Standalone Antenna Demonstration System

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ABSTRACT

Standalone Antenna Demonstration System

Alexander James Hempy

Antenna systems play a significant role in today’s electronic communications. They are essential for cell phones, satellites, radio, and radar among many other important applications. This paper describes the design, assembly, and operation of an antenna demonstration system designed to instill interest in the field of antenna design among high school and undergraduate college students. The system is portable, supplied solely by DC power supplies, easily reproducible, and includes rotational axes to illustrate antenna performance limitations and requirements. It provides a visual indication of wireless signal strength and demonstrates several antenna performance characteristics including polarization, gain and directivity, radiation patterns - nulls and maximums, and spreading loss. Several antenna types used in present-day applications (embedded and reflector antennas), in addition to structural barriers encountered in typical operating environments, are used to define wireless system performance. Students gain insight on radiating structure and orientation effects on antenna system characteristics and hopefully develop interest in future wireless studies.
ACKNOWLEDGMENTS

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Finally, I extend my thanks to the IEEE Antennas and Propagation Society for hosting and funding the First Ever IEEE Antenna Design Challenge.
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Chapter 1 Introduction

An antenna demonstration system to illustrate “The Magic of Antennas” was developed as a contest entry for the first ever IEEE Antenna Design Competition, sponsored by the Antennas and Propagation Society (AP-S). Three goals were given by the contest committee:

- Design an antenna demonstration system to teach how antennas operate.
- The system must be safe and durable, easily reproducible by others, inexpensive, and portable.
- The total cost for system replication must be less than $1500.

1.1 Antenna Selection

To illustrate antenna operation, four antenna types were selected: dipole, quagi, corner reflector, and patch antennas. These radiators illustrate polarization, radiation patterns, and gain and directivity. Spreading loss, simulated environmental effects on propagating signals, antenna implementation on printed circuit boards, and radiation pattern variations using simulated antenna arrays (corner reflector) are also demonstrated. Each antenna is straightforward to build and requires inexpensive parts available at local hardware stores or online vendors.

The transmit antenna is a dipole that rotates in the roll and azimuth directions; and is used in all demonstrations. An identical, but stationary, receive dipole antenna is used in conjunction with the rotating transmit dipole to demonstrate polarization and radiation pattern nulls. A $\lambda/2$ dipole radiates equally in the plane normal to the dipole axis, which can be verified by adjusting the rotating dipole azimuth.
A Quad and Yagi-Uda antenna combination, called a Quagi was selected to demonstrate gain and directionality. The Quagi is simple to build, provides gain beyond that of a dipole antenna, focuses energy in one direction, and yields performance comparable to a dipole Yagi-Uda antenna. The Quagi antenna rotates azimuthally to demonstrate H-plane half-power beamwidth.

A corner reflector was selected to demonstrate how gain and beamwidth vary with the number of effective elements (image theory) and element spacing. The platform rotates in the H-plane to measure radiation pattern and gain differences with respect to corner angle. The dipole antenna fixture allows spacing adjustment to demonstrate array factor maximums and nulls for each corner angle.

An embedded patch antenna illustrates implementation on printed circuit boards that include additional circuitry. The patch antenna exhibits a wider half-power beamwidth and greater gain compared to dipole antennas.

Four environmental barriers are used to simulate obstructions encountered by propagating signals. Electrostatic discharge material and copper mesh resemble barriers encountered in electronics manufacturing environments and are expected to substantially attenuate the received signal. Drywall and poultry net resemble residential walls; less attenuation is expected. Spreading loss is demonstrated with each antenna configuration by increasing the separation distance. Polarization effects are demonstrated by adjusting the rotating dipole roll.
1.2 Receiver and Transmitter Design

The system operates at 915MHz within the 902-928MHz (33cm) amateur band. This allows for relatively small antennas and eliminates requirements for components that operate above ultra-high frequencies (UHF). The system includes an oscillator, two RF amplifiers, and a bandpass filter. The signal strength indicator includes a peak detector, DC amplifier, and bar graph driver and visually displays received signal strength. The system block diagram is shown in Figure 1.1.

![System block diagram](image)

1.2.1 Component Requirements

A 0dB worst case antenna gain was used to calculate component requirements. The expected RF amplification is 15dB, while the expected oscillator output power is -2dBm from experience with similar models designed at Cal Poly.

Relatively low amplitude peak detected signals are amplified by a non-inverting op amp to adjust the bar graph display DC voltage and utilize the full LED range. To provide sufficient dynamic range, a minimum -6dBm is required at the signal strength indicator input based on a previously constructed peak detector.

A 3’ transmit/receive separation distance ensures that all antennas operate in the far-zone. The spreading loss for 3’ separation at 915MHz is 30.89dB. With the expected
oscillator output power, antenna and amplifier gains, and spreading loss, -2.89dBm will be applied to the bandpass filter. To provide a minimum -6dBm to the peak detector, the bandpass filter insertion loss must be less than 3.11dB. To attenuate signals outside the amateur band, 3dB minimum attenuation is required at 902MHz and 928MHz.

### 1.3 Contest Submissions

A two page proposal that includes an objective, preliminary design, and a bill of materials was due on October 1, 2009, see Appendix B. The 10 page final report that includes a detailed design description, list of parts and materials, photos, and assembly and operation instructions was due on May 1, 2010. The parts list is given in Appendix C. The final system is shown in Figure 1.2.

![Figure 1.2. Actual Wireless System.](image)
Chapter 2 Transmitter Design

2.1 Oscillator

A negative resistance oscillator modified from [4] is used to generate a 915MHz RF signal, see Figure 2.1. The oscillator provides -1.5 dBm when supplied with 9V DC input. The oscillator couples energy into a λ/2 10AWG copper wire resonator to generate oscillations. The schematic is given in Figure 2.2.
2.1.1 Negative Resistance

The oscillator operates on the principle of negative resistance, an unstable condition required to initiate oscillation. Negative resistance provides a loop gain greater than unity to increase signal amplitude. The initial signal is provided by thermal noise in the resonator. As the signal amplitude increases, the negative resistance magnitude decreases due to transistor gain roll-off to stabilize the RF signal. To create a negative input resistance, inductance is required at the transistor base.

To verify negative resistance at the emitter input, consider a small signal model of a common base transistor amplifier with a grounded inductor at the base, see Figure 2.3.

![Common base amplifier with grounded base inductor: small signal model.](image)

The impedance at the emitter input is [5]

\[
Z_{in}^E = \frac{r_\pi + j\omega L}{\beta(s) + 1}
\]

(2.1)

\(\beta(s)\) is defined as [5]

\[
\beta(s) \approx \frac{\beta_0}{1 + \frac{s}{\omega_\beta}}
\]

(2.2)

The magnitude and phase of \(\beta(s)\) for a typical transistor is plotted in Figure 2.4. If the amplifier is operated above \(10\omega_\beta\) and provides gain, a negative input resistance is realized. This is illustrated in equation (2.3).
At frequencies greater than $10\omega_B$, $\beta = -j|\beta(s)|$. Assuming $j\omega L \gg r_\pi$ and $-j|\beta| + 1 \approx -j|\beta|$ eq. (2.1) becomes

$$Z_{in}^E = \frac{j\omega L}{-j|\beta|} = \frac{-\omega L}{|\beta|} \tag{2.3}$$

which indicates a negative resistance at the emitter input. The BFR91 transistor was replaced with the Avago AT-41486 transistor since the BFR91 is now obsolete. The AT-41486 transistor provides gain up to $f_T \approx 8$GHz.

### 2.1.2 Assembly

The oscillator is assembled on 62 mil FR4 laminate. The components are soldered above the ground plane and normal to the resonator to minimize coupling with the resonator. Sufficient inductance (8.9nH) at the transistor base is achieved using a minimum 0.5” length lead, see Figure 2.2. Resonance occurs in a $\lambda/2$ (6.38”) 10AWG
solid copper wire coupled to the AT-41486 (Q1) transistor emitter lead, by a lead positioned between the resonator and ground plane at one end of the resonator, see Figure 2.5.

The coupling wire should be placed within 1” of the end; the resonator is shorted at each end where minimum voltage and maximum current exists. The copper wire is located approximately 0.1” above the ground plane.

**2.1.3 Operation**

The absolute maximum collector-emitter voltage is 12V DC. To prevent transistor damage, the oscillator is operated at 9V DC max. The oscillator peak output power vs. input DC voltage is plotted in Figure 2.6. The oscillator provides -1.5dBm at 9V DC.
2.2 RF Amplifier

The transmit (TX) RF amplifier milled on FR4 substrate provides approximately 15dB gain, see Figure 2.7. Matching networks minimize reflections and maximize amplifier efficiency.

Figure 2.6. Oscillator peak output power vs. input voltage.

Figure 2.7. A 15 dB gain microstrip amplifier.
2.2.1 Optimization

To optimize amplifier performance, transistor S-parameters were measured from 900MHz to 930MHz, ±15MHz about the 915MHz operating frequency. The amplifiers were modeled in ADS using the measured S-parameters. A matching network using 50Ω (121mil) open circuit stubs was designed to maximize gain and was optimized in ADS. The layout dimensions are given in Figure 2.8. The TX amplifier simulated and measured gain are shown in Figure 2.9.

![Figure 2.8. Amplifier layout dimensions.](image-url)
2.2.2 DC Biasing

The smallest allowable line width (15mil) high impedance ($Z_0 = 119.3\,\Omega$) quarter-wave lines are added to DC bias the high frequency AT-41435G transistor. The bias lines are AC shorted to ground through coupling capacitors near the biasing circuitry to create an open circuit at both RF transistor ports and minimize noise at the DC port, see Figure 2.10.
The amplifier requires 5V DC applied across the collector-emitter. The base-emitter junction is biased by adjusting potentiometer R1. Base-emitter voltage is directly proportional to the voltage drop across R1, see Figure 2.11, and requires approximately 0.7V DC for operation.

