A LOW-COST OMNIDIRECTIONAL ANTENNA FOR USE IN OUTDOOR WI-FI ACCESS POINTS

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ABSTRACT

This thesis details the design of a dual-broadband, omnidirectional antenna intended for use in outdoor Wi-Fi access points. The design utilizes a corporate-feed topology to ensure an in-phase feeding of two stacked dual-band elements over a wide band. The lengths of the traces that feed each element are minimized so that the design may be based around a higher loss-tangent (lower cost) substrate. The design of the dual-band element first accounts for the loading effect of the substrate using Jaisson’s approximation. This allows us to approximate the effective lengths of two serially-fed dipoles that constitute each dual-band element. We then calculate the far-field patterns and space the elements to achieve the maximum range under the constraint that the entire structure be at most 101.6 [mm] (4”) tall. Then, incorporating the balun proposed in [1], the structure is optimized in CST Microwave Studio to achieve the minimum azimuth plane ripple for its configuration. Next, we introduce the feed network and impedance match the antenna at 2.4 [GHz] using a shorted-shunt stub. The resulting antenna achieves peak gains of 3 dBi and 6 dBi in the 2.4 and 5 [GHz] Wi-Fi frequency bands and impedance bandwidths in excess of 25% over both bands. We conclude with a cost analysis of the antenna, and based on performance, size, and cost, we advocate its use in present and future Wi-Fi access points.
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CHAPTER I
INTRODUCTION

The desired client-serving radiation modes in outdoor (and indoor) Wi-Fi are determined by known client positions, client density, and the mesh network topology. For high-density deployments, operating in a pure line of sight scenario, a narrow beam, polarization diverse antenna isolates adjacent mesh cells and ensures a reliable link with a client whose up-link polarization is unknown and possibly changing. This scenario is similar to access point (AP) to AP back-haul, where a theoretical point-to-point polarization matched beam is optimal to maintain isolation and increase gain and hence range. Wall-mount APs and strand-mount APs are two other common mounting configurations. The preferred wall-mount radiation mode is hemispherical, much like that of a patch antenna or a sector, while that of a strand-mount AP is omnidirectional. In large urban areas, where the client in a cell may be anywhere in relation to a fixed AP, an omnidirectional radiation mode is employed. It is on this latter case, which is the most common and natural of all Wi-Fi deployment schemes, that we focus.

The growing number of Wi-Fi clients heightens the demands on system capacity and throughput. Dense urban areas, for example, often make use of a mesh network of APs to support these demands. Clearly an AP having more range and
capable of more (reliable) throughput reduces the amount of hardware in the network, which in turn lowers cost, shortens site installation time, and is less intrusive on the local aesthetic. Range is achieved by increasing the effective length of the antenna; capacity is enhanced, most simply, by adding antennas [2]. It has been shown that upgrading from three to four antennas to receive three spatial streams results in a redundancy gain that improves the reliability of the link, helps sustain higher data rates, and also increases range [3]. The fourth antenna helps further combat the multipath fading problem by providing the receiver with an additional information-carrying path to listen to.

Many current outdoor access points utilize a 2X3 (two-input, three-output) or 2X2, two spatial stream MIMO scheme. The supporting integrated antennas are generally monopoles or dipoles. A dual-band monopole is a fine choice if the primary objective is to produce the smallest unit possible. However, as capacity demands necessitate an increase in the number of antennas, monopoles are no longer an attractive choice. This is because monopoles that share the same ground plane are highly coupled, and the position of adjacent elements, the shape of the ground plane, and the stand-alone element pattern essentially determine the uniformity, or lack thereof, of the coverage pattern. Next generation outdoor access points plan to utilize a 4X4, three spatial stream MIMO scheme in an effort to meet contemporary throughput demands. This thesis proposes a low-cost, wideband, omnidirectional antenna to support this requirement. The antenna operates over the 2.4 and 5 [GHz] Wi-Fi
frequency bands and achieves a highly omnidirectional pattern with peak gains of 3 dBi and 6 dBi over 2.4 - 2.5 [GHz] and 5 - 6 [GHz], respectively.

The two most restricting constraints on the performance of the proposed antenna are material cost and antenna length. Since there are four antennas (and four related cable and mounting assemblies) required per AP, the individual cost of the antenna must be low. The design goal specifies that the height of the antenna must be at most 101.6 [mm] (4”). This is so the antennas may be integrated under a radome that in shape and form factor is sleek and proportionate to the size of the unit. However, setting an upper bound on length also sets an upper bound on performance. Since our target length is nearly 80% of a free-space wavelength at 2.4 [GHz], there are sure to be design challenges and compromises, especially in the lower frequency band. Some of the questions that we must address are:

1. How is the feed cable to attach from the radio to the antenna? This transmission line transition is made all the more sensitive by being the feed point of an omnidirectional antenna. The diameter of the coaxial cable that is used and how it is routed around and attached to the antenna all have consequences on H-plane pattern ripple. Also, however the feeding is accomplished, the method must be reproducible so that the antenna may be mass-produced.

2. If a dipole-type element excited in an approximate half-wavelength resonant mode is utilized, how is the surface current controlled so that an axially-directed current may be achieved? In other words, how is the azimuth plane pattern
ripple minimized, assuming an ideal attachment of the feed cable? The fundamental design challenge is to balance and direct the currents on the dipoles so that a uniform magnetic field is set up in the near field.

3. If a corporate-fed array is pursued, how are the elements to be fed to minimize propagation loss and comply with (2)? We seek the feed method that minimizes the length of the on-board transmission lines that feed the elements. This allows for the adoption of a lower cost (higher loss tangent) substrate.

4. How is an element to maintain a 50 [Ω] characteristic impedance over multiple bands? And what measures may be taken to reduce the resonant frequency of a dipole without increasing its length or jeopardizing its radiation patterns?

These inquiries not only form a measure with which we may assess the suitability of various design implementations, but also direct our literature search and guide us through the decision making that defines the early design phase.

Several dual-band antennas with current-balancing structures are available in the literature on printed antennas. He, Wang, and He [4] propose a wideband printed element that is shaped by etching the top surface in a Sierpinski-like periodic pattern they call a periodic cell. The periodic cells modify the current distribution so that a dual-band resonance is obtained. The antenna achieves wide bandwidths in both frequency bands, however the radiation mode in the upper band is undesirable. The dipole is fed using a meandering open-ended microstrip line that balances the currents on the dipole arms and performs the wideband impedance match. The antenna is
large (greater than 76.2 [mm] (3”) tall), especially for application over the S-band, the intended upper band. Zhang and Xin [5] end-load a single dipole element to achieve a dual-band response and suggest a general procedure for the design of dual-band baluns of arbitrary resonant frequencies. The wideband element is reduced to rather narrow resonances at 2.48 and 5.22 [GHz] after the inclusion of the balun. The presence of the balun in the plane of the element also has a detrimental effect on the 5 [GHz] radiation patterns, where in the direction of the balun, in the $x-y$ plane, the pattern exhibits a 40 dB null. A similar feeding structure to [4] is implemented in [6] by Delaveaud and Brocheton to obtain a dual-band response. A microstrip-fed 50 [Ω] line runs along the dipole axis and terminates near the end of one of the poles in an open-circuit. At the lower resonant frequency the length of the open-circuited stub is near a quarter-wavelength. The open-circuited stub forces a near short-circuit condition across a slot separating the poles, which accomplishes the lower band impedance match. In the upper band, the element operates in the full-wavelength resonant mode. The stub now appears as a high impedance transformer across the slot and matches the antenna to the line. The T-shaping of each pole reduces the overall length of the element and facilitates the dual-frequency operation of the antenna. The radiation patterns demonstrate excellent pattern uniformity in both bands. Although it is only alluded to, it is claimed that the cross-polarization discrimination suffers due to radiation from the open-circuited stub. It is this, the innate moderate upper-band impedance bandwidth (12.1%) (due to the variation of the impedance of the open-circuited stub over frequency), and the estimated length that complicate the
direct adaptation of the existing design to Wi-Fi. Chu and Popovic [7] tweak the design proposed in Chuang and Kuo [8] with an aim to reduce the size of the element without reducing the existing impedance bandwidth. The dipoles in [4] and [5] all share identical feeds that balance the dipole currents. A nearby slot cut in a ground plane of a driven microstrip line creates a strip with a current that is opposed to the microstrip return current. After a quarter-wavelength run of line, which divorces the board feed from the dipole feed, the line shorts to the opposite side of the board where a pole of a dipole resides. Since the slot current and driven current are in-phase, the balanced current distribution is obtained. The microstrip return current provides the balanced current on the other pole of the dipole. The element demonstrates a highly omnidirectional pattern with an impedance bandwidth of about 28%. The impedance bandwidth is mainly attributed to the width of the printed element (6 [mm]) as discussed in [7]. It is shown in [7] that nearly the same impedance bandwidth may be achieved for a smaller element by tapering the arms of the dipole in combination with the use of parasitic elements. In general, both of these tactics slow the impedance change with frequency, but can be responsible for other effects that may be less than desirable. The tapering of a dipole arm means that no longer is the element as tall as it can be which results in a lower $Q$-factor, or quality of resonance. This is because a tapered dipole does not confine the current tightly to a path. One of the most succinct and well-developed papers written in this area is that of Lindberg, et al. [9]. There they position two dipoles in close proximity to one another to obtain dual-resonances. The dipoles are sequentially fed by coplanar strips with the
longer element fed first. In the low frequency band the high-band element appears as a shunt capacitor and has little effect on the impedance or radiation pattern of the longer element. In the high frequency band, the high voltage condition at the end of the longer element is now seen across its input terminals, which forces the current up the shorter dipole, which as an input impedance close to 50\,[\Omega]. The current balance is a consequence of the propagating coplanar strip line that feeds the dipoles. The electric field in the gap is directed along the length of the antenna, which depending upon its direction, drives the current up or down the dipole. The wideband impedance match is accomplished by optimizing the electrical lengths and characteristic impedances of the input microstrip line, the coplanar strip line, and the open series stub, which helps facilitate the transition from microstrip to coplanar strips. The balanced feeding structure of Zhang, Chu, and Wang [10] is similar to that described in [9], but employs a defected ground structure summarized by Kim et al. [1]. The defected ground structure enforces open-circuit conditions on the coplanar strip line and its dimensions determine its operating frequency range and bandwidth. The energy propagating along the coplanar strip line couples to a pair of dipoles as in [9], with the longer element fed first and the shorter element fed second. The longer element is terminated in a horizontal stub which reduces the size of the total structure substantially, but perturbs the 5\,[GHz] radiation patterns, particularly in the elevation plane where too much energy is radiated to the sky (or to the ground, depending on the orientation). Still, because of the compact feeding mechanism, the
element, with some modifications, is an excellent candidate to incorporate into a stacked dipole array configuration.

