Novel Closed-Loop Matching Network Topology for Reconfigurable Antenna Applications

Dissertation

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Abstract

As technology progresses, mobile devices such as laptops, tablets, cell phones, and two-way radios have become smaller in size. Consequently antennas become electrically small to fit inside aggressive packaging requirements with rapidly changing real and imaginary impedances. As such, these antennas are very narrow in bandwidth with high-Q and input impedance which is very sensitive to environmental effects. The radiation efficiency of the device is drastically decreased as the antenna is detuned and signal quality is degraded. As the number of mobile devices we use increases, adaptive impedance tuners have and will become a bigger necessity, especially as more radios are integrated into a single device.

This dissertation presents novel improvements to closed loop tuning topologies from a system level perspective addressing impedance tuners, sensing techniques, and how they apply to different antennas. The biggest design hindrance to impedance tuners are losses due to small signal resistance ($R_S$), and loss due to circuit resonances and radiation. A detailed explanation of these loss mechanisms is developed, providing designers with the knowledge to minimize the impact of said losses and improve system efficiency. By exploiting loss mechanisms, a novel small and low cost VHF impedance synthesizer is presented to characterize impedance tuners in load pull measurements.
With full consideration of circuit loss mechanisms, a new directional coupler based tuning topology is presented. Traditional tuning topologies aim to minimize $|S_{11}|$ of the matching network. As demonstrated in this work, such a method has the potential to maximize losses in the circuit, especially in multi-stage tuners. Alternative directional coupler based topologies are presented which maximize the system transducer gain. Furthermore, a novel method of sensing a tuned state through the use of a near field probe that detects far field radiated power is introduced. A design guide is detailed with several examples for use with different types of antennas. Concepts developed in this dissertation are demonstrated in an adaptive tuning system where mechanical means of tuning is applied as a low loss tuner. An electrically small monopole is tuned using the power sensor to provide feedback over a 2.2:1 bandwidth (180 to 400MHz) where at the lowest tunable frequency the antenna is $\lambda/11.3$ in size.
Dedicated to my parents, Kevin and Csilla Smith.
Acknowledgments

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Lastly, thanks be to God for giving me wisdom and guidance through my life. (Romans 3:22-23; Acts 16:31) “For there is no distinction: for all have sinned and fall short of the glory of God, and are justified by His grace as a gift, through the redemption that is in Christ Jesus. Believe in the Lord Jesus, and you will be saved.” (ESV)
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Chapter 1: INTRODUCTION

1.1 Motivation, Challenges, and Objective

Mobile devices continue to become more integrated in our daily lives. As technology progresses, mobile devices such as laptops, tablets, cell phones, and two-way radios have become smaller in size. Consequently, antennas become electrically small in order to fit inside aggressive packaging requirements with rapidly changing real and imaginary impedances. As such, these antennas are very narrow in bandwidth with high-Q [1] and have an input impedance which is very sensitive to environmental effects [2,3]. The radiation efficiency of the device is drastically decreased as the antenna is detuned and signal quality is degraded.

As the number of mobile devices we use increases, adaptive impedance tuners will become a bigger necessity, especially as we integrate more radios into a single device. The explosive rate at which the use of mobile devices is growing is evident in the consumer market of tablets, smartphones, and personal computers. For the first time in 2011, the number of smartphones sold exceeded the number of PCs sold [4] as illustrated in Fig. 1.1 (recreated from [4]). Furthermore Fig. 1.1 illustrates the rapidly increasing rate at which tablet and smartphone sales are increasing. According to [4], this rate will not slow down in the near future as global consumers are shifting
from traditional “dumbphones” (5.6 billion subscribers) to smartphones (835 million subscribers). As such, the smartphone industry has begun integrating automatic impedance tuners into their devices through companies like Wispry, Paratek, and Peregrine, where tunable capacitor arrays are utilized at cell bands to compensate for environmental coupling effects of the handset. As real time tuning technology improves, we will be able to minimize the number of antennas on a handset and exploit better RF spectrum use. While demand for mobile devices like smartphones and tablets is increasing, adaptive antenna tuners are applicable for a wide range of platforms and systems.

Figure 1.1: Global Internet device sales.

Electrically small antennas are integrated in handsets and placed in a multitude of environments as depicted in Fig. 1.2: the device could be worn on the human body, or placed in or on a vehicle, used in an outdoor / indoor environment, or placed
on / near a metallic surface. With these possibilities in mind, it is clear that the matching condition in which the antenna’s input impedance is interfaced with 50 Ω equipment becomes unpredictable, outlining the necessity to obtain broad matching coverage across the Smith Chart. Some automatic matching networks aim to tune one or two matching conditions, such as the loading effects due to the human hand [5]. Specialized matching networks can be designed to account for specific loading environmental effects. However, for many systems, the environment is unknown yielding unpredictable antenna loading, especially for electrically small antennas that have rapidly changing real and imaginary impedance. Therefore, it is necessary to design an agile impedance matching system to account for all possible load impedances. For some applications, communication only takes place over a narrow bandwidth. A reconfigurable matching network is necessary to handle as many points on the Smith Chart as possible at a few frequencies. Some reconfigurable networks have been demonstrated using π matching with J-Inverters [6–8]. However, when the operating bandwidth is wide, impedance matching is typically compromised. For more aggressive bandwidths, a double stubbed tuner was implemented in [9] covering 10 to 20 GHz. However, double stubbed tuners lack broader impedance coverage due to blind spots [10,11]. Simultaneous impedance and frequency agility can be attained for a triple stub tuner configuration [12] and adopted in this work for broad Smith Chart coverage. As the antenna environment is unknown, look up tables to determine a tuning state are not an option making closed loop feedback systems a requirement.

Automatic closed-loop impedance matching is a system level problem with several components that must be designed in tandem. Systems are typically composed of a
In this dissertation, novel improvements are made to adaptive closed looped tuning topologies from a system level perspective addressing impedance tuners, sensing techniques, and how they apply to different antennas. While tuning algorithms are of importance, previous work has already provided many robust solutions. While the VHF/UHF 200 to 400MHz band is the focus for this work, concepts are applicable at all design frequencies. The key contributions of this dissertation are:

- Demonstration of the minimal number of parameters necessary to achieve broad Smith Chart matching coverage for various stubbed tuners is developed in Chapter 2. Design considerations and component limitations for realizable impedance tuners is provided as available components limit achievable performance.
The biggest design hindrance in automatic impedance tuners are losses due to small signal resistance ($R_S$) and loss due to circuit resonances and circuit radiation. It is generally assumed the goal should be to simply minimize losses in matching networks. While this is true, a detailed explanation for loss mechanisms is lacking. Chapter 3 details a new demonstration of how losses are manifested in matching networks. With full consideration of how circuit losses are incurred, a new directional coupler based tuning topology is presented to ensure transducer gain is maximized and validated in simulation / measurements. Traditionally $|S_{11}|$ minimizing topologies are used, as demonstrated in this work, such a topology can maximize the tuners losses, especially for multi-stage impedance tuners. Hence the importance of the alternative topologies presented in this dissertation.
• By exploiting loss mechanisms in matching circuits, Chapter 3 introduces a novel VHF impedance synthesizer to characterize impedance tuners in load pull measurements.