![Figure 2.11. Amplifier bias circuit schematic.](image)

### 2.2.3 Assembly

The amplifiers are milled on 62 mil thick, 2.5” x 5” FR4 laminate, see Figure 2.7. The board includes solder pads for DC biasing. The hand-made 60.5nH inductors are 0.7” in length and composed of 24AWG wire formed into 13 0.1” diameter loops. The inductance of a coil of wire is calculated using equation 2.4 [8] where N is the number of loops, \( \mu \) is the permeability of the inductor core, A is the cross-sectional area in meters, and l is the length in meters.

\[
L = \frac{N^2 \mu A}{l}
\]

A magnified view of the actual biasing circuitry including solder pads and wound inductors is shown in Figure 2.12.

![Figure 2.12. Amplifier bias circuit.](image)
Chapter 3 Receiver Design

3.1 RF Amplifier

The receiver also includes an RF amplifier to increase signal strength. The receive (RX) RF amplifier is identical to the TX amplifier. The RX amplifier provides approximately 0.8dB more gain than the TX amplifier due to individual transistor characteristics. Simulated and measured RX amplifier gain are shown in Figure 3.1.

![RX Amplifier Gain](image)

**Figure 3.1. RX amplifier gain.**

3.2 Bandpass Filter

A bandpass filter prevents signal reception at undesired frequencies. Several bandpass filter designs were considered including quarter-wave coupled resonator, capacitively coupled quarter-wave resonator, and cascaded lowpass and highpass filters using transmission line stubs. The parallel-coupled half-wavelength resonator bandpass filter was selected since it does not require lumped components, a detailed design method
is available [9], and it exhibits relatively low insertion loss compared to other filters, see Figure 3.2. Lumped components should be avoided at microwave frequencies as unavailable component values are frequently required.

![Parallel-coupled bandpass filter.](image)

### 3.2.1 Design

The coupled-line bandpass filter was designed using relations in [9] and optimized in ADS. The design process from [9] is as follows.

The characteristic admittances of coupled sections are calculated using [9]

\[
\frac{J_{01}}{Y_0} = \frac{\pi \text{FBW}}{\sqrt{2} g_0 g_1} \quad \text{(3.1)}
\]

\[
\frac{J_{j,j+1}}{Y_0} = \frac{\pi \text{FBW}}{2} \frac{1}{\sqrt{g_j g_{j+1}}} \quad \text{(3.2)}
\]

where \( j = 1 \) to \( n-1 \), \( n \) is the filter order, and

\[
\frac{J_{n,n+1}}{Y_0} = \frac{\pi \text{FBW}}{\sqrt{2 g_n g_{n+1}}} \quad \text{(3.3)}
\]

where the fractional bandwidth, FBW, is defined as

\[
\text{FBW} = \frac{\omega_2 - \omega_1}{\omega_0} \quad \text{(3.4)}
\]

and \( g_0, g_1, \) etc. are elements of a ladder-type lowpass filter with a unity normalized cutoff frequency and are available in tabular form [9]. \( Y_0 \) is the characteristic admittance.
The required even mode and odd mode characteristic impedances of the coupled line resonators are calculated using [9]:

\[
(Z_{0e})_{j,j+1} = \frac{1}{Y_0} \left[ 1 + \frac{J_{j,j+1}}{Y_0} + \left( \frac{J_{j,j+1}}{Y_0} \right)^2 \right] 
\]  

(3.5)

\[
(Z_{0o})_{j,j+1} = \frac{1}{Y_0} \left[ 1 - \frac{J_{j,j+1}}{Y_0} + \left( \frac{J_{j,j+1}}{Y_0} \right)^2 \right] 
\]

(3.6)

with \( j = 0 \) to \( n \).

Coupled line section lengths and widths are calculated. The next series of equations yields characteristic odd and even mode impedances for a given width and interelement spacing of a coupled line section. The dimensions are adjusted until the coupled section even and odd mode impedances match the impedances found using (3.5) and (3.6). Relations for calculating the coupled section impedance for a given length and width are listed in Appendix A. The actual length of each coupled line section is given by

\[
I_j = \frac{\lambda_0}{4 \left( \sqrt{(\varepsilon_{re}^e)_j \times (\varepsilon_{re}^o)_j} \right)^2} - \Delta I_j 
\]

(3.7)

where \( \varepsilon_{re}^e \) and \( \varepsilon_{re}^o \) are the even and odd mode effective dielectric constants, respectively, (see Appendix A). Fields extend beyond the open end of a microstrip line which is accounted for by [9]

\[
\Delta I = \frac{c Z_c C_p}{\sqrt{\varepsilon_{re}}} 
\]

(3.8)

where \( c \) is the speed of light in free space, and

\[
C_p = \varepsilon_e \varepsilon_r W / h 
\]

(3.9)
is the equivalent shunt capacitance of an equivalent length transmission line $\Delta l$ [9].

Even and odd mode impedances for a 0.1dB ripple Chebyshev 2nd order lowpass prototype filter are listed in Table 3.1. The coupled section widths, lengths, and separation distances are listed in Table 3.2.

Table 3.1. Coupled section even and odd mode impedances.

<table>
<thead>
<tr>
<th>$j$</th>
<th>$J_{j+1}/Yo$</th>
<th>$Z_{oe,j+1}$ (Ω)</th>
<th>$Z_{oo,j+1}$ (Ω)</th>
</tr>
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<tbody>
<tr>
<td>0</td>
<td>0.247</td>
<td>65.413</td>
<td>40.696</td>
</tr>
<tr>
<td>1</td>
<td>0.071</td>
<td>53.809</td>
<td>46.697</td>
</tr>
<tr>
<td>2</td>
<td>0.056</td>
<td>52.962</td>
<td>47.352</td>
</tr>
</tbody>
</table>

Table 3.2. Coupled line section dimensions: width, separation, and length.

<table>
<thead>
<tr>
<th>$j$</th>
<th>$W_j$ (mm)</th>
<th>$S_j$ (mm)</th>
<th>$l_j$ (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>2.010</td>
<td>0.255</td>
<td>0.05899</td>
</tr>
<tr>
<td>1</td>
<td>2.294</td>
<td>1.530</td>
<td>0.05810</td>
</tr>
<tr>
<td>2</td>
<td>2.273</td>
<td>1.705</td>
<td>0.05811</td>
</tr>
</tbody>
</table>

The filter was modeled and optimized in ADS and optimized. The resultant circuit model is shown in Figure 3.3; final dimensions are in Figure 3.4.

Figure 3.3. ADS bandpass filter circuit model.
Simulated results, shown in Figure 3.5, indicate a 0.12dB insertion loss at 915MHz and a 20MHz (2.19%) 3dB bandwidth, the frequency span over which $|S_{21}|$ is within 3dB of the peak magnitude.

![BPF Insertion Loss vs. Frequency](image)

**Figure 3.5. Measured and ADS simulated bandpass filter insertion loss vs. frequency.**

Experimental results, shown in Figure 3.5, indicate 2.67dB insertion loss at 915MHz and a 30MHz 3dB bandwidth (3.28%). The insertion loss meets the 3.11dB maximum specified in Chapter 1.
3.2.2 Assembly

The bandpass filter was milled on a 31 mil thick, 1” x 7.75” RT Duroid 5870 laminate. Two female SMA edge-mount connectors are soldered to the trace and the ground plane on both ends of the filter.

3.3 Signal Strength Indicator

The signal strength indicator is composed of an RF peak detector, an adjustable-gain non-inverting operational amplifier circuit, an LED bar graph driver, and LED array; see Figure 3.6. The peak detector converts the received RF signal to a DC voltage which is proportional to the received signal strength, which is applied to a non-inverting operational amplifier circuit for range adjustment. A bar graph display circuit drives a 30 LED array. The number of illuminated LEDs is proportional to the peak detector voltage and received RF signal strength.

The adjustable gain compensates for variation among RX antennas. When the TX and RX antennas are positioned for maximum transmission and reception, all 30 LEDs should be illuminated.

Figure 3.6. The signal strength indicator illuminates LEDs proportional to the received RF signal strength.
Figure 3.7. The signal strength indicator is comprised of Schottky diodes, LM3914 bar graph display ICs, op amps, and passive components.
3.3.1 Peak Detector

The peak detector is a rectifier which uses two HSMS-2822 Schottky diodes to rectify the incoming RF signal. The signal is applied to a 1000pF capacitor to filter the AC component and yield a DC voltage. A 100kΩ resistor is placed in parallel to discharge the capacitor when the input level decreases. Hence, the DC output voltage responds to RF amplitude variations for radiation pattern measurements. The schematic is given in Figure 3.8.

![Figure 3.8. The peak detector: Shottky diodes, capacitor, resistor.](image)

The Schottky diodes operate up to 1.5GHz and have a 12V reverse breakdown voltage that accommodates a maximum peak detector voltage of 3.75V. The full-scale bargraph driver input voltage (all 30 LEDs) is 3.75V. The peak detector DC output voltage is sufficient to illuminate all LEDs without amplification when 11dBm is supplied at the input. A plot of DC output voltage vs. RF input power is shown in Figure 3.9
Peak detector components are soldered onto a 1” x 1” 62mil FR4 laminate. The input is applied through a male SMA connector and 0.085” semi-rigid coax line soldered onto the FR4 substrate. A vectorboard contains the bar graph display, DC amplifier, and includes the FR4 circuit.

### 3.3.2 DC Amplifier

A gain-adjustable non-inverting DC amplifier is included between the peak detector and the bar graph display driver to provide gain adjustment. This allows receiver tuning for full-scale LED illumination when antennas are positioned for maximum signal reception. The amplifier also allows for antenna separation distance variations.