All of the aforementioned designs delegate to some structure the responsibility of balancing the currents on the dipole arms. This does not mean there is an integrated balun necessarily at work. A balun transforms an unbalanced line to a balanced line. A balanced line is a two-conductor transmission line having identical conductors and its line voltages to ground and line currents are equal and 180° out of phase at corresponding points along the line. Take for example the feeding structure found in [10]. Although the method is successful in balancing the currents on the dipole, it is not a balun, since the antenna is still driven between the unbalanced voltage and ground. This is a most prevalent misconception that equivocates the very meaning of a balanced line.

Despite the multitude of efforts put forth to balance dipole currents, there are many designs that disregard this as a crucial design step. Of these, there are a few whose compact shapes to obtain dual-resonances merit their mention. Zhang et al. [11] direct feed a dual-band element that consists of a T-shaped lower frequency element inserted between two shorter dipoles. The direct 50 [Ω] match is accomplished by tapering the element near the feed point and controlling the widths of the slots between the dipoles. Although the element is small and exhibits nice properties, its direct implementation into a corporate-fed array on a single board is not straightforward. Similar designs are found in Quan et al. [12] and Tze-Meng et al. [13] and both suffer from the same apparent limitation. In [12], the dual-band dipole
is spaced off a reflector to provide a directional radiation pattern. The antenna is probe fed and utilizes a capacitative coupling stub to impedance match the antenna in the lower band and couple the energy to the dipole. The upper band resonance is achieved by controlling the width of the coplanar strip line that is coupled to by the input stub. In [13] the complementary design approach to [11] is proposed. Instead of a T-shaped dipole inserted between two shorter dipoles, the 5 [GHz] resonator is placed between two L-shaped 2.4 [GHz] poles that form the element above a folded ground plane. The impedance match is achieved by optimizing the slots between the elements and tapering the element at the feed point, as in [11]. In [11] and [13] there is no mechanism to choke the currents from flowing down the shield and reradiating. Also, since they are both fed directly, the length of shield soldered to the lower pole impacts the flow of current on the pole and how the current reradiates. Although this is an unfavorable consequence of the chosen feeding method, it may be an acceptable tradeoff for designers who are seeking low-cost, low-profile antennas that, out of the gate, have wideband 50 [Ω] resonances, are easily fed, and have solid omnidirectional radiation patterns in both Wi-Fi frequency bands.

Drawing from mainly [7], [9], and [10] the remaining chapters are devoted to stepping through the design of the proposed antenna. In Chapter 2 we discuss the design of a corporate array feed network to feed two dual-band elements in-phase. The presence of the transmission lines, because of their electrical length and because they are current-carrying, alter the impedance looking into an element and likewise, have an effect on the radiation patterns. In Chapter 3 we discuss the design
of a dual-band element to integrate into a stacked dipole configuration. Then, we optimize the H-plane patterns to achieve the lowest possible pattern ripple given our element and configuration. Next we impedance match each element in the presence of the transmission lines. CST Microwave Studio is used to solve Maxwell’s Equations (employing the Transmission Line Method in the time domain) and simulation data is provided as appropriate. A complete set of measured S-parameter and radiation pattern data is provided in Chapter 4. We conclude with a cost analysis of the proposed assembly and in consonance with our findings, advocate its use in future access points.
CHAPTER II
FEED NETWORK DESIGN

The focus of the present chapter is the design of a corporate array feed network. The feed network consists of a coaxial to microstrip transition, a power division, and two tapered line sections that feed two dual-band elements in-phase. Prior to introducing the layout and developing the theoretical concepts necessary for the design, we begin with a discussion of the two basic feeding approaches and identify the pros and cons of each.

2.1 Corporate Versus Serial Feeding and Substrate Selection

Single-band stacked dipole configurations often make use of a serial feed. This is so the feed may run axially along the length of the dipole and the attachment of the feed cable may be kept near the bottom of the board, somewhat hidden from the antenna elements. Also, the pattern ripple experienced by a single dipole due to the presence of the feed-line is canceled by placing another dipole symmetrically about the feed-line as shown in Figure 2.1. This makes for a highly omnidirectional pattern when the dipoles are spaced less than a quarter-wavelength apart. This approach however, has two main limitations: first, an in-phase feeding of the antenna elements happens only at a single frequency, and is thus inherently narrow-band for the given
radiation mode and second, at high-frequency, the choice of the printed circuit board is critical. Long runs or phase-delayed lines yield unequal current amplitudes entering each element producing tilt in the elevation plane pattern and a reduction in overall antenna efficiency. The only way to maintain efficiency is to use a more expensive, low-loss substrate.
Since we are cost-constrained and desire high efficiency and slowly changing radiation modes over both frequency bands (especially over the upper band where the fractional bandwidth is 18%), we pursue an approach that feeds the elements corporately while minimizing the lengths of the traces to do so. This allows us to base the design around a less expensive substrate. We have chosen EM-888 ($\epsilon_r = 4$, $\tan\delta = 0.007$) to serve this purpose. The 0.7112 [mm] (0.028”) thickness of the material is chosen so that the trace widths that may be fabricated guarantee lines whose characteristic impedances fall within the range of 25 – 110 [Ω]. The use of 0.028” thick material also increases the strength of the antenna board and affords it further resistance to mechanical impact, shock, and vibration. An example of a corporate array feeding scheme is given in Figure 2.2.

The lone undesirable consequence of the adopted feeding style is that (by necessity) it resides in the plane of the elements. The presence of the feed line and feed cable increase azimuth plane pattern ripple and differentiate the self and driving-point impedance of the antenna element. In Chapter 3 we explore the effects of the feedline on antenna impedance and pattern in detail.

2.1.1 From the Feed-Point to the Antenna Element

We now present some visual references that aid us in the explanation of various parts of the design. The artwork is provided in Figure 2.3. The antenna is fed using a stripped 1.32 [mm] diameter coaxial cable. The use of a small outer-diameter cable minimizes the length of unshielded center conductor and is less a source of azimuth
Figure 2.3: The artwork of the proposed antenna including the coaxial feed.
Figure 2.4: Close-ups of the modeled coaxial to microstrip transition.

plane pattern ripple. The stripped end has 6 [mm] of the braid exposed, 0.2 [mm] of the dielectric exposed, and a pre-bent and tinned 1.5 [mm] run of stranded center conductor exposed. A close-up of the feed-point is given in Figure 2.4. The braid is soldered directly to the “ground side” of the antenna. The pad is dimensioned so that all 6 exposed [mm] of the braid may be soldered to the pad. This ensures a reliable physical connection. The pre-bent center pin then, being already routed through the non-plated through-hole in the board, is soldered to a short 39 [Ω] pad that allows the solder to collect locally on the pad rather than bleed out onto the 100 [Ω] lines at the T-junction. The 100 [Ω] lines exponentially taper back into a 55 [Ω] line over a 25 [mm] run. Each 55 [Ω] line terminates in a current balun that drives two serially-fed dipoles that achieve the desired dual-band resonances.

The remainder of this chapter is devoted to the design of the feed network. We start with a discussion of the Theory of Small Reflections and its application to continuously tapered lines. From this we extract the pass-band characteristic of the exponentially-tapered line. Once all impedance-controlling parameters have been
initialized, the feed network is optimized in CST Microwave Studio and the simulated S-parameters are provided.

2.2 The Theory of Small Reflections and Exponentially-Tapered Transmission Lines

We begin with a discussion of distributed elements and define the reflection coefficient assuming a sinusoidal time dependence throughout. Then for a single transmission line section terminated in an arbitrary load impedance, we approximate the total reflection coefficient looking into the section. Using this approximation we extend our approach to a multisection transformer that, in the limit as the number of discrete sections increases indefinitely, is a continuously tapered line. This means that the characteristic impedance varies continuously as a function of position along the line. We then consider the case of the exponentially-tapered transmission line and compute its passband profile. The material gathered for this section is compiled from Ida [14], Pozar [15], and Wadell [16].

