between 4.75 and 4.79. The $\varepsilon_r$ value of 4.79 is an 8.86% error from the assumed 4.4 value the design began with. This is a significant error and should be addressed in future designs.

**Possible Improvements**

To ensure accuracy in future microstrip antenna design, substrate material characterization should be conducted before the design begins. A 3” by 1” section of substrate can be removed for the test, leaving enough material still available for most projects. Because FR-4 dielectric permittivity is not tightly controlled, a section of each board may be tested prior beginning the design.

In order to avoid repeated dielectric permittivity tests and wasting material with each test, a substrate designed specifically for microstrip applications should be used. Duroid provides accurate and reliable substrate dielectric parameters.

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**Figure 27:** Thru-Line $|S_{11}|$ for $\varepsilon_r$ calculation
References


Appendix

A. Design Procedure

The patch dimensions were chosen by following the design outlined by [3]. Equations 9 through 12 are given in [3] and are used to determine patch width, W, and length, L. Materials and center frequency must be specified. FR-4 material, chosen to be used for this project, has a dielectric constant assumed to be $\varepsilon_r = 4.4$. The FR-4 board thickness, h, is 62 mils. Center frequency is $f_r = 2.437$ GHz (802.11g midband).

$$W = \frac{v_0}{2f_r} \sqrt{\frac{2}{\varepsilon_r + 1}}$$  \hspace{1cm} (9)

$$\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} \left[ 1 + \frac{12h}{W} \right]^{-1/2}$$  \hspace{1cm} (10)

$$\Delta L = h \times 0.412 \left( \frac{\varepsilon_{\text{eff}} + 0.3}{\varepsilon_{\text{eff}} - 0.258} \right) \left( \frac{\frac{W}{\lambda} + 0.264}{\frac{W}{\lambda} + 0.8} \right)$$  \hspace{1cm} (11)

$$L = \frac{v_0}{2f_r \sqrt{\varepsilon_{\text{eff}}}} - 2\Delta L$$  \hspace{1cm} (12)

Figure 28 shows the patch dimensions designed in this procedure. The effective dielectric constant causes fringing at the patch edges. The length L is chosen so that the electrical length of the patch, including fringing, is $\lambda/2$.

![Figure 28: Patch shape, showing design values and fringing area](image)

Equation 9 is used to find a “practical width that leads to good radiation efficiencies” [3]. Using the width calculated in Equation 9, the effective dielectric constant, $\varepsilon_{\text{eff}}$, can be calculated using Equation 10. Fringing effects in the dielectric and air boundary at the patch edges cause an increase of the patch electrical length. The patch length is increased by $\Delta L$ at each radiating edge. This length can be found using Equation 11, then used in Equation 12 to solve for the patch length. The length, L, given in equation 12 provides resonance at the design center frequency.
At the resonant frequency, the input impedance at the patch radiating edge is given by Equations 13 and 14, [3]. Equation 14 assumes $W << \lambda_0$ which is true for this project.

\[ Z_{in} = \frac{1}{2G_1} \]  \hspace{1cm} (13)

\[ G_1 = \frac{1}{90} \left( \frac{W}{\lambda_0} \right)^2 \]  \hspace{1cm} (14)

At resonance the input impedance is purely resistive, there is no reactive component. Using the edge input resistance, a probe feed location is determined. Input resistance at the probe feed location, $R_{in}$, varies with the probe feed location, $y_0$, as shown in Equation 15.

\[ R_{in} = Z_{in} \cos^2 \left( \frac{\pi}{\lambda_0} y_0 \right) \]  \hspace{1cm} (15)

To match the antenna to the 50 Ohm coax line feeding the antenna, $R_{in}$ is set to 50 Ohms and $y_0$ is calculated using Equation 15.