![Graph showing peak detector DC output voltage vs. input power.](image)

*Figure 3.9. Peak detector DC output voltage vs. 915MHz RF input power.*
The amplifier is comprised of LM324 op-amps which require a positive supply voltage, see Figure 3.10. The gain is adjusted by varying R5, a 1MΩ potentiometer. The amplifier gain provided by the amplifier is given by

\[
\frac{V_o}{V_{in}} = 1 + \frac{R5}{R4}
\]  

(3.10)

which provides a maximum gain of 100. This reduces the minimum required RF input peak detector power to illuminate all 30 LEDs to approximately -7dBm.

The amplifier includes a voltage follower at the amplifier input to reduce loading effects on the peak detector. The voltage follower output is applied to a R3 which should equal the parallel combination of R4 and R5 to minimize input offset voltage errors caused by input bias current.

![Non-inverting DC amplifier schematic.](image)

Figure 3.10. Non-inverting DC amplifier schematic.

The components are soldered to a 3.6” x 7” protoboard. The amplifier input is connected to the peak detector output and the amplifier output drives the bar graph display driver.
3.3.3 Bar Graph Display Driver

The bargraph display driver is composed of three LM3914 Dot/Bar display drivers. Each IC drives up to ten LEDs. The schematic was obtained from [10] and modified to cascade three LM3914s, see Figure 3.11.

The LM3914s are configured for bar mode to provide an enhanced visual indication of received signal strength. The circuit also includes a 2uF tantalum decoupling capacitor across the voltage supply which prevents oscillations caused by long wires from the supply to the common LED anode port [10].

The maximum input voltage is 35V [10] which is greater than the maximum 6V that can be supplied by the non-inverting op amp. The maximum power dissipation for each IC is 1365mW [10]. The circuit is designed to drive each LED with 5.7mA which results in 342mW dissipated per IC, well below the maximum allowed power dissipation. The current drawn through each LED in amperes is given by [10]

\[ I_{LED} \approx \frac{12.5}{R6} \]  

(3.11)

which is approximately 5.7mA with R6 = 2200Ω (note: R7 and R8 determine the current per LED in U3 and U4, respectively).
LEDs are illuminated in proportion to the DC input voltage. The step size for each additional illuminated LED is 125mV. At 0V DC, no LEDs are illuminated and at 3.75V DC all LEDs are illuminated. Bar graph driver linearity is shown in Figure 3.12.

Figure 3.12. The bargraph driver has a linear output.
The circuit is soldered to the vectorboard that includes the DC amplifier and peak detector. The ICs are positioned in-line with one another to minimize lead lengths. The LEDs are mounted in-line to emulate a signal strength ‘bar,’ see Figure 3.6.
Chapter 4 Antenna and Barrier Design & Assembly

4.1 Dipole Elements

4.1.1 Design

The dipole elements are designed for 50Ω input impedance. The outer and inner diameters are 3/8” and 5/32”. Selected tube dimensions, yield a 52.5Ω characteristic impedance calculated using equation 4.1 [6] where $b$ and $a$ are the outer and inner radii, respectively.

$$Z_0 = \sqrt{\frac{\mu_0}{\varepsilon_0} \frac{\ln(b/a)}{2\pi}}$$  \hspace{1cm} (4.1)

The dipole incorporates a split-type balun to minimize surface current radiation on the outer brass tube. The balun is created by cutting two $\lambda/4$ slots down the transmitting side of the dipole, see Figure 4.1. One dipole arm is AC shorted to the slot which creates an AC open $\lambda/4$ from the elements at the start of the balun, see Figure 4.1. The total electrical distance between the two copper elements is 180° resulting in radiation from the copper elements.
4.1.2 Assembly

The dipoles used in the stationary, rotating, and corner reflector antennas are composed of 5/32” and 3/8” diameter brass tubing, 10AWG copper wire, and RG174 SMA male coaxial cable, see Figure 4.1. Two 3.23” length slots (λ/4 at 915MHz) are cut into the 3/8” brass tubing using a Dremel cutting wheel to form a split balun. 1/8” diameter holes are drilled at one end of both the 3/8” and 5/32” brass tubes to accommodate one dipole arm, see Figure 4.2. The other dipole arm is soldered to the surface of the split balun. The initial 4” length dipole arms are trimmed in 1/32” increments until minimum |S₁₁| occurs at 915MHz.
Figure 4.2. Dipole elements are supported by split baluns.

The coaxial feed line is stripped and the center conductor is soldered to the 5/32” inner brass tube, whereas the outer coax conductor is soldered to the 3/8” outer brass tube. Nonconducting hot glue is applied inside the dipole end to mechanically stabilize the tip of the split balun, see Figure 4.3. The coaxial feed lines are approximately 2’ long for the stationary and rotating dipole antennas, and approximately 3’ for the corner reflector to allow for repositioning. The stationary and rotating dipole antenna tubing is 7” in length; the corner reflector, 20”.

Figure 4.3. Dipole feed connection and hot glue support.
4.2 Rotating Dipole

Dipole antennas are commonly used in wireless communications and are straightforward to assemble, and are well-documented in introductory antenna textbooks. Rotating and stationary dipole antennas are included to demonstrate dipole radiation patterns and polarization effects.

A rotating 915MHz transmit dipole antenna is used for all demonstrations, see Figure 4.4. Polarization effects and radiation pattern characteristics (including nulls) are illustrated for each receive antenna using the rotating transmit dipole.

![Figure 4.4. Rotating transmit dipole used for the entire demonstration.](image)

4.2.1 Design

The rotating transmit dipole antenna is mounted on two Lazy Susan bearings to enable rotation in two orthogonal directions; azimuth and roll. These rotational axes are
used to illustrate radiation pattern (received signal strength vs. incident angle) and polarization (TX/RX antenna alignment) effects, respectively.

4.2.2 Assembly

The rotating dipole mounting base is composed of wood components assembled using wood glue, except the two Lazy Susans, which are secured by wood screws. All dimensions are defined in Figure 4.5. A 13/32” hole accommodates the dipole. The
dipole feed is soldered to the brass tubing after it is inserted into the base. The mounting post is composed of two 1” thick 2” wide posts glued together as shown in Figure 4.5.

4.3 Stationary Dipole

The stationary dipole is identical to the rotating dipole without Lazy Susans and is used in the first demonstration, see Figure 4.6. The antenna is vertically polarized and mounted to a wooden base.

![Stationary dipole antenna](image)

Figure 4.6. Stationary dipole antenna: illustrates polarization and radiation pattern effects.

4.3.1 Design

The base is designed to position the dipole at the same height as the transmit rotating dipole antenna. Construction requires only wood, wood glue, and a single wood screw placed from the bottom of the base into the post. Stationary dipole base dimensions are given in Figure 4.7.
4.4 Quagi

The Quagi antenna is used to demonstrate directionality: how radiation is focused in a specific direction. The Quagi is a variation of the Yagi-Uda antenna (parasitic array with a single driven element) that substitutes square loops for dipole elements, see Figure 4.8. The Quagi antenna includes one active (driven) element and three parasitic elements - one reflector and two directors - which focus radiation from the driven element toward the directors. This results in a gain (in the maximum radiation direction) 1.84 dB greater than the stationary dipole antenna.
4.4.1 Design

The Quagi antenna was designed to allow rotation in the H-plane to demonstrate radiation patterns. Similar to the dipole antennas, the base was designed using wood and a Lazy Susan for straightforward construction, portability, and rotation capability.

Element perimeter lengths in feet are given by [11]

\[
\text{Reflector length} = \frac{1046.8}{f_{\text{MHz}}} \quad (4.2)
\]

\[
\text{Driven element length} = \frac{985.5}{f_{\text{MHz}}} \quad (4.3)
\]

\[
\text{Director length} = \frac{937.3}{f_{\text{MHz}}} \quad (4.4)
\]

Total element perimeter lengths and length per side values are given in Table 4.1.
Table 4.1. Quagi element lengths.

<table>
<thead>
<tr>
<th>Element</th>
<th>Element Length (in)</th>
<th>Side Length (in)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reflector (R)</td>
<td>13.729</td>
<td>3.432</td>
</tr>
<tr>
<td>Driven Element (DE)</td>
<td>12.925</td>
<td>3.231</td>
</tr>
<tr>
<td>Directors (D)</td>
<td>12.292</td>
<td>3.073</td>
</tr>
</tbody>
</table>

Interelement spacing for 915MHz was determined by frequency scaling a 144MHz Quagi [11]. The separation distances for the 915MHz Quagi are given in Table 3.1.

Table 4.2. Quagi antenna interelement spacing distances.

<table>
<thead>
<tr>
<th>Interelement Spacing (in)</th>
</tr>
</thead>
<tbody>
<tr>
<td>R-DE:</td>
</tr>
<tr>
<td>DE-D1:</td>
</tr>
<tr>
<td>D1-D2:</td>
</tr>
</tbody>
</table>

The driven element is fed on the side to vertically polarize the antenna. The current distribution is shown in Figure 4.9 which indicates radiating vertical sides. The horizontal sides are non-radiating due to zero net current distributions.

Figure 4.9. Driven element current distribution.
4.4.2 Assembly

The Quagi antenna is composed of a 5/16” wooden dowel, wood, and glue. All dimensions are defined in Figure 4.10. The boom is secured to the stand using a 5/16” wooden dowel.

The Quagi elements are squares formed from 10AWG solid copper wire. The directors, driven element, and reflector have 3.08”, 3.23”, and 3.43” sides, respectively. The reflector and directors are soldered together to form closed squares. The driven element is fed by RG174 coax on the side, see Figure 4.11. The gap dimension is 1/32” determined experimentally.