• A unique method of sensing a tuned state through the use of a near field probe that samples far field radiated power for closed loop tuning is presented in Chapter 4. Concepts used in the EMC / EMI fields are adapted to develop a general design guide for properly integrating said near field probes in reconfigurable feedback systems for a number of antennas including monopole, loop, slot, and patch antennas.

• In Chapter 5, a new low-loss mechanical tuner using small motors is designed to tune an electrically small monopole antenna. The integrated near field probe aids in tuning the small antenna over a 2.2:1 bandwidth (180 to 400MHz) where at the lowest tunable frequency the antenna is $\lambda/11.3$ in size. The fabricated prototype validates the low loss mechanical tuner as well as the novel near field probe sensor for feedback.

The dissertation is concluded with a summary of the important contributions and discusses future improvements of adaptive impedance tuning systems.
Chapter 2: RECONFIGURABLE IMPEDANCE MATCHING CIRCUITS

Adaptive matching networks have been demonstrated for antenna matching [6], telecommunication systems and front ends [13], and tunable power amplifiers [14]. As communication systems become more aggressive, said tuners must be able to attain broader impedance coverage over wider frequency bandwidths. As a result engineers must account for simultaneous impedance and frequency coverage, tuning time, power consumption, linearity, impedance sensing, device size, system cost, and circuit losses. This chapter investigates the minimum required parameters to gain full Smith Chart coverage from stubbed tuners. Discussion of realizable component design limitations and considerations is presented as they apply to impedance tuners.

2.1 Impedance Matching Circuits

There are numerous ways to implement switches, capacitors (C), inductors (L), and transmission lines to design an impedance matching circuit. Several are outlined in Fig. 2.1. Matching circuits can essentially be broken into 3 categories. The first being lumped networks (Fig. 2.1(a)-Fig. 2.1(c)), second being transmission line networks (Fig. 2.1(d)), and third being a hybrid of lumped and transmission line networks (Fig. 2.1(e)-Fig. 2.1(f)). For automatic tuning, typically either a lumped
approach or a hybrid approach is taken. The simplest is the Lumped Network (L-Network), such a network consists of either a series-shunt or shunt-series configuration with either a capacitor or inductor. Due to the difficulties in realizing electrically controlled inductors, typically the variable component is some form of capacitance. If variable inductance is necessary, a lumped inductor can be placed in conjunction with the variable capacitor to tune its inductance. While L-Networks are appealing due to their simplicity and small size, they are limited by the possible impedances they can match as only one or two components are used. A shunt-series configuration should be used for impedances inside the $1+jX$ circle on the Smith Chart, while a series-shunt is used for impedances outside the $1+jX$ circle [10].

![Impedance matching networks](image)

Figure 2.1: Impedance matching networks, (a) Lumped Network, (b) π - Network, (c) T-Network, (d) Stubbed Transmission Line, (e) Hybrid Lumped / Transmission Line, (f) Hybrid Lumped / Stubbed.

Broader Smith Chart coverage is attained through a π - Network and T-Network. In [6] a π-Network is able to obtain a wide impedance matching coverage from 380 to 400MHz. π and T networks can be interchanged as one can be converted into
an equivalent version of the other \[10\]. The big advantage of this is flexibility, the ability to choose different component values depending on the desired design. While the lumped networks are useful, they are limited in their bandwidth of operation.

To address wider bandwidths of operation, stubbed tuners are used. Essentially the susceptance of the stub (Fig. 2.1(d)) is varied by loading it with a tunable capacitor (Fig. 2.1(f)). A double stub tuner in \[9\] is used to achieve tuning over a 2:1 bandwidth, although it suffers from forbidden tuning regions inherent to double stubbed tuners \[10\]. Broader impedance coverage is gained through the use of a triple stub tuner and presented as part of the author’s research efforts in later sections.

Ultimately, there are an infinite number of configurations that can be used to achieve a useful tuner. While more matching stages, or distributive networks can provide more matching conditions vs frequency, circuit efficiency can drastically decrease due to added losses from components. Current commercially available variable C or L components have not seen the same leaps and bounds in performance / size as transistor technologies. The designer is fairly limited in what one can accomplish for an aggressive tuning system. The following sections will outline how to obtain impedance / frequency agility out of stubbed tuners, as well as design considerations and limitations one should be aware of due to component characteristics.

2.2 Stubbed Tuning

2.2.1 Triple Stub Tuner Parameters

Stubbed tuning is a common impedance matching approach due to the matching flexibility it provides. Traditionally both shunt and series stubs are discussed. Coplanar waveguide transmission lines (CPW) and microstrip transmission lines are
usually used to implement tuners. As such, series stubs cannot be implemented in these transmission lines in a traditional 2-port configuration, so only shunt stubs will be discussed. Fig. 2.2 will be used to give a brief description of a single, double, and triple stub tuner.

Figure 2.2: Single, double, and triple stub tuner ((a)-(c)), microstrip schematics ((d)-(f)).

Fig. 2.2(a)-(c) depict the traditional schematic representation of a single, double, and triple stub tuner respectively. Fig. 2.2(d)-(f) shows the microstrip equivalent which will be used for future discussion. Clearly the stubbed tuner can be analyzed by using admittances or impedances. The parameters that can be optimized to achieve resonance are depicted in Fig. 2.2(a)-(c): the inter stub spacings ($d_{stub}$), line admittance ($Y_d$ and $Y_s$), stub lengths ($L_{stub}$), and stub terminations which can be Short Circuited (S.C.) or Open Circuited (O.C.). It is noted that for a 50Ω system, where
the stubs are 50Ω, a S.C. stub between 0° − 90° is equivalent to shunt inductance, conversely an O.C. stub between 0° − 90° is equivalent to shunt capacitance.

There is an advantage to having a stub impedance higher than that of the characteristic impedance of the system. For example, consider two different short circuited stubs, each of different length and impedance that are to create a reactance, \( X = 20\Omega \). The input impedance of a short circuited stub, \( Z_{SC} \) is the following:

\[
Z_{SC} = jZ\tan(\beta\ell)
\]  

(2.1)

Where \( Z \) is the impedance of the short circuited stub, \( \beta \) is the wavenumber, and \( \ell \) is the length of the stub. If a stub has an impedance \( Z = 50\Omega \), to create \( Z_{SC} = 20\Omega \), the length must be \( \ell = 21.8^\circ \). However, if the stub has an impedance of \( Z = 100\Omega \), then a length of \( \ell = 11.3^\circ \) is required to obtain \( Z_{SC} = 20\Omega \). In this example it is observed that the exact same impedance is created with different impedance / length stubs. With a higher impedance stub, a shorter transmission line length can be used. This is advantageous to realize a more compact circuit where space is limited.

Another parameter to consider is \( d_{stub} \). If the impedance of \( d_{stub} \) is the same as the characteristic impedance of the system, i.e. 50Ω, then the length of the series transmission line rotates around a constant VSWR circle on the Smith Chart acting like a delay line. The series transmission line acts as an impedance transformer. This is evident through observing the equation for the input impedance (\( Z_{in} \)) of a lossless transmission line:

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

(2.2)

Where \( Z_0 \) is the characteristic impedance of the transmission line and \( Z_L \) is the load impedance. Two examples are compared in Fig. 2.3. Each case transforms a load
impedance $Z_L = 30 - 80 j \Omega$ impedance through a 50° long transmission line. As shown in Fig. 2.3(a), $Z_L$ begins on the VSWR 6.4:1 circle, the impedance of the transmission line is 50Ω, as such $Z_L$ is transformed on the constant 6.4:1 VSWR circle to $Z_{in} = 8.1 - 9 j \Omega$. In this case, the transmission line acts like a delay line. If the impedance of the series transmission line is 100Ω rather than 50Ω, the results will be that of Fig. 2.3(b). In this case, the transmission line acts as an impedance transformer moving $Z_L$ from the VSWR 6.4:1 to a VSWR 3:1 circle where $Z_{in} = 18.4 - 16.7 j \Omega$.