2.2.1 Distributed Elements and the Matched Condition

Distributed elements are circuit components or transmission line sections whose current and voltage waveforms vary in phase along their length. The variation in phase may be a substantial fraction of a period or it may vary many cycles. At 2.4 [GHz] (and 5 [GHz]) significant changes in phase occur over line lengths as short as a few millimeters. The famous telegrapher equations, developed by Oliver Heaviside, model
how sinusoidal waveforms propagate over transmission line sections or circuit com-
ponents that are electrically long. The time domain form of these equations are

\[
\frac{\partial v(z, t)}{\partial z} = -Ri(z, t) - L \frac{\partial i(z, t)}{\partial t}, \quad (2.1)
\]

\[
\frac{\partial i(z, t)}{\partial z} = -Gv(z, t) - C \frac{\partial v(z, t)}{\partial t}, \quad (2.2)
\]

where \(v(z, t)\) and \(i(z, t)\) are the voltage and current at any point along the line \(z\) at
any instance in time \(t\). The line parameters \(R, L, G,\) and \(C\) are the series resistance,
series inductance, shunt conductance, and shunt capacitance per unit length of the
line, respectively. The series resistance \(R\) is associated with conductive loss due to
the finite conductivities of the conductors. The shunt conductance \(G\) is associated
with dielectric loss as a result of the use of imperfect dielectrics. A perfect dielectric
experiences only the polarization of its molecules in the presence of an applied electric
field and has no free charge that may be directed by the field as a current. Since each
conductor carries a current, each conductor has some serial inductance per unit length
\(L\) and because the driven line is physically separated from the grounded conductor,
it experiences some shunt capacitance \(C\). These parameters completely describe the
electrical nature of the line.

Using (2.1) and (2.2) it is straightforward to obtain the following frequency
domain form of the wave equations for the voltage and current on the line:

\[
\frac{\partial^2 V(z)}{\partial z^2} - \gamma^2 V(z) = 0, \quad (2.3)
\]

\[
\frac{\partial^2 I(z)}{\partial z^2} - \gamma^2 I(z) = 0, \quad (2.4)
\]
where $\gamma^2 = (R + j\omega L)(G + j\omega C) = (\alpha + j\beta)^2$ is the propagation constant and $V(z)$ and $I(z)$ are the voltage and current phasors. The solutions to (2.3) and (2.4) are superpositions of forward and backward propagating waves given by

\begin{align*}
V(z) &= V^+ e^{-\gamma z} + V^- e^{\gamma z}, \\
I(z) &= I^+ e^{-\gamma z} + I^- e^{\gamma z}.
\end{align*}

Using (2.2), $I(z)$ may be expressed as

\begin{equation}
I(z) = \frac{\gamma}{R + j\omega L} [V^+ e^{-\gamma z} - V^- e^{\gamma z}].
\end{equation}

The characteristic impedance of a transmission line is defined as the ratio of the forward propagating voltage wave and the forward propagating current wave as

\begin{equation}
Z_0 = \frac{V^+}{I^+}.
\end{equation}

As a function of the line parameters the characteristic impedance is written as

\begin{equation}
Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}
\end{equation}

which is easily seen from (2.7). In general, $Z_0$ is complex and varies as a function of frequency. For many transmission lines, including microstrip, paired strips, and coplanar strips, it is conveniently safe to assume that $Z_0$ is real and does not depend on frequency. Now suppose a lossless line is terminated in an arbitrary load impedance at $z = 0$. This is shown in Figure (2.5). Because the line is lossless (2.5) and (2.7) simplify to

\begin{equation}
V(z) = V^+ e^{-\beta z} + V^- e^{\beta z},
\end{equation}
Figure 2.5: A lossless transmission line terminated in an arbitrary load impedance.

\[ I(z) = \frac{\gamma}{R + j\omega L} [V^+ e^{-j\beta z} - V^- e^{j\beta z}] \]  

(2.11)
since there is no attenuation and only propagation. At the load \( z = 0 \) the ratio of the backward traveling voltage wave to the forward propagating voltage wave is the reflection coefficient \( \Gamma \). To determine an expression for the reflection coefficient at the load, note that at the load the ratio of voltage to current is \( Z_L \). Evaluating (2.10) and (2.11) at the load and taking their ratio gives

\[ Z_L = \frac{V^+ + V^-}{V^+ - V^-} Z_0. \]  

(2.12)

Rearranging terms yields the following expression for the reflection coefficient at the load

\[ \Gamma = \frac{Z(f)_L - Z_0}{Z(f)_L + Z_0}. \]  

(2.13)
Notice that in general the reflection coefficient is complex and varies (ideally) only with the load which varies with frequency. Also, observe that the only way to eliminate a backward propagating wave is to impose that

\[ Z(f)_L = Z_0. \]  

(2.14)

This is the matched condition and it is equivalent to a forward propagating voltage wave traveling along a line of infinite length where there are no impedance discontinuities to generate reflections.

2.2.2 The Theory of Small Reflections

Now that we have an expression for the reflection coefficient, let’s consider a single transmission line section of characteristic impedance \( Z_0 \) and of electrical length \( \theta = \beta \ell \) terminated in a real load \( Z_L \) on one end and in a long line of characteristic impedance \( Z_1 \) on the other. The scenario is captured in Figure (2.6). We wish to

Figure 2.6: The multiple-reflection viewpoint of a single impedance transformer.
approximate the total reflection coefficient due to the two impedance discontinuities under the assumption that the discontinuities are small. Applying (2.13) at the first discontinuity gives

\[ \Gamma_1 = \frac{Z_0 - Z_1}{Z_0 + Z_1}. \]  

(2.15)

At the second discontinuity the reflection coefficient is

\[ \Gamma_2 = \frac{Z_L - Z_0}{Z_L + Z_0}. \]  

(2.16)

Therefore a quick approximation to the total reflection coefficient is

\[ \Gamma \approx \Gamma_1 + \Gamma_2 e^{-2j\beta\ell}. \]  

(2.17)

The factor \( e^{-2j\beta\ell} \) rotates \( \Gamma_2 \) to account for the incident wave traveling the distance \( \ell \) twice (up the line, then down the line). Now consider an \( N \)-section transmission line consisting of commensurate (equal length) sections. Applying (2.17) to each section gives that

\[ \Gamma \approx \Gamma_1 + \Gamma_2 e^{-2j\beta\ell} + \Gamma_3 e^{-4j\beta\ell} + \cdots + \Gamma_{N+1} e^{-2Nj\beta\ell}, \]  

(2.18)

or

\[ \Gamma \approx \sum_{i=1}^{N+1} \Gamma_i e^{-2(i-1)j\beta\ell}. \]  

(2.19)

Increasing the number of line sections indefinitely over a finite interval \([0, L]\) forces \( \ell \to 0 \) and in the limit gives

\[ \Gamma \approx \int_{x=0}^{L} e^{-2j\beta z} d\Gamma, \]  

(2.20)
which is a general approximation for the reflection coefficient of a continuously tapered line. In [15] \( d\Gamma \) is determined to be

\[
d\Gamma = \frac{1}{2} \frac{d}{dz} \left( \ln \frac{Z(z)}{Z_0} \right) dz
\]

(2.21)

where \( Z(z) \) is the impedance of the taper at any point \( z \) of the transition. Substituting (2.21) into (2.20) provides us with a useful way of computing the reflection coefficient for any continuous transition \( Z(z) \). The result is

\[
\Gamma \approx \frac{1}{2} \int_{z=0}^{L} e^{-2j\beta z} \frac{d}{dz} \left( \ln \frac{Z(z)}{Z_0} \right) dz.
\]

(2.22)

2.2.3 The Exponential Taper

The exponential taper takes advantage of the fact that the natural logarithm resides in the integrand of (2.22). To see that this simplifies the calculation of the reflection coefficient, let

\[
Z(z) = Z_0 e^{az}
\]

(2.23)

be the exponential taper over in the interval \([0, L]\). Evaluating \( Z(z) \) at \( z = 0 \) shows that we are matching a line of characteristic impedance \( Z_0 \) to a line of some other characteristic impedance, say \( Z_1 \). At \( z = L \) we must have that

\[
Z(L) = Z_1.
\]

(2.24)

This forces

\[
a = \frac{1}{L} \ln \left( \frac{Z_1}{Z_0} \right)
\]

(2.25)

which specifies the taper as

\[
Z(z) = Z_0 e^{\frac{1}{L} \ln \left( \frac{Z_1}{Z_0} \right) z}.
\]

(2.26)
Substituting (2.26) into (2.22) produces

\[ \Gamma \approx \frac{1}{2} \int_{z=0}^{L} e^{-2j\beta z} d z \left( \ln \frac{Z_0 e^{\frac{1}{2} \ln \left( \frac{Z_1}{Z_0} \right)}}{Z_0} \right) d z. \]  \hspace{1cm} (2.27)

This simplifies quickly to

\[ \Gamma \approx \frac{1}{2} \ln \frac{Z_1}{Z_0} e^{-j\beta L} \sin \beta L. \] \hspace{1cm} (2.28)

In Figure (2.7) we plot the reflection coefficient versus the line length in wavelengths.