Figure 4.10. Quagi antenna base dimensions.
The Quagi may also require tuning. The driven element side dimension is initially 0.2” greater than specified. File down one end of the driven element until $|S_{11}|$ is minimized at 915MHz. Maintain a 1/32” gap between the copper ends, see Figure 4.11.

Figure 4.11. The Quagi antenna is fed on the vertical side by a coaxial cable.

### 4.5 Corner Reflector

A corner reflector is used to demonstrate a simulated dipole antenna array, see Figure 4.12. Through image theory, the corner reflector illustrates how the number of effective elements and element spacing affect antenna gain and beamwidth.

Figure 4.12. Adjustable corner reflector antenna: simulates antenna arrays.
4.5.1 Design

The corner reflector includes slots positioned for 90°, 60°, and 30° corners that simulate antenna arrays with 4, 6, and 12 elements, respectively: see Figure 4.13. Element one represents the actual source for each corner configuration.

A dipole antenna, identical to the stationary and rotating dipoles, is used to receive signals with the corner reflector. The dipole images are required to satisfy the tangential E-field boundary conditions at the conducting planes. The support structure allows antenna positioning for optimum transmission with different corner angles. The corner reflector is mounted on a Lazy Susan which allows rotation to demonstrate simulated array radiation patterns.

![Figure 4.13. Left to right: 90°, 60°, and 30° corners including images.](image)

Field strength peaks occur in the maximum radiation direction with different dipole spacing. For 90° corner reflectors, normalized field strength peaks occur at λ intervals because the radiation from elements 1 and 3 in add in phase, see Figure 4.13. Elements 2 and 4 are equidistant to the far zone in the maximum radiation direction and
also add in-phase. Minimums occur in odd multiples of a half wavelength since the radiation will cancel between elements 1 and 3.

Field strength peaks occur at $2\lambda$ intervals for 60° corners which can be calculated using similar analysis [3]. For 30° corners, the field strength peak is “almost periodic” since the arguments of the trigonometric functions representing the array factors are related by irrational numbers [3].

### 4.5.2 Assembly

Cable ties and mounts secure the dipole and allow repositioning with the different corner angles, see Figure 4.14.

![Figure 4.14. Corner reflector adjustable dipole mount.](image)

For a 90° corner, the dipole is placed approximately 6.45” (0.5\(\lambda\)) from the vertex to achieve peak received field strength [3]. The dipole is placed 8.39” (0.65\(\lambda\)) and 15.49” (1.2\(\lambda\)) from the vertex for 60° and 30° corner angles, respectively [3]. The element spacing for the first maximum at each corner angle is listed in Table 4.3.

<table>
<thead>
<tr>
<th>Corner Angle</th>
<th>Spacing from vertex (in)</th>
<th>Spacing from vertex ((\lambda))</th>
</tr>
</thead>
<tbody>
<tr>
<td>90°</td>
<td>6.45</td>
<td>0.50</td>
</tr>
<tr>
<td>60°</td>
<td>8.39</td>
<td>0.65</td>
</tr>
<tr>
<td>30°</td>
<td>15.49</td>
<td>1.20</td>
</tr>
</tbody>
</table>
All corner reflector dimensions are defined in Figure 4.15. The 22” x 22” platform is attached to a 10” x 22” wooden base via a Lazy Susan.

![Figure 4.15. Corner reflector dimensions.](image)

### 4.6 Embedded Patch

The embedded printed circuit board (PCB) patch antenna is used to illustrate a scaled version of current cell phone antenna technology and the advantages of microstrip antennas: low profile, ease of fabrication, low cost, and PCB compatibility with other circuitry. The patch antenna is milled alongside a 915MHz amplifier and is matched to a 50Ω transmission line via a quarter-wave transformer, see Figure 4.16.
4.6.1 Design

The patch length was calculated using equation (4.5) found in [2] where $\lambda$ is the operating wavelength in air and $\varepsilon_r$ is the relative dielectric constant of the laminate. The patch length for $\lambda = 0.327$ m and $\varepsilon_r = 2.33$ is 4.14” from eq. (4.5).

$$L \approx 0.49 \frac{\lambda}{\sqrt{\varepsilon_r}} \tag{4.5}$$

Because the maximum allowable printed circuit board size is 9” x 12”, a 5” patch width was selected to allow 2” margins from the PCB edges and to minimize the patch input resistance.
The patch input resistance was found using Ansoft HFSS. A quarter wave transformer was designed to match the 50\(\Omega\) transmission line to the patch antenna input resistance. The quarter wave line impedance was calculated using equation (4.6) found in [1] with \(Z_0=50\Omega\) and \(R_L \approx 150\Omega\). The microstrip linewidth (36 mil) and trace length (2362 mil) for the quarter wave line impedance \((Z_1 = 86.6\Omega)\) were found using ADS LineCalc.

\[
Z_1 = \sqrt{Z_0 R_L}
\]  

(4.6)

Miter bends were implemented to connect the embedded RX patch antenna output to the RF amplifier input. Curved bends have an effective length less than the centerline which creates path length error. In addition, curved bends also add shunt capacitance which can affect sensitive high frequency circuits [7]. 90° bends also add shunt capacitance, however applying miter bends reduces the capacitance. This restores the line back to its original characteristic impedance. Miter bend dimensions are given in Figure 4.17. Relations for dimension calculations [7] are given in eqs. (4.1) to (4.3).

![Figure 4.17. Miter bend dimensions.](image-url)
\[ D = W \cdot \sqrt{2} \]  
\[ X = D \cdot (0.52 + 0.65 \exp \left(-1.35 \cdot \left(\frac{W}{h}\right)\right) \]  
\[ A = \left(X - \frac{D}{2}\right) \cdot \sqrt{2} \]

where \( W/h \) is the laminate width to height ratio, and

\[ 4.6.2 \text{ HFSS Model} \]

The HFSS patch antenna model is shown in Figure 4.18. Initial patch dimensions are taken from section 4.6.1. The patch length was adjusted for 915MHz operation. The return loss is shown in Figure 4.19. The model predicts a -26dB return loss at 915MHz.
Figure 4.19. Simulated patch antenna return loss.

The H- and E-plane radiation patterns are shown in Figures 4.20 and 4.21, respectively. The patch antenna model predicts a 4.51dB maximum gain normal to the patch (z direction): see Figure 4.18. The half power beamwidth, the angular range where the transmitted power is within 3dB of the peak, is approximately 82° and 77° in the H- and E-planes, respectively.
Figure 4.20. Simulated H-plane patch antenna radiation pattern.

Figure 4.21. Simulated E-plane patch antenna radiation pattern.
4.6.3 Assembly

The embedded PCB patch antenna was milled and the amplifier redesigned for operation on 31 mil thick, 9” x 12” RT Duroid 5870 laminate. The amplifier uses the FR4 version biasing circuit since the same transistor was used. The ADS layout is shown in Figure 4.22.

![Figure 4.22. Embedded patch antenna layout.](image)

4.7 Environmental Barriers

4.7.1 Design

Environmental barriers simulate actual wireless system environments and signal attenuation effects. The environmental barriers are represented by four 2’ x 2’ frames composed of drywall, poultry net, electrostatic discharge (ESD) material, and copper
mesh; see Figure 4.24. The drywall and poultry net represent building walls, while the ESD and copper mesh frames simulate electronics manufacturing environments.

The poultry net and copper mesh barriers can be modeled as waveguides. The waveguide dimensions are shown in Figure 4.23. The copper mesh and poultry net hole dimensions are given in Table 4.4.

![Figure 4.23. Waveguide dimensions.](image)

<table>
<thead>
<tr>
<th></th>
<th>a (m)</th>
<th>b (m)</th>
<th>z (λ)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Poultry Net</td>
<td>0.0381</td>
<td>0.0254</td>
<td>2.4 x 10^{-3}</td>
</tr>
<tr>
<td>Copper Mesh</td>
<td>0.000787</td>
<td>0.000635</td>
<td>1.94 x 10^{-3}</td>
</tr>
</tbody>
</table>

The TE_{100} attenuation is calculated using relations from [14]. The propagation constant in the z-direction for the TE_{100} mode is

\[ \beta_z = \pm \sqrt{\beta^2 - \beta_x^2 - \beta_y^2} \]  

(4.10)
where

$$\beta = \frac{2\pi f}{c}$$  \hspace{1cm} (4.11)

$f$ is the TE$_{100}$ mode frequency in Hz and $c$ is the speed of light in meters per second.

For the TE$_{100}$ mode, $\beta_y = 0$ and $\beta_x = \pi/a$ where $a$ is the width of the wave guide in meters.

The attenuation in the $z$-direction is

$$\alpha = e^{-j\beta_z z'}$$  \hspace{1cm} (4.12)

where $z'$ is the waveguide electrical length in the $z$ axis. The poultry net and copper mesh barriers attenuate a 915MHz signal -1.67dB and -22.06dB, respectively.

Figure 4.24. Environmental barriers simulate actual wireless system environments. From left to right: ESD, drywall, copper mesh, and poultry net barriers.

4.7.2 Assembly

The environmental barrier frames are assembled with 0.5” x 0.75” wood beams 2’ in length and nailed together at the corners. Wood staples are placed at each corner joint for increased strength.
The ESD barrier is assembled by cutting and taping 6”x10” bags into a 2’ x 2’ square and attaching it to the wood frame using a staple gun. The copper mesh barrier includes four copper mesh 2’ x 6” strips extended across the wood frame and also secured by staples, see Figure 4.24. The poultry net barrier includes one 2’ x 2’ poultry net section stapled to the frame. The drywall barrier is available pre-cut in a 2’ x 2’ sheet, 0.5” thick.
Chapter 5 Antenna Testing

To measure radiation patterns and patch antenna absolute gain, two identical patch antennas were milled. Patch antennas were selected for ease of fabrication through milling machine manufacturing.