When transforming an impedance through an $n$-length transmission line of higher impedance than that of the system, the transformation “curls” in towards the center of the Smith Chart. This is an appealing characteristic for an impedance matching network as the goal is to reach a VSWR 2:1 or better match. It is noted that a VSWR 2:1 is chosen to define a “matched” impedance as it provides an acceptable 89% efficiency in a lossless system. Of course a VSWR of 3:1, 4:1, or 5:1 (75%, 64%, and 56% efficiency respectively) may be acceptable for many applications.
If any combination of O.C., S.C., stub length \((L_{stub})\), line impedance, and stub spacing \((d_{stub})\) is possible, then any of the circuits in Fig. 2.2 can be optimized to tune any load impedance \((Z_L)\) so long as the real portion is non-zero. In order to picture the agility of each circuit, \(Z_L = 100 + 200j\Omega\) is matched to 50\(\Omega\) and shown in Fig. 2.4.

![Figure 2.4: Matching \(Z_L = 100 + 200j\Omega\) with (a) single, (b) double, and (c) triple stub tuner.](image)

Fig. 2.4 shows how each of the circuits of Fig. 2.2(a)-(c) can match \(Z_L = 100 + 200j\Omega\) with each parameter labeled on the Smith Chart. To aid in comparison the Q ellipses are plotted on the Smith Charts. It is observed that not only do multiple stages yield more matching options (agility), but depending on the matching configuration, one can yield a wider band match by ensuring each matching stage stays inside a particular Q circle. The single stage network reaches a \(Q = 5.2\) (less BW) circle whereas the triple stage network stays inside a \(Q = 2.5\) circle (more BW).
2.2.2 Full Smith Chart Matching Coverage

Single Stub Tuner

While specific tunable matching circuits can be made to handle certain regions of the Smith Chart, there are applications where full impedance coverage on the Smith Chart is desired. Thus it is useful to understand which and how many parameters are necessary for full Smith Chart matching coverage. Analysis of the single stub tuner of Fig. 2.2(d) is seen in Fig. 2.5. The parameters that are varied are $d_{stub1}$, $Z_1$ (impedance of $d_{stub1}$), $L_{stub}$, and whether or not the stub is S.C. or O.C. or both are possible. It is noted that the impedance of the stub is always set to 50Ω. The input impedance of a S.C. stub is $Z_{SC} = jZ_s \tan(\beta \ell)$, and for a O.C. stub $Z_{OC} = -jZ_s \cot(\beta \ell)$ where $Z_s$ is the stub impedance. By inspection of these equations, the input impedance which is fully reactive for a lossless line, is a function of the stub impedance and the stub length. The same reactance can be created by setting $\ell$ constant and varying $Z_s$, or conversely setting $Z_s$ constant and varying $\ell$ (assuming $0^\circ \leq \ell \leq 90^\circ$). So for simplicity, all stubs that will be analyzed maintain a constant impedance of 50Ω, and their length ($\ell$) is varied for tuning. The heat plots represent the $|S_{11}|$ matching condition after a particular impedance was matched. Regions for $|S_{11}| = -10$, -6, and -3dB are marked on each chart.

Each Smith Chart in Fig. 2.5 shows the matching performance contained within a VSWR 50:1 circle for specified parameter constraints described beneath each chart. For Fig. 2.5(a)-(b) full Smith Chart coverage is not attainable when $Z_1$ is 50Ω and the transmission lines are varied between $0^\circ - 90^\circ$ for either a S.C. or O.C. case.
Figure 2.5: Single stub Smith Chart matching inside a VSWR 50:1. Heat plot represents $|S_{11}|$ matching condition after tuning at each impedance.
When $d_{stub1}$ is limited to 45°, impedance coverage is greatly reduced as in Fig. 2.5(c)-(d). Rather, if $d_{stub1}$ can change between 0° − 180°, and $L_{stub}$ is S.C., full impedance matching coverage is attained for a constant $Z_1$, the same is true if the stub is O.C. (Fig. 2.5(e)). If $Z_1$ is allowed to vary between 20Ω − 50Ω, we can attain full impedance coverage when $d_{stub1}$ is between 0° − 120°, where the stubs can be O.C. and S.C (Fig. 2.5(f)-(h)).

A previous study showed that a series transmission line of impedance higher than that of the system “curled in” on the Smith Chart thus aiding in matching. This is true for some impedance / length conditions, however, certain higher impedance conditions will produce a worse match (i.e. 200Ω vs 100Ω) for a fixed transmission line length as they can rotate away from the center of the Smith Chart. Alternately, for a fixed series length and a lower impedance, the transformed impedance moves away from the center of the Smith Chart but also rotates more, i.e. looks like a longer transmission line. In the case of this analysis, for Fig. 2.5(f),(g), a longer series transmission line is necessary to achieve matching. Thus $Z_1$ varying from 20 to 50Ω is required to achieve full matching coverage.

It is clear that obtaining full impedance coverage for a single stub tuner requires a wide range of variability from the stub and distance to the load. Or, the ability to vary $Z_1$ over a wide range of impedances when $d_{stub1}$ is constrained between 0° − 120°. As such, the single stub tuner is not the best choice for a tuning circuit with full Smith Chart coverage as it is difficult to realize the required variable parameters.

**Double Stub Tuner**

The double stub tuner matching performance contained within a VSWR 50:1 circle is shown in Fig. 2.6. It was observed that the distance of the stub to the load
for the single stub tuner was a crucial parameter that must be varied to obtain full
Smith Chart coverage; this is not the case for a double stub tuner. For most tuners,
the distance from the load and first stub typically is not variable. As such, the double
and triple stub analysis will have a static length of $d_{stub1} = 10^\circ$ between the load and
the first stub. Practicality dictates that there is some delay line between the load
and the first stub which is why it is included. All transmission lines are 50Ω unless
otherwise noted for the impedance $Z_1$ of $d_{stub2}$ which is varied in Fig. 2.6(e),(f).

If both the inter-stub spacing ($d_{stub2} = 0^\circ - 90^\circ$) and the stub lengths ($L_{stub1,2} =
0^\circ - 90^\circ$) can be varied, we see that full Smith Chart coverage is obtained when
the stubs are O.C. (Fig. 2.6(a)). However, when the stubs are S.C. a significant
number of impedances are un-matched (Fig. 2.6(b)), it is not immediately apparent
why the O.C. case can match all impedances while the S.C. case cannot, but this
will be discussed later in further detail. When only the stubs are varied in length
($L_{stub1,2} = 0^\circ - 90^\circ$), and are O.C. or S.C. as in Fig. 2.6(c) and Fig. 2.6(d) respectively,
there are unmatchable regions on the Smith Chart. It is interesting to note that
Fig. 2.6(b) and Fig. 2.6(d) yield identical performance. This is better understood
by examining a step-by-step matching approach of Fig. 2.6(a) in Fig. 2.7(a), and
Fig. 2.6(b) in Fig. 2.7(b) where $Z_L = 10 + 15j \Omega$. An O.C. matching stub less than
90° is equivalent to a shunt capacitance, as such, it rotates clockwise on a constant
conductance circle. A delay line between stubs or the load rotates clockwise on a
constant VSWR circle (Fig. 2.7(a)). It is intuitive that for the final O.C. stub, the
impedance must be transformed to the upper half of the Smith Chart on a constant
conductance circle of 0.5 to 2 (normalized, not depicted).
Figure 2.6: Double stub Smith Chart matching inside a VSWR 50:1. Heat plot represents $|S_{11}|$ matching condition after tuning at each impedance.