We observe that the nulls in the pass-band occur when \( L = \frac{\lambda}{2} i, \ i = 1, 2, \ldots \), since \( \sin(\beta L) \) vanishes regardless of what line \( Z_0 \) is matched to line \( Z_1 \). Our design matches a \( Z_0 = 100 \ [\Omega] \) line with a \( Z_1 = 55 \ [\Omega] \) line over a line length of 26 [mm]. Computing the exponential growth constant from (2.25) gives

\[ a = -0.02299. \] \hspace{1cm} (2.29)

Figure 2.7: The magnitude of the reflection coefficient varying the taper length.
The voltage reflection coefficient looking into the tapered line section (from (2.28)) at 2.45 [GHz] is

\[ |\Gamma| = 0.0511. \] \hspace{1cm} (2.30)

If we normalize \( V^+ = 1 \) so that the incident power to the tapered section is \( \frac{1}{Z_0} \), then the reflected power down the line of characteristic impedance \( Z_0 \) is

\[ P_{\text{reflected}} = \frac{\Gamma^2}{Z_0}. \] \hspace{1cm} (2.31)

From this we define a power efficiency that does not include attenuation on the line and is solely due to the impedance mismatch looking into the tapered section. It is computed as

\[ \frac{P_{\text{in}} - P_{\text{reflected}}}{P_{\text{in}}} = \frac{1 - \Gamma^2}{Z_0} = 1 - \Gamma^2. \] \hspace{1cm} (2.32)

Substituting 2.30 into 2.32 gives the matching efficiency as

\[ \text{Matching Efficiency} = 99.74\%. \] \hspace{1cm} (2.33)

Therefore, nearly all of the power incident to the tapered section is transmitted to the matched line that feeds each antenna element.

2.3 Distributed Circuit Components and Tuning

Before we discuss the optimization of the feed network and the corresponding results, we first address an important consequence of the chosen feeding scheme. The probe-style feeding method is shown in Figure (2.4). The unshielded center conductor is the circuit equivalent of a series inductor. The series inductance depends on its length
and the operating frequency. Because the run is so short relative to a wavelength at 2.45 [GHz], the inductance is small and does not need to be accounted for. However, over 5 – 6 [GHz] some tuning is required on the board due to its relative length. The next section discusses the design of open and shorted transmission line stubs that are well-suited for this purpose.

2.3.1 Capacitors and Inductors from Transmission Line Stubs

Consider Figure (2.5) depicting a line of characteristic impedance \( Z_0 \) terminated in a load impedance \( Z_L \). Let us suppose that the line is lossless so the propagation constant depends only on the phase constant \( \beta \). The input impedance looking into the section down a length of line \( \ell \) from the load is

\[
Z_{in} = \frac{Z_0 Z_L + jZ_0 \tan \beta \ell}{Z_0 + jZ_L \tan \beta \ell}.
\] (2.34)

From basic circuit theory, a stub that is the distributed equivalent of a lumped capacitor or inductor ought to have a purely imaginary input impedance. One of the ways this is achieved from (2.34) is by setting \( Z_L = 0 \). This gives

\[
Z_{in} = jZ_0 \tan \beta \ell,
\] (2.35)

which clearly has the form \( j\omega L \). Therefore the inductance of the stub is

\[
L = \frac{Z_0 \tan \beta \ell}{2\pi f},
\] (2.36)

where \( f \) is the operating frequency. This is a simple but important result. An inductor depends upon a current to generate its magnetic field. Setting \( Z_L = 0 \) enforces a high
current point at the end of the stub. From (2.35), we observe that stubs less than \( \frac{\lambda_g}{4} \) are inductive and the inductance for a fixed length depends only on the characteristic impedance of the line. This observation will be useful later on when we design a shorted stub to impedance match the antenna.

Similar equations for a distributed capacitor are found by setting \( Z_L = \infty \). This satisfies the zero current (or high voltage) condition at the end of the stub. Simplifying (2.34) under this assumption yields the input impedance of an open-ended transmission line stub as

\[
Z_{in} = \frac{Z_0}{j \tan \beta \ell}.
\]  

Therefore the capacitance generated by the stub is

\[
C = \frac{\tan \beta \ell}{2\pi f Z_0}.
\]

Notice that as the characteristic impedance of the line decreases the shunt capacitance increases. For microstrip or paired strip transmission lines, this is equivalent to an increase of the surface area of the conductors the make up the shunt capacitor.

Prior to applying the results of this section, we note a few well-known and useful facts:

- In common printed transmission lines, it is much easier to fabricate shunt rather than series inductors and capacitors to accomplish impedance matching and tuning.

- To achieve the widest impedance bandwidth as a result of the match, the stub should be placed as close to the load as possible.
• Stubs should be kept as short as possible, especially open-ended stubs, which will radiate if their length is a significant fraction of the operating wavelength.

2.3.2 Tuning

To tune out the series inductance of the feed over the 5 [GHz] band we may either place a shunt inductor near the feed point or move a quarter wavelength away from the feed point toward the load and place a shunt capacitor. A shorted shunt stub that generates a small amount of inductance at 5.5 [GHz] looks like a short circuit at 2.45 [GHz]. Therefore, only the latter option ensures that we do not inadvertently detune the lower band.

To see how this works consider a small open shunt stub that is 1 [mm] in length and 1 [mm] in width. The 1 [mm] width guarantees a microstrip line that has a characteristic impedance of 62 [Ω] for our chosen board thickness and material. At 5.5 [GHz] the shunt capacitance generated by the stub is (from (2.38))

\[
C = 1.095 \times 10^{-13} \, [F] \tag{2.39}
\]

while at 2.45 [GHz] the impedance looking into the stub is

\[
Z_{in, 2.45\text{GHz}} = 1.2 \, [k\Omega]. \tag{2.40}
\]

Thus, the low frequency current is not perturbed by the presence of the shunt stub, while in the upper band the stub’s position, length, and characteristic impedance generate the proper shunt capacitance necessary for tuning.

In both (2.3) and (2.4) we observe the final positions of the shunt stubs. Near the feed point, a shunt stub is placed close to a quarter wavelength away. A
half-wavelength from there another shunt stub is placed (since the line conditions repeat every guided half-wavelength). From there, their position, length, and width are optimized for an acceptable impedance match.

2.3.3 Simulated Results - Feed Network

The line widths, exponential growth constant, and stub positions and lengths are optimized to attain a solid impedance match for our given discretization of the geometry and numerical method. The transmission lines feeding the elements are terminated in a port impedance equal to the line impedance to permit the calculation of insertion loss.

![Graph](image)

Figure 2.8: The simulated VSWR of the optimized feed network.

In Figure (2.8) we plot the voltage standing wave ratio (VSWR) with respect to frequency. The VSWR is the ratio of the maximum voltage on the line to the minimum voltage. For a perfectly matched line there is only a forward propagating
wave, and thus no standing wave, so the minimum and maximum voltages on the line are equal. Therefore a perfect match happens exactly when the VSWR = 1. In general the VSWR is defined in terms of the reflection coefficient as

\[
\text{VSWR} = \frac{1 + |\Gamma|}{1 - |\Gamma|}.
\]

(2.41)

Note that as \(\Gamma \to 1\) the VSWR \(\to \infty\), and there is full reflection from the load back towards the source. In general, a VSWR in the neighborhood of 1.25:1 over both bands suffices and is an acceptable foundation for the design.

![Figure 2.9: The simulated dielectric loss over frequency.](image)

In Figure 2.9 we plot the dielectric loss from the coaxial input port to the antenna element for four frequencies (as illustrated). As is expected, the more cycles traversed by the wave per unit time the more loss it accrues over the same physical length. The input power to the coaxial port is 0.5 \([W_{rms}]\). Thus, less than 4% of the power is dissipated in the substrate. It is apparent that if another feeding approach
were used, this figure would increase two-fold or more and reduce the amount of power supplied to the antenna to be radiated. It is worth mentioning that the losses in inexpensive dielectrics exceed the conductive losses by an order of magnitude. It is for this reason that we omit a plot of the power dissipated in the conductors.

![Graph of the simulated insertion loss from the coaxial port to each element.](image)

Figure 2.10: The simulated insertion loss from the coaxial port to each element.

The insertion loss of each path of the feed network is given in Figure (2.10). Insertion loss is a logarithmic quantity that is related to the transmission coefficient $T$ by

$$IL = -20 \log T.$$  \hspace{1cm} (2.42)

A lossless one-input, two-output power division gives an insertion loss of 3 [dB] per path. Therefore, per path we are losing only 0.2 [dB] at 2.45 [GHz] and 0.4 [dB] at the top end of the 5 - 6 [GHz] band. These are reasonable losses and guarantee a
high level of overall efficiency assuming that the antenna will be matched well to the line and will operate in an efficient radiation mode.