Experimental patch antenna radiation patterns were measured and verified against simulated results. However, the amplifier could not be biased inside the anechoic chamber. Two test patch antennas were milled without an amplifier for use inside the anechoic chamber, see Figure 5.1.

![Figure 5.1. Patch antenna used for antenna testing.](image)
The patch antennas were tuned to operate at 915MHz by scoring the top of the patch antenna, see Figure 5.1. The tuned length is derived from equivalent electrical lengths for the tuned and initial patches

\[ L_{tuned} = \frac{f}{f_{desired}} L \]  

(5.1)

where \( f \) and \( L \) are the original operating frequency and length, respectively, and \( f_{desired} \) is the desired operating frequency. The original and tuned return loss for patch A are plotted in Figure 5.2. Tuned return loss for patch B is given in Figure 5.3.

![Patch A Return Loss](image)

Figure 5.2. Patch A return loss.
5.1 Test Preparation

To compare HFSS far-field radiation patterns, the transmit and receive antennas were positioned in the far zone of each antenna; [1]

$$R_{ff} = \frac{2D^2}{\lambda}$$

(5.2)

where $D$ is the maximum antenna dimension and $\lambda$ is the operating wavelength. The far field distance at 915MHz for each antenna is listed in Table 5.1.
Table 5.1. Antenna far field distances.

<table>
<thead>
<tr>
<th></th>
<th>D (in)</th>
<th>Rff (in)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rotating Dipole</td>
<td>6.45</td>
<td>6.45</td>
</tr>
<tr>
<td>Stationary Dipole</td>
<td>6.45</td>
<td>6.45</td>
</tr>
<tr>
<td>Quagi</td>
<td>3.23</td>
<td>1.62</td>
</tr>
<tr>
<td>90° Corner Reflector</td>
<td>12.91</td>
<td>25.82</td>
</tr>
<tr>
<td>60° Corner Reflector</td>
<td>16.78</td>
<td>43.62</td>
</tr>
<tr>
<td>30° Corner Reflector</td>
<td>30.98</td>
<td>148.66</td>
</tr>
<tr>
<td>Patch antenna</td>
<td>5</td>
<td>3.87</td>
</tr>
</tbody>
</table>

The two patch antennas were mounted in the anechoic chamber and $|S_{21}|$ was measured. The absolute gain of each antenna is calculated through the Friis equation [1]

$$P_r(dBm) = G_r(dB) + G_t(dB) + P_t(dBm) + 20\log \left( \frac{4\pi R(m)}{\lambda(m)} \right)$$  \hspace{1cm} (5.3)

where $P_r(dBm)$ is the receive power, $G_r(dB)$ is the receive antenna gain, $G_t(dB)$ is the transmit antenna gain, $P_t(dBm)$ is the transmit power. The last term is the spreading loss where $R(m)$ is the separation distance in meters and $\lambda(m)$ is the wavelength in meters.

Rearranging the equation, we obtain

$$G_r(dB) + G_t(dB) = P_r(dBm) - P_t(dBm) + 20\log \left( \frac{4\pi R(m)}{\lambda(m)} \right)$$  \hspace{1cm} (5.4)

Since transmit and receive antenna theoretical gains are identical the gain of each antenna is

$$G_r(dB) = G_t(dB) = \frac{1}{2} \left[ L_1(dB) + L_2(dB) + |S_{21}(dB)| + 20\log \left( \frac{4\pi R(m)}{\lambda(m)} \right) \right]$$  \hspace{1cm} (5.5)

where $P_r(dBm) - P_t(dBm)$ is replaced with $L_1(dB) + L_2(dB) + |S_{21}(dB)|$ where $L_1(dB)$ and $L_2(dB)$ are the loss terms for cables connected to ports 1 and 2, respectively. The
absolute gain of each patch antenna is approximately 2.07dB. The HFSS model predicted a 4.51dB gain, a 2.44dB difference. The error is due in part to non-identical antennas. Because the antennas required tuning, they have different return loss values at the operating frequency.

5.2 Rotating Dipole

The rotating dipole was mounted to the positioner via a wooden base used for the rotating and stationary dipoles and Quagi, see Figure 5.4. The antenna was secured using multiple cable ties and tested to ensure position stability. The rotating dipole exhibits a -10.6dB return loss at 915MHz, see Figure 5.5.

Figure 5.4. Rotating dipole positioner mount.
The rotating dipole antenna provides a 1.96dB maximum gain (expected = 2.15dB). The measured H-plane radiation pattern varies by a maximum of 5dB, but exhibits lobes when positioned away from the transmit antenna (90° to -90° including 180°), see Figure 5.6. This is most likely caused by positioner head attenuation since it is between the transmit and receive antennas for $\varphi = -100^\circ$ to $100^\circ$ including $180^\circ$. The measured E-plane shows a null at approximately $90^\circ$. Two nulls are expected at $\pm90^\circ$. The absent null at -90° is most likely caused by reflections from the positioner mount and Lazy-Susans on the antenna base since they are between the transmit and receive antenna between $0^\circ$ and $180^\circ$ including $-90^\circ$. 

Figure 5.5. Rotating dipole return loss.

The rotating dipole return loss is shown in the graph. The data points are plotted against frequency in MHz, with the Y-axis showing the return loss in dB. The graph shows a minimum return loss of approximately -10dB at around 900 MHz, with a peak return loss of about -2dB at 940 MHz.
Figure 5.6. Rotating dipole measured H-plane radiation pattern.
5.3 Stationary Dipole

The stationary dipole uses the rotating dipole mounting fixture, see Figure 5.8. Cable ties were used to secure the dipole. The positioner was rotated 90° about the x axis to verify stationary dipole mounting, see Figure 5.8 for coordinate system definition.
The stationary dipole exhibits -12.8dB return loss at 915MHz, and operates over a 44MHz wider frequency range than the rotating dipole. A 2.2dB improved return loss occurs at 915MHz.

**Stationary Dipole Return Loss**

![Stationary Dipole Return Loss Diagram](image)

**Figure 5.8. Stationary dipole positioner mount.**

**Figure 5.9. Stationary dipole return loss.**
The stationary dipole provides a -0.56dB maximum gain at $\varphi = 0^\circ$ in the H-plane, a -2.71dB error from the expected gain. This could be caused by excessive solder where the split balun meets the copper elements. The solder extends into the split balun which may shift the location of the short circuit. The stationary dipole exhibits side lobes similar to the rotating dipole when the positioner is between the transmit and receive antennas. The side lobes are unexpected since dipole antennas should radiate equally in all H-plane directions.

The stationary dipole exhibits both nulls in the E-plane since Lazy Susans do not exist in the stationary dipole base, see Figure 5.11. The nulls are located at approximately $\varphi = 90^\circ$ and $\varphi = -95^\circ$. The expected locations are $\varphi = \pm 90^\circ$, a $5^\circ$ error for the -95° null. This may be caused by orientation errors due to imperfect soldering methods, see Figure 5.8. The HPBW in the E-plane is approximately 96°; 21° wider than for the rotating dipole. This difference may be caused by an impedance mismatch from the solder inside the split balun.
Figure 5.10. Stationary dipole measured H-plane radiation pattern.
Figure 5.11. Stationary dipole measured E-plane radiation pattern.
5.4 Quagi

The Quagi was also secured using cable ties, see Figure 5.12, and exhibits -10.06dB return loss at 915MHz, see Figure 5.13.
The Quagi exhibits a main lobe with a 1.28dB maximum gain at $\varphi = 0^\circ$ (H-plane), see Figure 5.14. The second largest lobe is $180^\circ$ opposite the main, yielding a 10.93dB front to back ratio. The H-plane HPBW is 56°.

The Quagi has a 49° HPBW in the E-plane, see Figure 5.15. The pattern contains three distinct lobes, the main lobe at $\varphi = 0^\circ$, and two smaller lobes at $\varphi = \pm 145^\circ$. The pattern exhibits two nulls at $\pm 90^\circ$ which is expected.

![Quagi Antenna H-Plane](image)

Figure 5.14. Quagi measured H-plane radiation pattern.
Figure 5.15. Quagi measured E-plane radiation pattern.

5.5 Corner Reflector

Due to the weight of the corner reflector, a stronger base was required, see Figure 5.16. The base was assembled using walkway foam and wooden planks to position the corner reflector at the transmit antenna height. The walkway foam was stacked to avoid damage to the fiber optic control line. The rotating base was locked in position using nonconductive duct tape. Because of the large corner reflector size, only H-plane patterns were measured.
The corner reflector exhibits the best matching of all dipoles, a -21.16dB return loss at 915MHz, see Figure 5.17.

Figure 5.16. Corner reflector positioner base.

Figure 5.17. Corner reflector return loss.
The 90°, 60°, and 30° co-pol H-plane patterns are shown in Figure 5.18. The 90° corner reflector exhibits the largest maximum gain, 6.70dB at \( \varphi = 0^\circ \). The 90° corner reflector is expected to provide the largest gain since the 90° corner has the widest effective aperture: area which captures energy from an incident electromagnetic wave. The 90° corner reflector H-plane HPBW is approximately 54°.

![Corner Reflector H-Plane](image)

**Figure 5.18.** Corner reflector measured H-plane co-pol radiation patterns for all corner angles.

The 60° corner has the second largest maximum gain, 5.97dB at \( \varphi = 0^\circ \) and approximately 55° HPBW. The 30° corner exhibits the smallest maximum gain, 3.96dB.
at $\varphi = 0^\circ$. Due to a design flaw, the 30° corner dipole extends past the corner edges, see Figure 5.19. This results in reduced gain and a larger HPBW (96°) caused by dipole radiation that is not directed by the reflector planes. To solve this problem, reflector planes larger than 2’ x 2’ are required; however, this is not possible since portability is a project goal.