- **(a)** $d_{stub2} = 0^\circ - 90^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (O.C.)
- **(b)** $d_{stub2} = 0^\circ - 90^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (S.C.)
- **(c)** $d_{stub2} = 90^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (O.C.)
- **(d)** $d_{stub2} = 90^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (S.C.)
- **(e)** $d_{stub2} = 90^\circ$, $Z_1 = 20\Omega - 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (O.C.)
- **(f)** $d_{stub2} = 90^\circ$, $Z_1 = 20\Omega - 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (S.C.)
- **(g)** $d_{stub2} = 45^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (S.C.&O.C.)
- **(h)** $d_{stub2} = 90^\circ$, $Z_1 = 50\Omega$, $L_{stub1,2} = 0^\circ - 90^\circ$ (S.C.&O.C.)
$d_{stub2}$ may not be long enough ($0^\circ - 90^\circ$) to transform the impedance to the necessary location before the final O.C. stub; since the first O.C. stub rotates clockwise, it can be used in tandem with $d_{stub2}$ to move $Z$ to the proper region before the final stub. Thus, matching to a VSWR of 2:1 or better is achievable. However, when the stubs are limited to $0^\circ - 90^\circ$ in length and are S.C. (Fig. 2.6(b) and Fig. 2.7(b)), matching is not always feasible. A S.C. stub less than $90^\circ$ in length is equivalent to a shunt inductance and moves counter-clockwise on a constant conductance circle.

Figure 2.7: Step-by-step matching of $Z_L = 10 + 15j \Omega$ for (a) O.C. stubs $d_{stub1} = 10^\circ$, $L_{stub1} = 41^\circ$, $d_{stub2} = 4.5^\circ$, $L_{stub2} = 34^\circ$, and (b) S.C. stubs $d_{stub1} = 10^\circ$, $L_{stub1} = 70^\circ$, $d_{stub2} = 90^\circ$, $L_{stub2} = 66^\circ$.

For a S.C. $L_{stub2}$ to finish matching $Z_L$, $Z_L$ must be transformed to the bottom half of the Smith Chart on a constant conductance circle of 0.5 to 2 (normalized, not depicted). Since the S.C. stubs move counter-clockwise on the Smith Chart, and the series delay line between the stubs ($d_{stub2}$) moves clockwise on the Smith Chart, rather than working in tandem as in the O.C. case, they work against one another. As such, $L_{stub1}$ in Fig. 2.7(b) cannot help $d_{stub2} = 90^\circ$ move $Z_L$ into the proper region on the
bottom half of the Smith Chart. As such $L_{stub2}$ cannot match $Z_{in}$ to 50Ω. In order for the matching network in Fig. 2.6(b) with S.C. stubs between $0^\circ - 90^\circ$ to match all load impedances on the Smith Chart, $d_{stub2}$ must provide a positive delay rather than a negative delay, or vary between $90^\circ - 180^\circ$. Since such a variable matching network would be much larger in size, it would be better to use an O.C. rather than a S.C. double stubbed matching network. It is for these reasons that Fig. 2.6(b) and (d) are the same. Even though Fig. 2.6(b) allows for a variable inter stub length, it does not extend to $90^\circ - 180^\circ$, thus its best performance is bounded by the longest inter stub length which is $90^\circ$. Fig. 2.6(d) has a static inter stub spacing of $90^\circ$ and thus yields the same performance. If Fig. 2.6(b) was bound to $0^\circ - 80^\circ$, it would have poorer matching coverage than that of Fig. 2.6(d).

Moving on, it is seen that Fig. 2.6(e),(f) have similar matching coverage where $Z_1$ is varied between $20\Omega - 50\Omega$, but does not yield full matching coverage. While with a wider $Z_1$ tuning range, full coverage is feasible, it is not practical as low impedances relate to very wide traces which may not be feasible in design. Fig. 2.6(g),(h) allows for both S.C. and O.C. $L_{stub1,2} = 0^\circ - 90^\circ$ tuning range and depicts the forbidden matching region (region where impedances cannot be matched) inherent to double stub tuners with a static $d_{stub2}$. It is also seen how the size of the forbidden region can be made larger or smaller at a particular frequency depending on the length of $d_{stub2}$. It should be noted that it is also difficult to realize a variable stub length that can be made O.C. and S.C., typically the stub is either one or the other, not capable of being both. Not shown is the case where one stub can only be S.C. and one only O.C. .
In summary, for full Smith Chart coverage from a double stub tuner, it is necessary to have both variable stub lengths and inter-stub spacing as shown in Fig. 2.6(a).

**Triple Stub Tuner**

Triple stub analysis is shown in Fig. 2.8 where performance is contained within the VSWR 50:1 circle. With the added tuning stage, less parameter variance is required to obtain full coverage for a single frequency unlike the single and double stub tuning cases. For all of these simulations, the impedance for all transmission lines (stubs and inter-stub spacing) is 50Ω. In Fig. 2.8(a),(b) \( L_{stub1,2,3} \) are O.C. and S.C. respectively and vary between \( 0^\circ - 80^\circ \). There are several points on the perimeter of the Smith Chart that are unmatched. Full coverage is realized by extending the stub lengths to \( 0^\circ - 90^\circ \) for both circuits. It is interesting to note that \( d_{stub2,3} \) is much shorter for the O.C. case as opposed to the S.C. case. The explanation for this is the same from section 2.2.2, the S.C. stub rotates counter clockwise, while \( d_{stub2,3} \) rotates clockwise. Thus a longer \( d_{stub2,3} \) is required for a S.C. stub to obtain full matching agility.

Fig. 2.8(c),(d) are useful in determining the minimum variable length to realize full matching coverage. While each of the depicted simulations misses a few points on the edges of the Smith Chart, it is found that for full coverage, Fig. 2.8(c) requires \( L_{stub1,2,3} = 50^\circ - 90^\circ \) (40° of variation), and Fig. 2.8(d) requires \( L_{stub1,2,3} = 10^\circ - 90^\circ \) (80° of variation). Thus full coverage is achievable with less variability than that of Fig. 2.8(a),(b).

While it is difficult to create a tuning circuit that can have both S.C. and O.C. termination, there are other ways to create the same admittances with only a S.C. or O.C. termination. Fig. 2.8(g) shows the variability of a single stub from Fig. 2.8(f) in a 50Ω system on the constant admittance circle \( (Z_L = 50\Omega) \). The \( S.C. = 0^\circ - 20^\circ \)
variation on the top half of the Smith Chart highlighted in blue. Then the $O.C. = 70^\circ - 90^\circ$ variation on the bottom half of the Smith Chart highlighted in red. Rather than changing the stub termination, one could alternatively use an $S.C. = 160^\circ - 200^\circ$ stub, or $O.C. = 70^\circ - 110^\circ$ stub. Thus capacitive and inductive tuning can be realized without changing the stub termination. This is important to note as it will be discussed in the following sections where a varactor loaded S.C. stub is able to achieve the same type of variability. Again, full coverage is achieved for Fig. 2.8(f).