This concludes our discussion of the design of the feed network. We now proceed to Chapter 3 where we develop the antenna element and impedance match it in the presence of the transmission lines.
CHAPTER III
DUAL-BAND ANTENNA DESIGN

Our design of a dual-band element first considers Jaisson’s approximation that accounts for the loading effect of a thin substrate. This allows us to approximate the loaded length of each dipole relative to its free space wavelength. The normalized far-field power patterns are then calculated to determine, under ideal conditions, the expected radiation modes. Then, drawing from Lindberg et. al [9], we feed 2.4 [GHz] and 5 [GHz] dipoles in series to attain a high level of coexistence between the radiators. The balun developed by Kim et. al [1] is utilized to drive balanced currents along the dipoles. Once the feeding scheme is defined, we provide the simulated input impedance and radiation patterns with and without the feed network present and observe its influence on both parameters. Next, we impedance-match the elements in the presence of the feed line and provide the results. In Chapter 4 we compare these results with measured data.

3.1 Jaisson’s Approximation

Loading an antenna with a thin substrate of permittivity $\varepsilon_r > 1$ increases the length of the antenna with little effect on the operating radiation mode. A 70 [Ω] resonant printed dipole in free-space may be reduced in size when loaded so that the resonant
frequency does not change. One consequence of loading a dipole is a reduction in the real part of the input impedance (at resonance), which is linked to the increased effective permittivity of the near-field medium. The effective permittivity may be defined as the value of the permittivity of an infinite homogeneous medium that maintains all of the metrics that characterize the performance of the antenna in the finite loaded case. In effect, it transforms the heterogeneous medium into a homogeneous one.

The effect of loading a printed dipole on a thin substrate is explored in Jaisson [17]. There, Jaisson uses Silvester’s method of partial images (found in [18]) to develop a closed-form approximation for the effective permittivity. The result is valid for strip radiators, or thin (printed) dipoles that satisfy the following criteria:

- The width $w$ of the dipole and thickness of the substrate $h$ are much less than half the dipole length $\ell$.
- The thickness of the copper $t$ is much less than $w$.

![Figure 3.1: A strip dipole loaded by a thin substrate separated by a small gap.](image)
An example of a strip dipole is given in Figure 3.1. Our dipoles are physically bounded as follows: $0 < w \leq 3$, $h = 0.7112$, $t = 0.03556$, and $\ell > 8$ with the units given in millimeters. Jaisson’s approximation allows us to express the effective permittivity over frequency for various strip widths so that we may calculate, for instance, the length of a loaded half-wavelength dipole at $5.5 \text{ [GHz]}$. The effective permittivities for multiple strip widths are given in Figure 3.2, assuming the strips are loaded on $0.7112 \text{ [mm]}$ EM-888. Observe that the effective permittivity increases as a function of frequency. This is intuitive and is a consequence of the board “appearing” thicker at higher frequencies and thinner at lower frequencies. Also, as the trace width increases the effective permittivity decreases. Smaller traces are more enveloped by, or seemingly embedded in, the dielectric. This is observed clearly in Figure 3.3, where the calculation frequencies are $2.45$ and $5.5 \text{ [GHz]}$. We deduce from Figure 3.3

![Image of Figure 3.2: The effective permittivity of a strip dipole for multiple dipole widths.](image-url)
that dipole tapers to increase length are actually not that effective in increasing the electrical length, since the tapered section does not benefit from as much of a size reduction as a thin strip. In spite of this, tapering (or increasing width) does reduce the resonant frequency by extending the current path, as we shall see.

Let’s apply Jaisson’s approximation to our 2.45 and 5.5 [GHz] dipoles. The mean widths over the dipole lengths are 2.28 [mm] and 1.55 [mm] at 2.45 and 5.5 [GHz], respectively. The estimated effective permittivities are $\epsilon_{eff,2.45} = 1.30$ and $\epsilon_{eff,5.5} = 1.46$. Taking the longest current path on each dipole, the effective dipole lengths are

$$\ell_{eff,2.45} = \frac{0.03788}{\lambda_{eff,2.45}} = 0.35$$

at 2.45 [GHz], and

$$\ell_{eff,5.5} = \frac{0.01738}{\lambda_{eff,5.5}} = 0.39$$

at 5.5 [GHz].

Figure 3.3: The effective permittivity versus trace width at 2.45 and 5.5 [GHz].
at 5.5 [GHz], where $\lambda_{\text{eff}, f} = \frac{\lambda_0}{\sqrt{\varepsilon_{\text{eff}, f}}}$, and the longest current paths are 37.88 and 17.38 [mm] on the 2.4 and 5 [GHz] dipoles, respectively. Note that a single frequency point over the 5 [GHz] band suffices, since the effective permittivity only varies a few tenths over the band, regardless of the strip width. The results given in (3.1) and (3.2) set the upper bound on performance. The 2.4 and 5 [GHz] dipoles are small (30% and 22% shorter than half-wave dipoles, respectively) and consequently, we expect input impedances whose real part is less than 73 [Ω] and low gains resulting in high beamwidths. This is the classic tradeoff between antenna size and gain. Now that we have computed the effective dipole lengths, let’s approximate the radiation modes of our dipoles using the classical theory.
3.2 The Radiation Mode of Stacked Short Dipoles

An approximation to the far-field power patterns may be accomplished by assuming an infinitesimally thin, short dipole has the following normalized linear current distribution

\[ I(z') = (1 - |z'|^2 \ell_{\text{eff}}), \]  
(3.3)

where \( z' \) varies along the length of the antenna with the antenna centered at the origin of a coordinate system. Note that the zero current condition at the ends of the antenna is satisfied. Under these assumptions, the far-field electric field may be written as (see [14])

\[ \mathbf{E} = \hat{\theta} j\eta \beta e^{-j\beta R} \frac{e^{j\beta R}}{4\pi R} \sin \theta \int_{-\ell_{\text{eff}}/2}^{\ell_{\text{eff}}/2} (1 - |z'|^2 \ell_{\text{eff}}) e^{j\beta z' \cos \theta} dz'. \]  
(3.4)

Noting that

\[ \int_{-\ell_{\text{eff}}/2}^{\ell_{\text{eff}}/2} e^{j\beta z' \cos \theta} dz' = \frac{2 \sin(\frac{\beta \ell_{\text{eff}} \cos \theta}{2})}{\beta \cos \theta} \]  
(3.5)

and in general

\[ \int_{a}^{b} z' e^{j\beta z' \cos \theta} dz' = \frac{z' e^{j\beta z' \cos \theta}}{j\beta \cos \theta} \bigg|_{a}^{b} + \frac{1}{\beta^2 \cos^2 \theta} \left[ e^{j\beta z' \cos \theta} \right]_{a}^{b}, \]  
(3.6)

(3.7) simplifies to

\[ \mathbf{E} = \hat{\theta} j\eta \beta e^{-j\beta R} \frac{e^{j\beta R}}{\pi R} \sin \theta \left[ \cos(\frac{\beta \ell_{\text{eff}} \cos \theta}{2}) - 1 \right]. \]  
(3.7)

The far-field magnetic field intensity is related to the electric field intensity as

\[ \mathbf{H} = \frac{\mathbf{E}}{\eta}, \]  
(3.8)
where \( \eta \) is the intrinsic impedance of the medium. The Poynting vector at point \((R, \theta)\) is then

\[
\mathbf{P}_{avg} = \frac{1}{2} \text{Re}\{\mathbf{E} \times \mathbf{H}^*\} = \hat{R} \frac{\eta}{2 \pi^2 R^2 \beta^2 \ell_{eff}^2} \sin^2 \theta \left[ \cos(\frac{\beta \ell_{eff} \cos \theta}{2}) - 1 \right]^2.
\]  

(3.9)

Setting

\[
\mathbf{P}_0 = \frac{\eta}{2 \pi^2 R^2 \beta^2 \ell_{eff}^2}
\]

(3.10)

the normalized power pattern is

\[
\mathbf{P}_n = \frac{\mathbf{P}_{avg}}{\mathbf{P}_0} = \hat{R} \sin^2 \theta \left[ \cos(\frac{\beta \ell_{eff} \cos \theta}{2}) - 1 \right]^2,
\]

(3.11)

which varies only with the elevation angle \( \theta \). Using the effective lengths calculated in (3.1) and (3.2), we plot the normalized elevation plane power patterns in Figure 3.5 on a decibel scale. The patterns are nearly identical since the antennas have similar

Figure 3.5: The power patterns of thin wire dipoles of our effective lengths.
effective lengths. The beamwidths are about 84° in both cases, which is a little more
than the 78°-beamwidth of the sinusoidally-driven half-wavelength dipole. Thus, the
shorter dipole has less directivity, as we expect.