![Diagram](image)

**Figure 5.19. Radiation not directed from 30° corner reflector.**

None of the antennas exhibit nulls in the reflector plane directions. Nulls are expected for infinite reflector planes since an opposite polarity dipole image is located the same distance away from the corner. The nulls are non-existent due to the finite corner size since a dipole image is not present for all reflection angles and diffraction occurs at the reflector plane edge.
5.6 Embedded Patch

The patch antenna positioner fixture is shown in Figure 5.20. The patch was mounted using a wooden backboard and secured using nylon screws.

![Patch antenna A positioner mount.](Image)

Figure 5.20. Patch antenna A positioner mount.

Patch antenna A return loss is shown in Figure 5.2 and has a -12.82dB return loss at 915MHz, 13.18dB greater than the HFSS prediction. The H-plane radiation pattern is shown in Figure 5.21. The patch antenna has an 82° H-plane HPBW, matching the HFSS prediction. However, the maximum gain is 2.44dB less than the HFSS prediction. This is could be caused by the diffraction at the patch edge. The measured and simulated co-pol H-plane radiation patterns are shown in Figure 5.22.
Figure 5.21. Patch antenna measured H-plane radiation pattern.

Figure 5.22. Measured and predicted patch H-plane co-pol comparison.
The measured E-plane radiation patterns are shown in Figure 5.23. The measured and simulated co-pol E-plane patterns are shown in Figure 5.24. The E-plane HPBW is 80°, 3° wider than the HFSS prediction. Although the co-pol gain differs by less than 4dB from the predicted co-pol gain, the cross-pol gain is 34dB greater than HFSS results. This is caused by reflections from the positioner.

![Patch Antenna E-Plane](image)

**Figure 5.23.** Patch antenna measured E-plane radiation pattern.
Figure 5.24. Measured and predicted patch E-plane co-pol comparison.
Chapter 6 System Analysis

6.1 Cascaded Components

The wireless system is designed to maximize the received signal amplitude and the signal-to-noise ratio at the signal strength indicator. The components were designed for minimum insertion loss and maximum gain.

Another measure taken to increase the signal-to-noise ratio is the component order. The noise figure of two cascaded components is

\[ F = F_1 + \frac{F_2 - 1}{G_1} \]  

(6.1)

where \( F \) is the total noise figure, \( F_1 \) and \( F_2 \) are the noise figures of the first and second components, and \( G_1 \) is the gain of the first component.

An amplifier should be positioned before a bandpass filter, since the bandpass filter noise figure is divided by the amplifier gain. This is the order used in the actual system.

The total noise figure for both configurations was calculated using the noise figure and gain for an AH-1 amplifier and the constructed bandpass filter. The AH-1 provides 13.5dB gain and has a 2.7dB noise figure. The bandpass filter has -2.7dB gain and a 2.7dB noise figure at 915MHz. Placing the bandpass filter before the amplifier results in a 5.4dB overall noise figure, however placing the amplifier first reduces the overall noise figure to 2.79dB.
6.2 SNR and Power Levels

Maintaining a large signal to noise ratio is important in all systems to maximize received signal reception. This section analyzes the signal to noise ratio and power levels from the oscillator to the signal strength indicator input.

The available noise power is [13]

\[ P_N = kTB \]  

(6.2)

where \( k \) is Boltzmann’s constant, \( T \) is the temperature in K, and \( B \) is the bandwidth in Hz. The available noise power for a 1Hz bandwidth is as -174dBm at room temperature (290K) [13]. The worst case system bandwidth is 53MHz, the operational bandwidth of the Quagi antenna. Converting 53MHz to dBHz using \( 10 \cdot \log(BW) \) and adding to -174dBm yields the available noise power for a 53MHz bandwidth signal, -96.76dBm.

The noise and signal power levels at each stage for the worst case system configuration are shown in Figure 6.1. The worst case configuration is the transmit dipole, -0.56dB gain at 0° in the H-plane.
The initial SNR is 100.0dB. The spreading loss attenuates the signal 17.6dB more than the noise power because the noise power is reduced to the noise floor. This results in an 82.4dB received SNR. The input power at the signal strength indicator is -3.93dBm which is sufficient for operating all 30 LEDs. With the 90° corner reflector directed at the TX antenna, the signal strength indicator receives 6.24dBm, 6.14dB greater than the stationary dipole.

Figure 6.1. Worst case transmit power level diagram.
Chapter 7 System Demonstration

7.1 System Setup

An open workspace that allows for antenna separation (3ft) and DC power supplies is used for the system demonstration. Conducting objects are removed to minimize reflections. The oscillator, amplifier, and stationary dipole are interconnected using SMA coax cables. 9V DC and 6V DC is applied to the oscillator and signal strength indicator, respectively. The amplifier potentiometer is set to 0Ω and 5V is applied across the collector-emitter bias port. The potentiometer resistance is increased until the base-emitter voltage reaches 0.7V.

The corner reflector without reflector planes is located 3ft from the rotating dipole which is oriented as shown in Figure 4.4. The stationary dipole antenna is connected to the bandpass filter and signal strength indicator. Gain is increased by adjusting the potentiometer resistance until all LEDs are illuminated. If maximum gain is applied and some LEDs are not responding, reduce the gain to 0 (decreasing R5 to 0Ω) and insert the second RF amplifier between the receive antenna and bandpass filter. Increase signal strength indicator gain until all 30 LEDs are illuminated. The system is now ready for demonstration.

NOTE: Occasionally, the negative resistance oscillator operation is intermittent. If no LEDs are illuminated after inserting the RF amplifier, replace the oscillator with the Mini-Circuits ZX95-928BCA oscillator. Apply 5.1V to $V_{\text{tune}}$ and 8V to $V_{\text{cc}}$, and connect
the output to the transmit amplifier input. Repeat the setup procedure to initialize the system. The oscillator provides 3.3dBm at 915MHz.

7.2 Corner Reflector

The corner reflector demonstrates antenna arrays and radiation pattern effects. Position the corner reflector as depicted in Figure 7.1 and remove reflector panels. Gradually rotate it about the z axis and observe signal strength. The signal strength should remain constant.

Introduce the 90° corner and slowly rotate it about the z axis to observe how the signal attenuates when the main beam is directed away from the transmit antenna. The received signal strength should decrease as the corner is rotated. Adjust the dipole distance from the vertex to demonstrate maximums and nulls (adjust along x-axis). Repeat the procedure with a 60° and 30° corner angle. The 90°, 60°, and 30° corners have 54°, 55°, and 96° H-plane half power beamwidths (HPBW), respectively.
7.3 Embedded Patch

![Embedded Patch Diagram]

Figure 7.2. Embedded patch demonstration setup.

The embedded PCB patch antenna demonstrates modern wireless communications technology. The antenna is milled adjacent to RF circuitry on the same PCB. If flexible laminate is used, the antenna can conform to nonplanar surfaces.

The amplifier is biased as described in section 7.1. The patch antenna is located 3ft from the transmit antenna. Observe the H-plane radiation pattern by rotating the patch antenna in the azimuthal direction (adjust $\phi$). The H-plane HPBW is 82°. Observe the E-plane radiation pattern by rotating the patch antenna about the y-axis. The E-plane HPBW is 80°
7.4 Quagi

![Figure 7.3. Quagi demonstration setup.](image)

The Quagi antenna demonstrates a directional antenna and associated radiation patterns. The Quagi’s main beam maximum is directed toward the transmit antenna. Record how many LEDs are illuminated for comparison to other antennas. Slowly adjust the Quagi azimuthal angle (rotate in $\phi$ direction) and observe the received signal strength angle dependence on main beam. The H-plane HPBW is 56°, indicated by 9 illuminated LEDs.
7.5 Dipole Antenna

The rotating and stationary dipoles demonstrate signal reception, polarization, radiation patterns, and antenna pattern nulls. With all 30 LEDs illuminated, adjust rotating antenna roll angle to demonstrate polarization effects on received signal strength. Gradually adjust the roll angle 90° about the x-axis, see Figure 7.4. Signal strength will gradually decrease until no LEDs are illuminated at 90° roll. Polarization is illustrated for all other demonstrations by adjusting the transmit dipole roll angle.

To demonstrate dipole antenna radiation pattern nulls, position the rotating dipole as depicted in Figure 7.1. Rotate the transmit antenna 90° in both θ and φ directions. The tip of the dipole should be directed at the transmit antenna. In this orientation, zero received power should be indicated. The H-plane radiation pattern is demonstrated by adjusting the rotating dipole about z axis.

Spreading loss is demonstrated by increasing the distance between the TX and RX antennas to 5ft, which reduces the indicator from 30 to 18 illuminated LEDs. The spreading loss (last term in eq. 5.3) increases by 4.46dB, which reduces the peak detector output voltage to 2.4V (Figure 3.9) and illuminated LEDs to 18 (Figure 3.12).
7.6 Environmental Barriers

The environmental barriers can be used in conjunction with any of the first four demonstrations. Placing the barriers midway between transmit and receive antennas (far-field) demonstrates attenuation effects on received signal strength.

Because ESD bags, copper mesh, and poultry net are composed of conducting components, relatively large signal attenuation occurs. For the stationary dipole configuration, waveguide analysis (Section 4.7.1) indicates 21 or 0 illuminated LEDs are expected with the poultry net or copper mesh barrier in place, respectively.

For the stationary dipole configuration, Figure 6.1 indicates -1dBm is applied to the peak detector which provides 0.68V DC to the bar graph driver, see Figure 3.9. To illuminate all LEDs, a 5.51 magnitude gain is required from the DC amplifier. With the poultry net barrier in place, -2.67dBm is applied to the peak detector which provides 0.49V to the DC amplifier. The DC amplifier provides 2.7V to the bar graph driver which illuminates 21 LEDs, see Figure 3.1 Using similar analysis for the copper mesh, -23.06dBm is applied to the peak detector which results in 0 illuminated LEDs.
Chapter 8 Conclusion

The wireless system successfully demonstrates many antenna characteristics with minimal support instrumentation (DC power supply). The system can be constructed with basic woodworking and hand tools and can be assembled in as little as four weeks with proper equipment. The largest component measures 2’ x 2’ which enables system portability for demonstration at multiple schools.