Typically variation in the tuner will be realized in the stubs, not the inter stub spacings. While varying inter stub spacing and transmission line impedances will give more degrees of freedom to match across a wide band, they are not necessary for single frequency Smith Chart coverage. To achieve full coverage at a single frequency, only the stubs need to be varied. The circuit from Fig. 2.8(c) need only be modified to allow the O.C. stub to vary $40^\circ$ between $50^\circ - 90^\circ$. As such, the triple stub tuner is more attractive for agile tuning than the single or double stub tuner.
Figure 2.8: Double stub Smith Chart matching inside a VSWR 50:1. Heat plot represents $|S_{11}|$ matching condition after tuning ((a)-(f)) where $d_{stub1} = 10^\circ$ and T-line $Z = 50\Omega$ and (g) single $L_{stub1,2,3}$ representation of (f) with $Z_L = 50\Omega$. All impedances matched to $|S_{11}| = -12.7\text{dB}$ or better.
2.3 Design Considerations and Component Limitations

Several conventional tunable components for impedance matching networks are summarized in this section. The first are semiconductor varactor diodes and have been widely used [7,15] for a variety of applications. Ferroelectric varactors have also been used as low as 850MHz and above 10GHz and are suited for higher power applications [16–18]. Micro-electro-mechanical systems (MEMS) have been integrated into many tuning systems, especially for MMIC applications [5,9,11,12]. MEMS tunable capacitors are also being developed by companies like Wispry, Paratek, and Peregrine Semiconductor for commercial cellular phone integration. MEMS are typically implemented to act as a switch on a transmission line or in a tunable capacitor bank. Likewise, PIN diodes have been used in the same manner [6]. Another way to implement an RF switch is through the use of a transistor (FET switch) [19]. Alternatively, larger mechanical RF switches can be used as well. Each technology has its own strengths and weaknesses, performance criteria are summarized in Table 2.1 (compiled from [20–23]). Table 2.1 can be broken into two types of technologies, MEMS switches and semi-conductor devices. In general MEMS strengths include near-zero power consumption, high isolation, low insertion loss and low inter-modulations as they are highly linear devices with up to 30-60dB improvements in distortion when compared to FET switches or PIN diodes [24]. Being a mechanical device rather than a semi-conductor, MEMS switches have a lower lifespan, typically tens of thousands of cycles, although recent improvements have extended their lifespan to 1 million cycles [25]. Furthermore, they suffer from packaging difficulties which increase their price and require high operating voltages making drive circuits (charge pumps) necessary. Component parameters of interest include DC bias current, DC bias voltage, linearity,
tuning range (Cmin/Cmax), loss (Rs), and RF power handling. Since the author’s research has involved hyperabrupt junction varactor diodes, design considerations using varactor will be discussed further.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>RF MEMS</th>
<th>PIN</th>
<th>FET</th>
<th>Varactor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage [V]</td>
<td>20-90</td>
<td>3-5</td>
<td>3-5</td>
<td>5-40</td>
</tr>
<tr>
<td>Power Consumption [mW]</td>
<td>0.05-0.1</td>
<td>5-100</td>
<td>0.05-0.1</td>
<td>0.01-0.2</td>
</tr>
<tr>
<td>Switching Time</td>
<td>1-300 μs</td>
<td>1-100ns</td>
<td>1-100ns</td>
<td>1-150ns</td>
</tr>
<tr>
<td>Loss [dB]</td>
<td>0.05-0.5</td>
<td>0.3-1.5</td>
<td>0.4-2.5</td>
<td>0.1-2</td>
</tr>
<tr>
<td>RF Power Handling [W]</td>
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<td>&lt;10</td>
<td>&lt;10</td>
<td>&lt;5*</td>
</tr>
<tr>
<td>Linearity</td>
<td>excellent</td>
<td>good</td>
<td>good</td>
<td>poor*</td>
</tr>
</tbody>
</table>

1Includes voltage up-converter and or drive circuitry.
*Can be improved with various configurations.
Table reports typical values.

Varactor diodes have many attractive qualities for impedance tuning applications. There are numerous commercially available low-cost varactors that come in a variety of packaging configurations. Also, varactors are solid-state components which are based on a mature technology and have high reliability. The two most important inter-related design parameters for impedance tuning are the quality factor (Q) and associate capacitance ratio Cmax/Cmin. Fig. 2.9 shows the simulated Q vs Frequency relationship for a typical hyperabrupt junction varactor. Different curves correspond to different varactor bias voltages. The Q value decreases significantly as frequency increases (and as the bias voltage decreases). At 400MHz and a high bias voltage, relating to a low capacitance of 3pF, a Q of ≈7500 is observed. By lowering the bias voltage and raising the diode capacitance just 1pF to 4pF, Q drops by a factor of 10 to 750. As the bias voltage decreases to a large capacitance of 30pF, Rs dominates.
resulting in a very low Q of 13.5 and the varactor begins to appear more like a resistor than a capacitor at 400MHz. In an ideal matching network, Q should be very high to minimize loss. Therefore it is important to set a lower limit on Q which in turn sets the capacitance ratio. In switched inductor or capacitor arrays using FET’s, PIN Diodes, or MEMS switches, a tradeoff between circuit footprint and complexity would limit the tunable range.

Figure 2.9: Typical Q variation with frequency and bias voltage of a hyperabrupt junction varactor. Q decreases as frequency increases and bias voltage decreases relating to higher circuit loss.

Fig. 2.10 shows a simple L/C circuit with a large shunt ideal inductor and a single or multiple shunt varactors with Fig. 2.9 properties. The objective is to use the varactor to tune out the inductance and maximize transmission (|S_{21}|) at 300MHz. In the single varactor case, the capacitance must be large enough to cancel the inductance.
which results in a maximum $|S_{21}|$ of -4.53 dB due to a low $Q \approx 6$ observed in Fig. 2.10. However, when using four varactors in parallel to achieve the equivalent capacitance, each varactor operates at a higher bias level with much higher $Q \approx 30$. This results in a maximum $|S_{21}|$ of -1.1 dB. Thus, depending on the required capacitance ratio, placing multiple varactors in parallel will exponentially increase $Q$ and improve the efficiency of the circuit. Therefore an acceptable trade off between component count, circuit complexity, tuning ratio, and $Q$ must be assessed when designing a matching network with varactor diodes.

![Figure 2.10](image)

Figure 2.10: One way to increase $Q$ by using multiple shunt varactors instead of a single varactor.