The simplest way to radiate more energy to the horizon is to vertically stack
elements. The array factor of two vertically stacked antennas driven in phase is (given
in Balanis [19])

\[ AF(\theta) = 1 + e^{-jkd \cos \theta}, \]  

(3.12)

where \( k = \omega \sqrt{\mu_0 \varepsilon_0} \) and \( d \) is the spacing (from center to center) of the antennas. The
array factor may be normalized by noting that

\[ 1 + e^{-jkd \cos \theta} = e^{-\frac{1}{2} jkd \cos \theta} (e^{\frac{1}{2} jkd \cos \theta} + e^{-\frac{1}{2} jkd \cos \theta}) \]

(3.13)

\[ = 2e^{-\frac{1}{2} jkd \cos \theta} \cos(\frac{1}{2} kd \cos \theta), \]

\[ |e^{j\xi}| = 1 \text{ for all real } \xi, \text{ and the maximum of (3.13) occurs at all odd multiples of } \pi/2 \]

and is 2. Therefore, the normalized array factor is

\[ AF(\theta)_n = \cos(\frac{kd \cos \theta}{2}). \]  

(3.14)

The total electric field may be conveniently written as the product of the element
pattern and the array factor (see [14]). Thus the normalized electric field magnitude
radiated by the dipoles is

\[ E_n = \hat{E} \sin \theta \left[ \cos(\frac{\beta_{eff} \cos \theta}{2} - 1) \right] AF(\theta)_n. \]

(3.15)

The normalized time-averaged radiated power expressed as a function of the field
magnitude is

\[ P_n = \hat{R} |E_n(\theta)|^2, \]

(3.16)
where (3.8) and $EE^* = |E|^2$ were used. The normalized E-plane power patterns varying the element-to-element separation $d$ are provided in Figure 3.6. We assume that the dipole length is $0.3\lambda$. As the distance between the elements increases, the beamwidth decreases. Thus, to attain the maximum gain in the azimuth plane, the elements should be spaced close to a wavelength apart. Our design utilizes a 52 [mm] spacing between the centers of the two elements. This is a spacing of $0.42\lambda_0$ at 2.45 [GHz] and a spacing of $0.95\lambda_0$ at 5.5 [GHz]. The elevation plane power patterns incorporating the effective lengths and the array factor are given in Figure (3.7). The E-plane beamwidths have been reduced to $55^\circ$ and $29^\circ$ at 2.45 and 5.5 [GHz], respectively, as desired. These are the pure radiation modes of our antenna assuming no coupling between any pair of dipoles.

Figure 3.6: The power patterns of stacked thin wire dipoles for various separations.
Figure 3.7: The power patterns of stacked thin wire dipoles of our effective lengths.

Integrated antennas are often mounted above ground planes to divorce the radiating elements from sensitive RF components. In the case of dipoles, this not only improves gain, but also provides a slight elevation tilt in the peak gain location, which may be desirable. We determine the spacing of the stacked dipole above an infinite, perfectly conducting ground plane using an argument from images. However, for the sake of brevity, we do not include simulated or measured data of this case; although it is the author’s experience that the VSWR is unaffected by the introduction of the ground plane (at the appropriate spacing), and in general, the radiation patterns have less ripple and more gain in this scenario.
3.3 The Effects of Ground Plane Proximity on Input Impedance and Pattern

Our dual-band topology is given in Figure 3.8 above a perfectly conducting ground plane. The elements are spaced a distance $d$ from center to center and the element closer to the ground plane is spaced off a distance $s$. Clearly, the capacitative relationship between the lower 2.4 [GHz] dipole and the ground plane determines the impedance variation experienced by the element. However, as we shall see, dipoles are rather tolerant.

![Diagram of dual-band array](image)

Figure 3.8: Our dual-band array placed above an infinite perfectly conducting ground.

The method of images provides a solution for the electric field in the required domain by removing the ground plane and replacing it with a specified number of line currents or point charges at mirrored points in space so that the potential along the plane of the once-removed ground is zero. The solution of the equivalent problem is often much simpler and is valid in the hemisphere above the ground plane only.
The equivalent problem to that in Figure 3.8 is depicted in Figure 3.9, where we consider only the 2.4 [GHz] dipole stack. The dipole may be viewed as two oscillating, electrically connected point charges (illustrated by + and −) that are 180° out of phase. Zero potential is maintained in the plane of ground by mirroring each positive (negative) “charge” by a negative (positive) “charge” across that plane. This results in four line currents all having the same direction. Thus, for an infinite perfectly conducting ground, and the appropriate spacings, a significant gain enhancement on the horizon is possible. In the previous section we demonstrated (see Figure 3.7) that a near-wavelength spacing at 5.5 [GHz] produces an omnidirectional pattern with much of the energy directed to the horizon. Therefore, to maximize H-plane gain in the upper band, given our constrained size, we select that $s = d/2$. The array factor
becomes (see [19])

\[ AF_n(\theta) = \sum_{n=1}^{4} \cos\left(\frac{(2n - 1)kd \cos \theta}{2}\right), \]  

(3.17)

where it is assumed that the current amplitudes are normalized and the elements are all driven in phase. Note that (3.17) maximizes to the number of antennas when \( \theta = \pi/2 \).

![Graph showing real part of the self and driving-point impedances of a 2.4 GHz dipole.](image)

**Figure 3.10**: The real part of the self and driving-point impedances of our 2.4 [GHz] dipole.

The real and imaginary parts of the self and driving-point impedance of our 2.4 [GHz] dipole are given in Figures 3.10 and 3.11. The circular ground plane is one wavelength in diameter at 2.4 [GHz]. It is observed that the input impedance of the dipole at 2.4 [GHz] varies little from one case to the other, with the greatest change being the real part of the input impedance, which increases when the ground plane is introduced. Also, notice that over the 5 [GHz] band, the 2.4 [GHz] element has large
real and imaginary parts to its impedance. The high impedance allows for the dipoles to be serially-fed with minimal interaction between the low and high-band radiators.

In Figure 3.12 we plot the free-space pattern of our 2.4 [GHz] dipole against the driving-point pattern when the antenna is mounted on a one-wavelength diameter ground plane and \( s = 26 \text{ [mm]} \). The pattern is tilted about 38° above the horizon when the ground plane is present. Also, the peak gain increases by about 0.5 dB due to reflections off the ground plane, but on the horizon there is about a 2 dB reduction in gain. Fortunately, the other radiating element helps “knock down” the tilted beam so that the peak gain falls closer to the horizon.
3.4 The Tapered Dipole

Dipole tapers are an effective way to reduce the resonant frequency of an antenna without jeopardizing its radiation beamwidths or radiation efficiency. Consider Figure 3.13. We define the Taper Width as the extra width profited by performing a linear taper. In Figure 3.14 we plot the return loss over frequency for various taper widths. Observe that in Figure 3.14 the % bandwidth in every case is about 10%. As the taper width increases, the $Q$-factor and resonant frequency decrease. Thus, one of the tradeoffs of tapering is, potentially, greater impedance mismatch loss over a wide band. We utilize a 5 [mm] taper width in the design of our 2.4 [GHz] dipole. This balances the width increase of the element with about a 400 [MHz] reduction in
Figure 3.13: The 2.4 [GHz] tapered dipole.

Figure 3.14: The 50 [Ω] return loss of the 2.4 [GHz] dipole over frequency for multiple taper widths.
resonant frequency. Also since the 2.4 - 2.5 [GHz] band is somewhat narrow (< 5%),
there is an excellent chance that a quality impedance match may be sustained. In
its pass band, the 5 [mm] taper spans over 100 [MHz] at a VSWR of 1.4:1 or less.
We also note that in each case the directivities and elevation plane beamwidths were
approximately 1.95 dBi and 84°, respectively, demonstrating that the patterns are
unaffected by small taper widths.

3.5 The Balun

The (current) balun, proposed by Kim et. al [1], is provided in Figure 3.15. It is a

1:1 impedance transformation that depends on the characteristic impedance of the
input microstrip line, the characteristic impedance of the coplanar strip line, and the

Figure 3.15: The microstrip to coplanar strip conversion balances the currents on the
dipoles.

1:1 impedance transformation that depends on the characteristic impedance of the
position and width of the output microstrip line. The balun is utilized effectively by Zhang et. al [10] to feed two dipoles in series and is a convenient way to obtain an axially-directed electric field at the dipole feed point. Consider Figure 3.15. The back of the balun (light grey) is at ground potential since the shield of the feed cable solders there (see Figure 2.4). The propagating microstrip mode shorts to the opposite side of a coplanar strip line that is terminated at both ends in an open circuit. This enforces that the energy propagates in the direction of the dipoles initially, and that finally, the electric field drives the current along the 5 [GHz] dipole. The short creates the potential difference across the strips, which gives direction to the field and the field in the gap drives the current along the dipole. The effectiveness of the capacitative coupling depends highly upon the input impedance of the dipole at a given frequency.

When the balun is used to drive currents along dipoles, the shaping of the rectangular cut-outs that form the open circuits may be optimized so that a wide bandwidth may be achieved in both bands. Even more so than the strip separation of the coplanar strips, the cut-outs control the balun’s current distribution and in combination with the resonant dipoles, determine the overall bandwidth of the element. This is demonstrated in [10]. Also, because of the quick conversion from coplanar strips to obtain the balanced dipole currents, it is mainly the characteristic impedance of the input microstrip line and the dipole impedances that determine the impedance looking into the dual-band element over frequency. Our input line has a 55 [Ω] characteristic impedance and each dipole, because of its small size, has a real part to its input impedance in the range of 50 – 70 [Ω]. Therefore, we expect to be
reasonably matched over both the 2.4 and 5 [GHz] frequency bands right out of the gate.

The equivalent circuits of the balun with the dual-band element at 2.45 [GHz] and over 5 - 6 [GHz] are provided in Figures 3.16 and 3.17. At 2.45 [GHz] the upper band dipole looks like a low-value shunt capacitor. On the other hand, over 5 - 6
the longer dipole has a high input impedance (as noted in Figures 3.10 and 3.11). This is because the longer dipole is approaching a full wavelength at 5.5 [GHz] and the high voltage condition at the end of the dipole is seen now across its input terminals. Therefore, the lower band element looks like an open circuit to the high frequency current. The block in Figure 3.17 represents the $RLC$ network that models the upper element’s impedance variation over frequency.