Antenna parameters including polarization, radiation pattern, gain, and barrier-induced signal attenuation effects are illustrated using four antenna types; dipole, corner reflector, Quagi, and embedded patch, which represent dominant antenna types used in current applications. Visual received signal strength indication is provided by an LED array driven by an RF amplifier, peak detector, and bar graph driver. All receive antennas are mounted to rotational platforms to allow investigation of antenna performance. The transmitted 915MHz signal is produced by student-designed components: an oscillator, amplifiers, and a filter.

The stationary dipole differed in performance from the rotational version with regard to gain and return loss. These differences were due to inconsistent solder joints. Excessive solder was applied to the stationary dipole antenna which may have created an impedance mismatch.

The Quagi antenna exhibited lower than expected gain. This is due to element size imperfections. The Quagi feed gap is specified by 1/32” determined experimentally.

The corner reflector radiation pattern did not include nulls at the plane angles. This was caused by finite reflector plane dimensions. 2’ x 2’ dimensions are
recommended to avoid source placement outside the reflector planes, i.e.: 30 deg corner angle case.

Spreading loss was demonstrated by increasing the separation distance between the antennas for each configuration. For the stationary dipole demonstration, 18 illuminated LEDs corresponds to a TX/RX antenna separation distance of 5’.

Dipole construction difficulties included tuning the dipoles to 915MHz, applying consistent solder, and replicating split balun dimensions. These issues caused performance to degrade with regard to gain and return loss. These problems can be mitigated by carefully installing the dipoles in the bases to avoid breaking solder joints, and by using a consistent method for drilling and cutting the tubing.

Actual barrier implementation compared well to expected performance. Only one LED remained illuminated with the ESD or copper mesh barriers, 29 LEDs remained illuminated with the drywall barrier, and 8 LEDs remained illuminated with the poultry net barrier. Conductivity differences for the four barrier materials correlate to signal attenuation. These results can be used to estimate system performance in residential or manufacturing environments.

The bandpass filter design required calculating the spacing, length, and widths of coupled sections and optimizing them using ADS. Several filter topologies were considered including \(\lambda/4\) coupled resonator, cascaded lowpass and highpass, and capacitively coupled \(\lambda/4\) resonator filters. These types were evaluated using design methods in [1]. The other considered designs were not well documented or required unavailable components.
The oscillator experienced intermittent performance potentially caused by temperature variation and limited coupling. It is recommended that this circuit be reconstructed with attention to the coupling wire to enhance performance reliability.

The antenna demonstration system can be used to familiarize high school and undergraduate college students with wireless system operation and radio communications. System durability and open access in a laboratory setting enables student investigation of wireless system reception through physical rotation, orientation, and positioning of antenna components.

Through wireless system experimentation and performance observations, students gain insight on the causes of signal reception variation in cell phones, wireless internet, television, radio, and other common wireless devices. Supplemental explanations of antenna operating principles to describe wireless system performance characteristics will enable an improved understanding of “The Magic of Antennas.”
Future Project Recommendations

1. Model and optimize a 915MHz Quagi antenna
2. Model and optimize 915MHz dipoles that incorporate split baluns
3. Model and optimize a corner reflector with finite reflector planes
4. Develop a signal strength indicator that illuminates LEDs logarithmically proportional to received signal strength
5. Develop a reliable oscillator that incorporates a phase locked loop for stability
References


Appendix A: Bandpass Filter Design Equations

To obtain coupled section even and odd mode impedances for a bandpass filter, the width and spacing of the coupled sections are adjusted. The even and odd mode impedances are calculated based on width, spacing, and laminate properties in the following equations. The effective even-mode dielectric constant is [9]:

\[ \varepsilon_{re}^e = \varepsilon_r + \frac{1}{2} \left( 1 + \frac{10}{v} \right)^{-a_{re}} \]  \hspace{1cm} (A.1)

where

\[ v = \frac{u(20 + g^2)}{10 + g^2} + g \exp(-g) \]  \hspace{1cm} (A.2)

\[ a_{e} = 1 + \frac{1}{49} \ln \left[ \frac{\frac{v^4}{v^4 + 0.432} + 1}{18.7} \ln \left( \frac{v}{18.1} \right)^3 \right] \]  \hspace{1cm} (A.3)

\[ b_{e} = 0.564 \left( \frac{\varepsilon_r - 0.9}{\varepsilon_r + 3} \right)^{0.053} \]  \hspace{1cm} (A.4)

where \( u \) is the coupled line width divided by the substrate height and \( g \) is the coupled line spacing to substrate ratio. The effective odd mode dielectric constant is calculated using [9]:

\[ \varepsilon_{ro}^o = \varepsilon_{re} + [0.5(\varepsilon_r + 1) - \varepsilon_{re} + a_o] \exp(-c_0 g^{d_e}) \]  \hspace{1cm} (A.5)

where

\[ a_o = 0.7287[\varepsilon_{re} - 0.5(\varepsilon_r + 1)][1 - \exp(-0.719u)] \]  \hspace{1cm} (A.6)

\[ b_o = \frac{0.747 \varepsilon_r}{0.15 + \varepsilon_r} \]  \hspace{1cm} (A.7)
\begin{align*}
c_o &= b_o - (b_o - 0.207) \exp(-0.414u) \quad (A.8) \\
d_o &= 0.593 + 0.694 \exp(-0.526u) \quad (A.9)
\end{align*}

and

\begin{align*}
e_{re} &= \frac{\varepsilon_e + 1}{2} \left(1 + \frac{10}{u}\right)^{-ab} \\
a &= 1 + \frac{1}{49} \ln \left(\frac{u^4 + \left(\frac{u}{52}\right)^2}{u^4 + 0.432}\right) + \frac{1}{18.7} \ln \left[1 + \left(\frac{u}{18.1}\right)^3\right] \quad (A.11)
\end{align*}

\begin{align*}
b &= 0.564 \left(\frac{\varepsilon_e - 0.9}{\varepsilon_e + 3}\right)^{0.053} \\
\text{The even-mode microstrip line characteristic impedance is given by} \ [9] \\
\begin{align*}
Z_{ce} &= Z_c \sqrt{\frac{\varepsilon_{re}}{\varepsilon_{re}}} \\
&= \frac{Z_c \sqrt{\varepsilon_{re} / \varepsilon_{re}}}{1 - Q\sqrt{\varepsilon_{re} \cdot Z_c / 377}} \quad (A.13)
\end{align*}
\end{align*}

where

\begin{align*}
Z_c &= \frac{377}{2\pi \sqrt{\varepsilon_{re}}} \ln \left[\frac{F}{u} + \sqrt{1 + \left(\frac{2}{u}\right)^2}\right] \quad (A.14) \\
F &= 6 + (2\pi - 6) \exp \left[-\left(\frac{30.666}{u}\right)^{0.7528}\right] \quad (A.15) \\
Q_1 &= 0.8685u^{0.194} \quad (A.16) \\
Q_2 &= 1 + 0.7519g + 0.189g^{2.31} \quad (A.17)
\end{align*}
\[ Q_3 = 0.1975 + \left[ 16.6 + \left( \frac{8.4}{g} \right)^6 \right]^{-0.387} + \frac{1}{241} \ln \left[ \frac{g^{10}}{1+(g/3.4)^{10}} \right] \] (A.18)

\[ Q_4 = \frac{2Q_1}{Q_2} \cdot \frac{1}{u^{Q_3}} \exp(-g) + [2 - \exp(-g)]u^{-Q_3} \] (A.19)

The odd-mode microstrip line characteristic impedance is given by [9]

\[ Z_{eo} = \frac{Z_c \sqrt{\varepsilon_{re}/\varepsilon_{co}}}{1-Q_0 \sqrt{\varepsilon_{re} \cdot Z_c/377}} \] (A.20)

where

\[ Q_5 = 1.794 + 1.14 \ln \left[ 1 + \frac{0.638}{g + 0.517g^{2.43}} \right] \] (A.21)

\[ Q_6 = 0.2305 + \frac{1}{281.3} \ln \left[ \frac{g^{10}}{1+(g/5.8)^{10}} \right] + \frac{1}{5.1} \ln(1+0.598g^{1.154}) \] (A.22)

\[ Q_7 = \frac{10+190g^2}{1+82.3g^3} \] (A.23)

\[ Q_8 = \exp[-6.5-0.95\ln(g)-(g/0.15)^5] \] (A.24)

\[ Q_9 = \ln(Q_7) \cdot (Q_8 + 1/16.5) \] (A.25)

\[ Q_{10} = Q_7 - \frac{Q_5}{Q_2} \exp \left( \frac{Q_6 \ln(u)}{u^{Q_3}} \right) \] (A.26)
Appendix B: Contest Proposal

IEEE Antenna Design Challenge
By Alex Hempy and Michael Civerolo

Objective

We plan to build a homemade wireless system that is compatible with dipole, loop Yagi-Uda, and PCB embedded microstrip antennas. We will use the system to demonstrate polarization, radiation patterns, and environmental effects on wireless transmission and reception. The receiver includes an LED indicator to display signal strength and identify optimum antenna configurations: see Fig. 1.

Both the TX and RX dipole antennas will have an SMA center feed and be constructed with 10 AWG copper wire. The TX dipole antenna is used in each demonstration and is fixed vertically.