When it comes to selecting components, often times the manufacturer only provides $Q$ or small signal resistance ($R_S$), not both. It may be tempting to select the
component with the lowest Q assuming it will be the part with the least amount of
loss. Quality factor can be calculated from the following equations.

\[
Q = \frac{|Im(Z)|}{R_S} \tag{2.3}
\]

For Capacitors: \(Q_{cap} = \frac{1}{2\pi f C R_S}\) \(\tag{2.4}\)

For Inductors: \(Q_{ind} = \frac{2\pi f L}{R_S}\) \(\tag{2.5}\)

In short, from equation (2.3), Q is the ratio of the reactive (stored) energy to
the dissipated energy. Any energy dissipated by our components will be in the form
of heat, i.e. loss. Equations (2.3)-(2.5) are used to complete the data in Fig. 2.11
where three commercial lumped components are analyzed at 300MHz. The ceramic
capacitor has the highest Q of 212, two wire air-core inductors with Q’s of 177 and 70
are also quoted. The component with the highest Q (ceramic capacitor) also has the
most loss with \(R_S = 0.417\Omega\). Conversely the component with the lowest Q has the
least amount of loss of \(R_S = 0.07\Omega\) for the 0906-3 inductor. If only Q were observed,
it may be tempting to say the capacitor would have the least loss which is not the
case. Quality factor is a ratio of reactance to resistance, hence a higher Q would not
necessarily mean lower loss, it may only mean that the reactance is quite high relative
to the resistance, as is the case for this example. While it is true we want energy
stored in these components, we also want the loss to be as low as possible. Thus the
designer must choose the component with the lowest \(R_S\), not the component with the
highest Q.
### Analysis at 300 MHz

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
<th>Q-Factor Equation</th>
<th>Quality Factor</th>
<th>$R_S$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>DiLabs C17AH006 Capacitor</td>
<td>6pF</td>
<td>88.4</td>
<td>212</td>
<td>0.417</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.417</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Coilcraft A05T Inductor</td>
<td>18.84nH</td>
<td>35.5</td>
<td>177.5</td>
<td>0.20</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.2</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Coilcraft 0906-3 Inductor</td>
<td>2.58nH</td>
<td>4.86</td>
<td>70</td>
<td>0.07</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.07</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure 2.11: Comparison of Q and $R_S$ for different commercial components.

Matching network size is of concern, especially if an electrically small antenna requires adaptive tuning. The stubbed tuning approach requires $n$-length transmission lines which can be of appreciable electrical size. As explained previously, for stubs of electrical length less than 90° a short circuited stub provides inductance, and an open circuited stub provides capacitance. As such, a short circuited stub and open circuited stub could be replaced by a lumped inductor or capacitor respectively. The series transmission line between the stubs acts as an impedance transformer of some length / impedance. A series transmission line can be replaced with a phase shifter in the form of a π-network. An example of these concepts are illustrated in Fig. 2.12. The simulations for each model were implemented in microstrip with realistic losses. A few impedances on the VSWR 20:1 circle are chosen to analyze the difference between the two circuits. After an impedance is selected, it is set as $Z_L$.
for each circuit, the capacitor banks in the circuit model are optimized to tune to a perfect match, and circuit loss is recorded in dB. Without any matching network the miss-match loss would be 7.4dB for any impedance on the VSWR 20:1 circle. For the transmission line circuit (Fig. 2.12 top), a worst case loss of 1.3dB is observed. Replacing the transmission line model with equivalent realistic lumped components (and associated $R_s$, Fig. 2.12 bottom) yields up to 7.2dB of loss. The losses from all the lumped components add up quickly and create a very lossy circuit. While this poorer performance is indeed from $R_s$, the loss manifests itself dominantly due to circuit standing waves and resonances as will be discussed in Chapter 3. This study again illustrates the trade-off between circuit size, complexity, and loss. Component losses can greatly hinder performance and must be considered during design.

Continuing discussion on component design quantities, one drawback of varactors is the requirement of a wide range of control voltages which could be in the range of 5 to 40 volts. In higher power applications commonly used in power amplifiers or VCO’s, the maximum voltage can be on the order of 100 volts. A varactor’s ability of providing continuous capacitance variation is an advantage over other technologies. In addition, varactors also feature low current draw which is typically on the order of nano-amps, making them suitable for battery operated devices.

Varactors are generally non-linear devices which can cause inter-modulations, especially in higher power transmit applications. Although this is avoidable with proper design considerations summarized in [26,27], this is generally not an issue with MEMS technology. Depending on signal integrity requirements, the non-linearity can limit the RF power handling of the varactor. Typically, the maximum RF input power rating is not provided in data sheets, and thus must be investigated depending on
Figure 2.12: Loss comparison for triple stub varactor loaded tuner (top) transmission line implementation (bottom) lumped circuit equivalent.
the application. However, as the transmit power is increased, RF voltage may further bias the varactor especially when it is operated in a low DC bias region. Thus operating voltage range is certainly a factor in selecting a device based upon the desired transmit power.
Chapter 3: IMPROVED CLOSED LOOP TUNING
TOPOLOGY FOR IMPEDANCE TUNING

3.1 Realizing Tunability With Commercial Components

Thus far analysis has been conducted assuming that a circuit can be varied in a number of favorable ways. However, it can be quite a challenge to realize a desired tunable range from a stubbed tuner. Usually it is desired to electronically control variable components in a circuit to achieved a tuned state. As such, it is important to understand how to implement commercially available components in a realizable tuning circuit. One approach is to load the transmission line stubs with a variable capacitor or inductor as in [9], alternatively switches can be distributed across the length of the stub shorting to ground in different locations [28]. Since low loss commercially available electronically controlled inductors are hard to come by, variable capacitors are typically used, or switched inductor arrays.

As discussed previously, changing the length of a stub changes the susceptance of the line and provides the ability to change equivalent shunt capacitance or inductance depending on if the shunt stub is short or open circuited for a line length less than 90°. Variability is achievable by placing a varactor diode on the shunt stub as depicted in Fig. 3.1(a)(c).
For the loaded short circuited (S.C) shunt stub, the susceptance of the transmission line is:

\[ B = \frac{j\omega C}{1 - Z_t\omega Ct} \]  

(3.1)

Where \( t = \tan(\beta \ell) \), \( \beta \) is the phase constant, \( \ell \) is the length, and \( Z_t \) is the impedance of the transmission line which will be 50\( \Omega \), and \( \omega = 2\pi f \) the angular frequency. The
admittance looking into the circuit is:

\[ Y_{in} = G_L + (B + B_L) \]  \hspace{1cm} (3.2)

Where \( G_L \) and \( B_L \) are the load conductance and susceptance. Likewise for an open circuited stub (O.C.) the susceptance of the transmission line is:

\[ B = \frac{j\omega Ct}{t + Z_t C\omega} \]  \hspace{1cm} (3.3)

The S.C and O.C. stubs without the capacitor are plotted in Fig. 3.1(b)(d), where \( \ell = 60^\circ, f = 300MHz \) and compared to their capacitor loaded counter parts. The capacitance is varied from 1pF to 30pF, and \( Z_L \) is set to be fully real at 50Ω. A S.C. stub of length 60° results in a shunt inductance as seen in Fig. 3.1(b)(d) (red marker). By adding the capacitor in series on the shunt stub, and varying it over the stated range, it is seen that both variable capacitance and inductance is achieved on the constant conductance circle. This continuously electrically controlled variable capacitance / inductance is greatly desired as it contributes a large amount of tuning agility to the circuit. Conversely, the equivalent O.C. stub results in shunt capacitance. By adding the capacitor, tuning through the shunt capacitance region on the constant conductance circle is achieved (Fig. 3.1(d)). However, this is already feasible with only a shunt capacitor without the O.C. transmission line. Thus we observe the benefits of utilizing a short circuited stub over that of the open circuited stub when the transmission line is loaded with a variable capacitor. Another advantage of a SC stub over a OC stub is that SC stubs are less susceptible to undesired radiation.