![Image](image.png)

Figure 3.18: The electric field across the coplanar slot at 2.45 [GHz].

The electric field in the slot viewed from the perspective of the coplanar stripline at 2.45 [GHz] and 5.5 [GHz] is given in Figures 3.18 and 3.19, respectively. The load in each case is the antenna element. It is demonstrated that the field is directed along the length of the dipoles, as desired. In Figures 3.20 and 3.21 we observe the surface current distributions on the balun in both bands. It is noted that the currents on each pole are equal in magnitude and opposite in direction. This
Figure 3.19: The electric field across the coplanar slot at 5.5 [GHz].

Figure 3.20: The balun’s current distribution at 2.45 [GHz].
Figure 3.21: The balun’s current distribution at 5.5 [GHz].

guarantees a low-ripple omnidirectional pattern in the plane around the antenna. The current-balancing property of the balun is demonstrated in Figure 3.22 where the output ports are Port #2 and Port #3 shown in Figure 3.15. The output currents are a near 180° out of phase over a wide band.

3.6 The Elements Driven Simultaneously in Free-Space

The next step in the design is to simultaneously drive both elements and observe the consequences on input impedance and pattern. In general, this is an appropriate place in the design to optimize the radiation patterns of the antenna. The results we present in this section are for an element that is optimal in the sense that it has minimum azimuth plane ripple for its configuration.
Figure 3.22: The 180° phase-shifted currents through the output ports of the balun.

Figure 3.23: Two dual-band elements driven in-phase, alone in space.
The dual-band element in a stack configuration is shown in Figure 3.23. The elements are driven simultaneously in-phase. In Figure 3.24 we plot the simulated azimuth plane directivity patterns at 2.45 and 5.5 [GHz]. The peak directivities are 3.4 and 6.95 dBi, respectively. Also, the mode variation over the 5 [GHz] band is negligible, with a slight decrease in beamwidth over the band being the lone variant.

In Figures 3.25 and 3.26 we plot the elevation plane directivity patterns at 2.4 and 5.5 [GHz]. These compare well with the theoretical patterns provided in Figure 3.7. The beamwidths are about 55° at 2.45 [GHz] and 27° at 5.5 [GHz]. We note that the \( \phi = 90^\circ \) plane is the cut that has the smallest peak directivity over the 5 [GHz] band. This is because the balun and the 2.4 [GHz] dipole reside in this plane. Still, the azimuth plane pattern has a maximum of 3 dB of ripple over the 5 [GHz] band.
Figure 3.25: The simulated $\phi = 0^\circ$ plane directivity patterns at 2.4 and 5.5 [GHz].

Figure 3.26: The simulated $\phi = 90^\circ$ plane directivity patterns at 2.4 and 5.5 [GHz].
Figure 3.27: The simulated return loss over frequency of a single array element.

The return loss of the antenna over frequency is given in Figure 3.27. The reference impedance is 55 [Ω], since this is the characteristic impedance of the input microstrip line. As expected, each element is reasonably matched, however some tuning is required. The lower band resonance needs tuned lower and the upper band resonance needs tuned higher. However, it is not worth impedance matching at this point, since the inclusion of the feed network rotates the input impedance looking into each element. We take this up in the following section.

3.7 Impedance Matching

The setup to perform the impedance matching exercise is shown in Figure 3.28. Each unused transmission line is terminated in a resistor equal to the characteristic impedance of the line. The port impedance looking into each element is 55 [Ω]
and is thus the impedance by which all other impedances are normalized. In Figure 3.29 we plot the reflection coefficient looking into a single port over frequency. We conduct the impedance matching analysis on one port only, since the broadband behavior of the two ports is nearly identical. Observe that at 5.5 [GHz] the antenna is reasonably matched. Therefore, we begin by impedance matching the antenna at 2.45 [GHz] understanding that any acceptable solution must sustain the upper band match. The reflection coefficient at 2.4 [GHz] is (from the plot)

\[ \Gamma = -0.0175 - j0.1987. \] (3.18)

The input impedance looking into the section is

\[ Z_L = Z_0 \frac{1 + \Gamma}{1 - \Gamma} = 49.1 - j20.3 [\Omega]. \] (3.19)
Since it is much easier to fabricate shunt stubs in microstrip, we normalize the input impedance, convert to its admittance, and use the Smith Chart as an admittance chart. The normalized admittance is

\[ y_L = 0.96 + j0.4. \quad (3.20) \]

Moving toward the generator a distance \( d = 0.01\lambda_g \) rotates the input admittance to

\[ y_L = 1 + j0.42. \quad (3.21) \]

Thus, we need a stub having a normalized susceptance of \( b = -0.42 \). The shortest possible stub that provides the required susceptance is a shorted shunt stub. Moving from the short circuit point clockwise to the \( b = -0.42 \) point on the \( |\Gamma| = 1 \) circle gives a total rotation of \( 0.187\lambda_g \), which is equivalent to 11.45 [mm] in the dielectric
at 2.45 [GHz]. Using (2.35), the input impedance of this shunt stub at 5.5 [GHz] is

\[ Z_{\text{in}} = j55 \tan \beta \ell = -j30.3 \, \Omega. \tag{3.22} \]

The result is negative because 11.45 [mm] is greater than a quarter-wavelength in the dielectric at 5.5 [GHz]. In order to not disrupt the match at 5.5 [GHz], the input impedance of the shorted shunt stub must be large. Clearly, the stub designed as is does not meet this qualification. However, if we could generate the “right” amount of inductance at 2.45 [GHz] using a shorted shunt stub that is a quarter wavelength long at 5.5 [GHz], we would have the solution in hand. From (2.36), the inductance of the 11.45 [mm] stub is

\[ L = \frac{Z_0 \tan \beta \ell}{2\pi f} = 8.6 \, \text{nH}. \tag{3.23} \]

We note that a guided quarter-wavelength at 5.5 [GHz] is 6.8 [mm]. Rearranging the terms in (2.36), we calculate the characteristic impedance necessary to generate the 8.6 [nH] of inductance at 2.45 [GHz] using a 6.8 [mm] line as

\[ Z_0 = 2\pi f \frac{L}{\tan \beta \ell} = 157.9 \, \Omega. \tag{3.24} \]

Our board thickness and dielectric constant prohibit us from designing lines that are greater than 110 [\Omega]. Also, since the stub is shorted, the line is terminated in a via. This too prevents us from designing a line that is less than about 0.3 [mm] in width. Nevertheless, we implement the shorted shunt stubs, leaving the stub width and position as a parameters to be optimized.

The simulated return loss looking into the coaxial port of the antenna given in Figure 2.3 is plotted in Figure 3.30. The spectrum allocated for outdoor Wi-Fi is
Specifically 2412 - 2483 [MHz] and 5250 - 5875 [MHz]. The maximum return losses over these bands are about -14 dB and -12 dB, respectively. This translates to VSWRs of at most 1.5:1 over the 2.4 [GHz] band and 1.7:1 over the 5 [GHz] band. Two of the -10 dB (1.92:1 VSWR) crossover points are shown and occur at 3.05 [GHz] and 5.12 [GHz]. It is observed that the antenna has a wide impedance bandwidth in both the upper and lower frequency bands.

3.8 The Simulated Radiation Patterns

The results presented in this section are for the antenna shown in Figure 2.3. Therefore, the results reflect the modeling of a short feed cable. The length of the feed cable is the sole discrepancy between the simulated and measured data. We plot the realized gains over a single angle in three orthogonal planes. The “realized gain”
quantity takes into account all losses, including loss due to impedance mismatch. In short, realized gain is the gain we expect to measure in the chamber.

Figure 3.31: The simulated $\theta = 90^\circ$ plane patterns over frequency.

The inclusion of the feed network results in a greater variation of the radiation patterns over the 5 [GHz] band. So that the reader observes the frequency dependence, the patterns are provided at 2.45, 5.2, 5.5, and 5.8 [GHz]. The azimuth plane patterns are given in Figure 3.31. The feed network is located in the $\phi = 270^\circ$ direction relative to the elements. As expected, this is the direction of maximum ripple. The H-plane pattern ripple is 1, 5, 4, and 3 dB at 2.45, 5.2, 5.5, and 5.8 [GHz], respectively. The peak gains of the antenna at 2.45 and 5.5 [GHz] are 2.9 and 6.5 dBi. The $\phi = 0^\circ$ and $\phi = 90^\circ$ plane patterns are plotted in Figures 3.32 and 3.33. The $\phi = 0^\circ$ plane compares well with the theoretical result given in Figure 3.7. The effects of the feedline on the $\phi = 90^\circ$ pattern are observed in Figure 3.33. Despite the increased
Figure 3.32: The simulated $\phi = 0^\circ$ plane patterns over frequency.