The first demonstration involves transmit and receive dipole antennas. The receive antenna’s mounting fixture allows rotation along two orthogonal directions to demonstrate polarization and pattern effects (Fig. 2a). A corner reflector composed of two sheet metal planes connected with hinges is placed a quarterwave from the vertical receive dipole antenna. The corner reflector is set in a slotted wooden base (Fig. 2b) to allow discrete corner angles simulating various numbers of array elements. This will demonstrate how gain and patterns can be modified by effectively changing the number of antennas.

The second demonstration involves a loop Yagi-Uda receive antenna mounted on a rotating fixture (Fig. 2c). This allows the antenna and its main beam to be steered towards or away from the transmitting dipole. The antenna loops are formed with 10 AWG copper wire and attached to a wooden boom. This demonstrates polarization, radiation patterns, and directionality effects.

The third and final demonstration utilizes an embedded microstrip dipole receive antenna, which is milled onto the RF amplifier printed circuit board (Fig. 2d). This antenna emphasizes advantages of microstrip antennas: low profile, ease of fabrication, low cost, and implementable on printed circuit boards with other circuitry.

Sections of drywall, chicken wire, copper mesh, and ESD material (2ft x 2ft) will be placed in the transmitting antenna far-zone and will illustrate the effects of environmental barriers on propagating signals. The entire system (Fig. 1), including all antennas, will be student-built and will demonstrate signal interference from environmental barriers, polarization effects, radiation patterns, a simulated array (corner reflector), gain, and directivity.
Preliminary Design

![Block diagram of the wireless system](image)

**Figure 1:** Block diagram of the wireless system

The signal will be generated by a 915MHz negative resistance oscillator constructed from copper wire, a high frequency BJT, RF choke, resistors, and capacitors. For temperature stability, the RF source will incorporate a phase locked loop referenced to a 10MHz Colpitts crystal oscillator.

Three amplifiers will be built to amplify signals from the transmitter, the receive embedded dipole antenna, and the dipole and Yagi-Uda receive antennas.

Two 915MHz microstrip bandpass filters will be designed and constructed to suppress out-of-band transmissions by at least 10dB. Both bandpass filters will consist of cascaded Chebyshev highpass and lowpass filters with nominal cutoff frequencies of 900MHz and 930MHz, respectively. The RX bandpass filter and LED signal strength indicator will be on the same PCB.

The LED signal strength indicator includes a peak detector connected to a high frequency Schottky diode and RC lowpass filter to capture the peak voltage. An LED bar graph converts this peak voltage to a visual display of the received signal strength. The number of illuminated LEDs is proportional to the signal strength magnitude.

![Antennas](image)

**Figure 2:** Rotating dipole antenna (a), corner reflector (b), Yagi-Uda antenna, design from ARRL Antenna Handbook (c), and Embedded dipole antenna, concept from SPEAG (d)
## Bill of Materials

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<td>6.99</td>
<td>2</td>
<td>13.98</td>
<td>Local hobby store</td>
<td></td>
</tr>
<tr>
<td>3' length 5/32&quot; diameter brass tubing</td>
<td>3.19</td>
<td>2</td>
<td>6.38</td>
<td>Local hobby store</td>
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</tr>
<tr>
<td>4' RG174 SMA male male coax</td>
<td>7.58</td>
<td>2</td>
<td>15.16</td>
<td>CO-174SMAX200</td>
<td><a href="http://www.cablesondemand.com">www.cablesondemand.com</a></td>
</tr>
<tr>
<td>2' RG174 SMA male male coax</td>
<td>7.10</td>
<td>2</td>
<td>14.20</td>
<td>CO-174SMAX200</td>
<td><a href="http://www.cablesondemand.com">www.cablesondemand.com</a></td>
</tr>
<tr>
<td>Misc. screws, nails, wood glue</td>
<td>10.00</td>
<td>1</td>
<td>10.00</td>
<td></td>
<td>Local hardware store</td>
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<tr>
<td>0.0865&quot; semi-rigid coax, 1ft</td>
<td>24.62</td>
<td>1</td>
<td>24.62</td>
<td>UT-85-TP-M17</td>
<td><a href="http://www.microstock-inc.com">www.microstock-inc.com</a></td>
</tr>
<tr>
<td>SMA straight plug 0.085&quot; connector</td>
<td>2.32</td>
<td>2</td>
<td>4.64</td>
<td>132101</td>
<td><a href="http://www.mouser.com">www.mouser.com</a></td>
</tr>
<tr>
<td><strong>Rotating Dipole Antenna</strong></td>
<td></td>
<td></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>3&quot; x 3&quot; Lazy Susan bearing</td>
<td>$1.50</td>
<td>2</td>
<td>$3.00</td>
<td></td>
<td>Local hardware store</td>
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<tr>
<td><strong>Corner Reflector Antenna</strong></td>
<td></td>
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<td></td>
</tr>
<tr>
<td>6&quot; x 6&quot; Lazy Susan bearing</td>
<td>$3.50</td>
<td>2</td>
<td>$7.00</td>
<td></td>
<td>Local hardware store</td>
</tr>
<tr>
<td>16 mil thick 12&quot; square copper sheets (2)</td>
<td>39.99</td>
<td>1</td>
<td>39.99</td>
<td><a href="http://www.basiccopper.com">www.basiccopper.com</a></td>
<td></td>
</tr>
<tr>
<td><strong>Quagi Antenna</strong></td>
<td></td>
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</tr>
<tr>
<td>5/16&quot; diameter 2&quot; long wood dowel</td>
<td>$0.20</td>
<td>1</td>
<td>$0.20</td>
<td>Local hardware store</td>
<td></td>
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<tr>
<td><strong>Oscillator</strong></td>
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<tr>
<td>Avago AT-41486 transistor</td>
<td>$2.59</td>
<td>1</td>
<td>$2.59</td>
<td>AT-41486-BLKG</td>
<td><a href="http://www.digikey.com">www.digikey.com</a></td>
</tr>
<tr>
<td>6800pF capacitor</td>
<td>0.22</td>
<td>5</td>
<td>1.10</td>
<td>C410C682K1R5TA7200</td>
<td><a href="http://www.digikey.com">www.digikey.com</a></td>
</tr>
<tr>
<td>RF choke</td>
<td>0.12</td>
<td>1</td>
<td>0.12</td>
<td>BL01RN1A1D2B</td>
<td><a href="http://www.mouser.com">www.mouser.com</a></td>
</tr>
<tr>
<td><strong>Amplifiers (x2)</strong></td>
<td></td>
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<td></td>
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<tr>
<td>Avago AT-41435G transistor</td>
<td>$5.00</td>
<td>3</td>
<td>$15.00</td>
<td>AT-41435G</td>
<td><a href="http://www.avnet.com">www.avnet.com</a></td>
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<tr>
<td>220pF SMT capacitor</td>
<td>0.05</td>
<td>12</td>
<td>0.60</td>
<td>06035A221J4T2A</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
</tr>
<tr>
<td>25 turn 0.5W 10kΩ potentiometer</td>
<td>1.95</td>
<td>3</td>
<td>5.85</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td><strong>Signal Strength Indicator</strong></td>
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<tr>
<td>HSMS-2822 Schottky diode</td>
<td>$0.67</td>
<td>1</td>
<td>$0.67</td>
<td>HSMS-2822-BLKG</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
</tr>
<tr>
<td>LM3914N-1 linear bar graph driver</td>
<td>1.95</td>
<td>3</td>
<td>5.85</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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</tr>
<tr>
<td>Protoboard</td>
<td>13.95</td>
<td>1</td>
<td>13.95</td>
<td>2852PCB-R</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
</tr>
<tr>
<td>Mounting hardware kit</td>
<td>2.95</td>
<td>1</td>
<td>2.95</td>
<td>106551</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td>Green LEDs</td>
<td>0.10</td>
<td>10</td>
<td>1.00</td>
<td>LG13740</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td>Yellow LEDs</td>
<td>0.15</td>
<td>10</td>
<td>1.50</td>
<td>LY3330</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td>Red LEDs</td>
<td>0.11</td>
<td>10</td>
<td>1.10</td>
<td>LTL-307E</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
</tr>
<tr>
<td>LM324N operational amplifier</td>
<td>0.29</td>
<td>1</td>
<td>0.29</td>
<td>LM324N</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td>18 pin IC socket</td>
<td>0.14</td>
<td>4</td>
<td>0.56</td>
<td>CA-18SDL-1T</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
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<tr>
<td>1000pF capacitor</td>
<td>0.12</td>
<td>1</td>
<td>0.12</td>
<td>SA105A102JAA</td>
<td><a href="http://www.jameco.com">www.jameco.com</a></td>
</tr>
<tr>
<td><strong>Environmental Barrier</strong></td>
<td></td>
<td></td>
<td></td>
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</tr>
<tr>
<td>1/2&quot; x 3/4&quot; pine, 1' length</td>
<td>$0.74</td>
<td>24</td>
<td>17.76</td>
<td></td>
<td>Local hardware store</td>
</tr>
<tr>
<td>2&quot; x 2', 0.5&quot; thick drywall</td>
<td>3.98</td>
<td>1</td>
<td>3.98</td>
<td></td>
<td>Local hardware store</td>
</tr>
<tr>
<td>6&quot; x 10&quot; ESD bags (100ct)</td>
<td>13.40</td>
<td>1</td>
<td>13.40</td>
<td>48602</td>
<td><a href="http://www.mouser.com">www.mouser.com</a></td>
</tr>
<tr>
<td>2&quot; wide poultry net</td>
<td>5.88</td>
<td>1</td>
<td>5.88</td>
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<td>Local hardware store</td>
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<tr>
<td>Copper Mesh #60 X 60 .0075&quot; 24&quot; x 24&quot;</td>
<td>22.90</td>
<td>1</td>
<td>22.90</td>
<td><a href="http://www.amazon.com">www.amazon.com</a></td>
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