As stated in Chapter 1, one goal of this research is to develop a frequency and impedance agile tuning system from 200 to 400MHz. Thus the impedance tuner should cover as much of the Smith Chart as possible. It is known from [10] that while a
double stub tuner is effective, the spacing between stubs creates a forbidden matching zone. To address this, a microstrip triple stub tuner is developed to maximize the impedance agility of the circuit. However, the inter stub spacing and physical stub lengths limit the frequency agility. To achieve full 200 - 400MHz coverage, a two state microstrip phase shifter is necessary to adjust stub length thus covering the full band as illustrated in Fig. 3.2.

![Diagram of triple stub microstrip layout with meandered microstrip phase shifter 200-400MHz.](image)

Figure 3.2: Triple stub microstrip layout with meandered microstrip phase shifter 200 - 400MHz.

Where for State 1, $\ell_{1\text{stub}} = 60^\circ$ and $\ell_{1\text{series}} = 45^\circ$ at 250MHz. For State 2, $\ell_{2\text{stub}} = 60^\circ$ and $\ell_{2\text{series}} = 45^\circ$ at 350MHz. The microstrip phase shifter is realizable using a single-pole double-throw switch properly placed on a meandered microstrip trace. All of the transmission lines in Fig. 3.2 are of 50Ω line widths. The model in Fig. 3.2 is simulated to demonstrate the achieved simultaneous frequency and
impedance matching agility. The capacitors are implemented with a bank of 4 variable capacitors in parallel with one another. A worst case Q of 200 at 400MHz is chosen setting $R_S = 0.34\Omega$ resulting in a capacitive tuning range of 2.8pF - 6pF from each realizable varactor diode and a tuning range of 11.2pF - 24pF for each bank. Fig. 3.3 illustrates the achieved impedance and frequency matching agility of this configuration, where load impedances are matched to a VSWR of 2:1 or better in the band of interest. State 1 is capable of covering 200 to 300MHz (1.45:1) and State 2 is capable of covering 275 to 400MHz (1.5:1). The design in Fig. 3.2 will be used to analyze circuit loss and tuning topologies in section 3.2 and section 3.3.

![Figure 3.3: Impedance and frequency matching agility of varactor loaded triple stub tuner, high band (top), and low band (bottom) performance.](image)

3.2 Losses in Realizable Tunable Circuits

So far discussion of tunable circuits has not addressed circuit losses. The goal is to understand how losses in the tuning circuit can affect performance, and to compare an ideal design to a realizable circuit. First, an understanding of matching circuit loss mechanisms will be discussed.
The most obvious source of loss in a reconfigurable matching network are those found in lumped circuit components and connecting transmissions lines (i.e. ohmic losses, conduction losses, etc). For lumped components, the series or parallel resistance, $R_S$, is typically characterized by the manufacturer and varies with frequency. However, for tunable components such as variable capacitors, $R_S$ changes with frequency and bias voltage (different capacitance values). It is therefore difficult to track the losses associated with each component at each tuning state and frequency.

Previous research addresses designing matching networks that account for lumped element losses. It has been shown that optimal high efficiency matching networks can be synthesized in the presence of realistic component losses [29,30]. However, these works refer to static rather than reconfigurable matching networks. While these models are useful in accounting for component losses, they do not necessarily compensate for troublesome reflections in multi-stage networks.

Circuit losses are manifested in three ways:

1. Lumped element losses due to $R_S$, ohmic and conduction losses from the PCB (loss as $R_{\text{circuit}} \rightarrow R_L$).

2. Loss due to inter-stage impedance mismatches and reflections.

3. Circuit resonances and radiation losses.

To illustrate the effects of these different (but related) loss mechanisms, two simple impedance matching circuits are analyzed (see Fig. 3.4 and Fig. 3.5).
Figure 3.4: Single and multi-stage networks used to match the load having \( R_L = 13\Omega \), \( X_L = 23\Omega \), \( Z_o = 50\Omega \), and \( Z_S = 50\Omega \) at the source, (a) single-stage matching with \( \ell = 38.6^\circ \), \( C = 38.6\text{pF} \), \( R_S = 2\Omega \), and (b) triple-stage matching with \( \ell_1 = 10^\circ \), \( C_1 = 10\text{pF} \), \( R_{1s} = 0.5\Omega \), \( \ell_2 = 39^\circ \), \( L = 25\text{nH} \), \( R_{2s} = 1\Omega \), \( \ell = 16^\circ \), \( C_2 = 4.42\text{pF} \), \( R_{3s} = 0.5\Omega \), \( Z_1 = 16 - j23\Omega \), \( Z_2 = 17 + j65\Omega \), \( Z_3 \approx 50\Omega \).

![Diagram](image)

Figure 3.5: Single stage and multi stage \(|S_{21}|\) response.

Both circuits in Fig. 3.4 were implemented using HFSS v12 full wave simulator with lumped elements on 50Ω microstrip lines on a Rogers TMM 10i substrate \((\varepsilon_r = 9.56 \text{ with } \tan\delta = 0.0016, \text{ and thickness of } 50\text{mil})\). Also, each of the lumped components in Fig. 3.4(a) and (b) are modeled with a series resistance (not depicted). For
Fig. 3.4(a) and (b), the total series resistance is 2Ω allowing for a fair loss comparison. Both the single and multi-stage networks are intended to match a load impedance $Z_L = 13 + j23\Omega$ to the 50Ω source. Accounting for only lumped element losses, $|S_{21}|$ can be calculated from the efficiency equation:

$$|S_{21}| = 10\log_{10}[\eta] = 10\log_{10}\frac{|Z_L|}{|Z_L| + R_{STotal}}$$

(3.4)

Here $\eta$ denotes efficiency and $R_{STotal}$ represents the total resistance for all components in the circuit. Solving (3.4) with $R_{STotal} = 2\Omega$, yields $|S_{21}| \approx -0.3dB$. This loss is observed in the single-stage matching network and therefore attributes the loss to the lumped element, $R_{STotal}$. In this case, the finite copper conductivity and tan$\delta$ from the PCB had negligible contribution to loss. From equation (3.4), it is concluded that the circuit efficiency will decrease as $R_{STotal}$ approaches the antenna’s radiation resistance. This further indicates the difficulty in realizing an efficient matching network for electrically small antennas with low radiation resistance. Even a small amount of loss can drastically decrease efficiency for low radiation resistances.

While the matching circuit in Fig. 3.4(b) has the same system loss as in Fig. 3.4(a), $|S_{21}|$ suffers 0.6dB more loss than that calculated by equation (3.4). Indeed, the additional losses are due to inter-stage mismatches. It is seen that looking into the circuit at $Z_1$, the impedance is $16 - j23\Omega$, and at $Z_2$ the impedance is $17 + j65\Omega$. It is not until $Z_3$ that the circuit is matched to $50 + j0\Omega$. As a result of these inter-stage mismatches, power reflected within the circuit leads to increased loss. This is depicted in the simplified circuit in Fig. 3.6 where interstage reflections / standing waves exist in Fig. 3.6(b) between Stage 1 and Stage 2, as well as between Stage 2 and Stage 3. In general, the more mismatched the antenna is, the larger the reflections through the
different stages of the network, resulting in higher losses. This can be problematic in matching networks involving multiple stages with lossy components.