Figure 3.33: The simulated $\phi = 90^\circ$ plane patterns over frequency.
pattern dependence on frequency, a great amount of the energy is radiated to the horizon in each case. In the upper band, the inclusion of the feed network suppresses the energy radiated in elevation angles off the horizon and reduces the peak gain values. This is the chief consequence of our chosen feeding scheme and is certainly tolerable. Recall that in Section 2.1 the feeding scheme was chosen so that a high level of total efficiency might be achieved without the use of an elite substrate material. In Figure 3.34 we plot the total efficiency at 2.4, 5.2, 5.5, and 5.8 [GHz]. The result indicates that although the elements are short and the substrate material is lossy, the antenna radiates 85% of the power supplied to it, nominally. Since the leading cost driver of the design is the substrate material, this is both a high-performance and cost-effective design approach.
CHAPTER IV
MEASURED RETURN LOSS AND RADIATION PATTERN DATA

In our final chapter we present the return loss and radiation pattern data of our proposed antenna over the 2.4 and 5 [GHz] frequency bands. The antenna is fed with standard 1.32 [mm] micro-cable that is terminated in an AMC connector. The antenna is shown in Figures 4.1 and 4.2. A close-up of the feed transition is provided in Figure 4.1: The frontside of the fabricated antenna.

Figure 4.1: The frontside of the fabricated antenna.

Figure 4.3. The radiation pattern measurements are performed in the 3 [m] anechoic chamber shown in Figure 4.4. The gains of the antenna element do not include the insertion loss of the feed cable; however, the effect of the feed cable on the antenna patterns is demonstrated. We note that the antenna’s gain and ripple depend on the
Figure 4.2: The backside of the fabricated antenna.

Figure 4.3: A close-up of the transition at the antenna feed-point.
4.1 Measured Return Loss Data

Return loss is related to the reflection coefficient by

\[ RL = -20\log_{10}\Gamma, \]  

(4.1)

and measures the amount of loss due to the impedance mismatch between the load and the line. In practice, the measured reflection looking into the feed cable is a
superposition (magnitude and phase) of reflections occurring at several points along on the line (as discussed in Section 2.2.2).

The return loss of the antenna is shown in Figure 4.5 over 2 - 6.6 [GHz]. It is measured on an Agilent E5071C network analyzer. The results compare well with the simulated results given in Figure 3.30. The antenna has 10 dB impedance bandwidths of 34% in the low band and 27% in the high band. Over the 2412 - 2483 [MHz] and the 5250 - 5875 [MHz] bands the antenna has a return loss of at most 14 dB. This translates to a VSWR of 1.5:1 or less. This guarantees that the reflected power back down the line to the transmitter is tolerable.
4.2 Measured Radiation Pattern Data

The radiation patterns of the antenna are provided in Figures 4.6 - 4.8. The patterns compare well with the simulated results of Section 3.8. The H-plane patterns are perturbed slightly by the presence of the feed cable, but remain quite circular. The presence of the feed network differentiates the elevation plane patterns shown in Figures 4.7 and 4.8. In the $\phi = 0^\circ$ plane the elements radiate without obstruction and the natural mode calculated in Section 3.2 is closely approached. In the $\phi = 90^\circ$ plane the feed network scatters some of the energy radiated by the elements as expected. Note that the beamwidths in the direction of the feed network ($270^\circ$ on the $\phi = 90^\circ$ plot) are less than the elevation plane beamwidths measured in our other cases. The feed network appears to the respective radiating dipole as a reflector. The effect is most discernable at $2.4$ [GHz], where the elevation plane pattern is similar to a driven

Figure 4.6: The measured $\theta = 90^\circ$ plane patterns over frequency.
Figure 4.7: The measured $\phi = 0^\circ$ plane patterns over frequency.

Figure 4.8: The measured $\phi = 90^\circ$ plane patterns over frequency.
dipole in front of parasitic reflector that is about 110% the length of the driven dipole. At high frequency the pattern undergoes more scatter over the entire $\phi = 90^\circ$ plane, but since the feed network is further away, there is less squinting of the main beam. The measured effects observed in this plane are also witnessed in the corresponding simulation result provided in Figure 3.33. The resemblance of the two results is impressive and demonstrates the power of the CST simulation tool. We enumerate the performance specifications of the antenna in the following section.

4.3 Performance Specifications and Cost

The performance specifications of the antenna are summarized below in Table 4.1. The peak gains are the maximum gains of the antenna and occur in the $\theta = 90^\circ$ plane. We note that the elevation plane beamwidths are the same as those calculated in Section 3.2 using the closed-form result that assumes a linear current distribution.

In general, the use of printed circuit board to design integrated antennas is an expensive approach. However, when the performance demands are high and a single radiating element cannot achieve the desired coverage pattern or range, an array of antennas must be used. The tight tolerances associated with transmission line widths and impedance matching stubs are best controlled using printed circuit board technology, and, as we have seen, the non-unity relative permittivity substrate affords the dipoles some small size reduction that decreases the overall size of the array by an amount proportional to the number of antennas. Therefore, finding a
cost-effective way to leverage the technology is at the core of the proposed design approach.

Table 4.1: Performance Specifications

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Nominal Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operational Frequency Range</td>
<td>2.412 – 2.483 &amp; 5.25 – 5.875</td>
<td>GHz</td>
</tr>
<tr>
<td>Input Impedance</td>
<td>50</td>
<td>Ω</td>
</tr>
<tr>
<td>Return Loss</td>
<td>14</td>
<td>dB</td>
</tr>
<tr>
<td>2.45 [GHz] Peak Gain</td>
<td>3</td>
<td>dBi</td>
</tr>
<tr>
<td>2.45 [GHz] H-Plane Ripple</td>
<td>3</td>
<td>dB</td>
</tr>
<tr>
<td>2.45 [GHz] E-Plane 3-dB Beamwidth</td>
<td>55</td>
<td>°</td>
</tr>
<tr>
<td>5.5 [GHz] Peak Gain</td>
<td>6</td>
<td>dBi</td>
</tr>
<tr>
<td>5.5 [GHz] H-Plane Ripple</td>
<td>4.5</td>
<td>dB</td>
</tr>
<tr>
<td>5.5 [GHz] E-Plane 3-dB Beamwidth</td>
<td>29</td>
<td>°</td>
</tr>
<tr>
<td>Dimensions</td>
<td>101.6 × 40.3</td>
<td>mm</td>
</tr>
<tr>
<td>Board Cost</td>
<td>1.50</td>
<td>USD</td>
</tr>
</tbody>
</table>

The number of antennas in MIMO systems are growing to support ever-increasing capacity demands. A $2 antenna board in a 4 × 4 MIMO system quickly encroaches upon the allocated budget for the antenna platform, which also must include mounting fixtures and cable assemblies for each antenna. If utilizing an antenna array is necessary, then the simplest way to reduce the cost of each antenna
is to select a higher loss tangent substrate material. Assuming 25,000 (4 × 4 MIMO) access points sell per year, the EM-888 antenna board costs $1.50. Therefore, the entire antenna platform may be produced for as little as $11.00, assuming $4.00 for four micro-cables and $0.25 for each mounting fixture. Thus, the affordability, low-profile, and solid performance of the antenna make it an excellent choice to be integrated into present and future MIMO access points.

4.4 Conclusion

Many of the Wi-Fi antennas proposed in the literature are compact, low-gain, and omnidirectional, and as such are well-suited for indoor applications. Clearly, a low-gain antenna (with some downtilt) is favorable, since many deployments utilize a mesh network of access points to increase the coverage area, and less antenna gain further isolates adjacent mesh cells. The proposed design addresses an entirely different problem. The mesh cells of outdoor Wi-Fi are as large as possible within the performance constraints of the system. Thus, the antennas must be designed to achieve excellent range and uniform coverage. The range of an antenna is maximized when its peak gain falls on the horizon. The proposed design radiates a uniform, omnidirectional broadcast pattern with peak gains on the horizon of 3 dBi and 6 dBi over 2.4 and 5 [GHz], respectively, without incurring an appreciable increase in size or cost. The size of the antenna is reduced by tapering the 2.4 [GHz] dipole and impedance matching the dual-band element using a shorted-shunt stub which reduces the low-band resonant frequency and sustains the high-band impedance match. The size reduction of
the antenna also permits the design of a product that is less expensive and has an appearance that is aesthetically pleasing and well-proportioned. Thus, the proposed design retains some of the advantages more commonly associated with indoor access point antennas. Ever-increasing throughput demands motivate an increase in the number of antennas to support MIMO and MU-MIMO (Multi-User MIMO) systems. The proposed antenna, because of its low-cost, small size, and high performance, is a fine choice for cost and size-constrained outdoor access point designs that integrate several antennas to meet these requirements.

4.5 Future Work

It is clear (from Figure 4.8) that feeding the elements is the main challenge in the design of low-ripple omnidirectional antennas. Our approach minimizes propagation loss in the substrate and guarantees an in-phase feeding of the array elements over a wide band. Consequently, the antenna retains a high level of efficiency and its radiation patterns do not vary considerably over frequency. This allows the antenna to be broadbanded in both the 2.4 and 5 [GHz] bands.

The author is unaware of any design that utilizes a purely axial corporate-feed network to feed dual-band elements on a single two-layer printed circuit board. If this could be done within the constraints of impedance matching and obtaining the proper radiation modes, the omnidirectionality of the antenna would depend exclusively on the inherent omnidirectionality of each element. Therefore, it is straightforward to imagine a planar antenna having similar peak gains to the proposed antenna, but less
azimuth plane ripple at the same cost. Of course, the design challenges implicit in
the suggested feeding scheme are left to be uncovered by the pursuer.
BIBLIOGRAPHY