Another severe loss mechanism is due to circuit resonances that cause transmission line radiation. Such resonances can occur in multi-stage matching networks where long transmission lines may exist. Radiation may also occur in the most extreme mismatch conditions where large reactive load impedances may appear.

Due to the reasons outlined above, it can be quite difficult to determine the amount of loss a multi-stage matching network will have without knowing the load impedance. To illustrate this, a multi-stage matching network capable of matching all load impedances on a VSWR 20 circle to a VSWR 2 or better are individually matched and plotted in Fig. 3.7 for a 50Ω source. The color of each data point represents the $|S_{21}|$ associated with the load impedance after tuning. While the return loss is
identical for each load impedance, the simulated $|S_{21}|$ after tuning changes drastically from 0.1 to 1.7 dB of loss. It is noted that the load impedances on the left side of the Smith Chart (closer to the short circuit) have less loss, where the impedances on the right side of the chart have more loss. On the left side of the Smith Chart on the VSWR 20 circle, the load resistance is low (minimum of 2.5Ω), and the load reactance is also low. Where the VSWR 20 circle intersects the $\pm j1$ reactance curves, the impedance is still relatively low, $5\pm j50\Omega$, despite having twice the load resistance, a different mismatch condition is created where the circuit realizes 1.2 to 1.6 dB of loss as opposed to 0.4 dB of loss for the 2.5Ω load.

Figure 3.7: $|S_{21}|$ results after matching the load impedance for a multi-stage tuning circuit. Loss varies up to 1.7 dB for a constant VSWR=20 circle.

If only the the absolute load impedance and lumped circuit loss is considered, it would be expected that the most loss would be observed for the lowest load impedances (left side of the Smith Chart) assuming the matching circuit was able
to match all impedances on the VSWR 20 circle to a VSWR 2:1. However, with an understanding of how interstage mismatches and reflections create different loss conditions, the results in Fig. 3.7 make sense.

With this understanding of loss mechanisms, the triple stub tuner presented in Fig. 3.2 can be analyzed with and without losses. It was seen in Fig. 3.3 that full Smith Chart coverage was achievable between the two matching circuits from 200 to 400MHz where each impedance is matched to a VSWR 2:1 or better. The circuit is re-analyzed at 250MHz considering the transducer gain after tuning in Fig. 3.8, transducer gain will be summarized in detail in the following section. Each point on the Smith Chart is matched using a genetic algorithm (GA) in Advanced Design System (ADS). Fig. 3.8(a) shows the transducer gain after tuning when no losses are present in the circuit. Since this circuit can match all impedances to a VSWR 2:1 or better, the worst case transducer gain would be -0.51dB and entirely be a result of the impedance mismatch due to a VSWR of 2. Hence all the impedances after matching in Fig. 3.8(a) result in \( \approx -0.5 \)dB of loss or less.

The transducer gain is plotted again after matching in Fig. 3.8(b) for the same circuit accounting for losses. The variable capacitors in Fig. 3.2 are replaced with commercial varactor diodes. Included in the simulation are the small signal resistance in the varactors where the component SPICE model is used. Also present are the conduction and ohmic losses from the PCB for a microstrip implementation. Since simulations are done using ADS, and are not full wave, any microstrip line coupling or radiation is not accounted for. As such, transducer gain is a result of mismatch loss (less than half a dB), ohmic / conduction losses, and \( R_s \) from the varactors / bias network, which are amplified due to standing waves in the circuit. Up to 6 dB and
worse loss is now observed (note the change in scale between Fig. 3.8 (a) and (b)). This demonstrates the difficulty in designing wide band agile matching networks. Commercially available components still require improvement for such aggressive frequency and impedance agility. While the lossless design had good performance, the design appears less appealing once losses are considered. Accurately modeling a particular matching network with losses over a broad frequency and impedance spectrum is cumbersome and time consuming as full wave simulations should be used.

![Graph showing transducer gain comparison between lossless and lossy scenarios.](image)

Figure 3.8: Simulated transducer gain for Fig. 3.2 low band circuit at 250MHz, (a) no losses (b) includes varactor and PCB losses.

### 3.3 Improved Tuning Topologies

Traditional closed loop tuning topologies such as those in [5,6] are illustrated in Fig. 3.10(b). A micro-controller adjusts the tuning circuit to minimize reflected power (B) and maximize (A) monitored with a directional coupler and RF detectors. This approach can effectively minimize the reflection coefficient. Though, it does
not necessarily maximize the power delivered to the antenna in practical multi-stage matching networks with ohmic losses as the following analysis will show.

An effective impedance matching network should maximize the transducer power gain (TPG). Transducer power gain is simply the ratio of power delivered to the load and the available power from the source. To better understand transducer power gain, the circuit model in Fig. 3.9 from [10] is studied.

![Figure 3.9: Two-port network with general source and load impedances with associated reflection coefficients and voltages.](image)

The two port network $[S]$ is connected to source and load impedances $Z_S$ and $Z_L$ respectively. From [10], the reflection coefficient seen looking toward the load is:

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (3.5)$$

Where $Z_0$ is the characteristic impedance for the two port network and the reflection coefficient seen looking toward the source is:

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} \quad (3.6)$$

Typically the input impedance of the network will be mismatched with reflection coefficient, $\Gamma_{in}$, which can be determined from the analysis of Fig. 3.9. Starting with
the definition of $\Gamma_L$ in terms of voltages:

$$V_2^+ = \Gamma_L V_2^-$$  \hspace{1cm} (3.7)

$$V_1^- = S_{11} V_1^+ + S_{12} V_2^+ = S_{11} V_1^+ + S_{12} \Gamma_L V_2^-$$  \hspace{1cm} (3.8)

$$V_2^- = S_{21} V_1^+ + S_{22} V_2^+ = S_{21} V_1^+ + S_{22} \Gamma_L V_2^-$$  \hspace{1cm} (3.9)

From (3.8) and (3.9):

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12} S_{21}}{1 - S_{22} \Gamma_L} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$  \hspace{1cm} (3.10)

$Z_{in}$ is the impedance seen looking into port 1 of the network. Likewise the reflection coefficient seen looking into port 2 when port 1 is terminated in $Z_S$ is:

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12} S_{21} \Gamma_S}{1 - S_{11} \Gamma_S}$$  \hspace{1cm} (3.11)

With an understanding of $\Gamma_L$, $\Gamma_S$, $\Gamma_{in}$, and $\Gamma_{out}$, the transducer power gain is (from [10,31]):

$$G_t = \frac{P_L}{P_{avs}} = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)(1 - |\Gamma_S|^2)}{|1 - S_{22} \Gamma_L|^2 |1 - \Gamma_{in} \Gamma_S|^2}$$  \hspace{1cm} (3.12)

However, if the system has a matched source, i.e. $Z_S = Z_0$ which often is the case for antenna impedance matching systems where $Z_S = Z_0 = 50\Omega$, $\Gamma_S = 0$ then the transducer power gain simplifies to:

$$G_t(\text{MatchedSource}) = \frac{P_L}{P_{avs}} = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{|1 - S_{22} \Gamma_L|^2}$$  \hspace{1cm} (3.13)

In (3.12), $P_L$ is the power delivered to the load (antenna), $P_{avs}$ is the power available from the source, and $\Gamma_L$, $\Gamma_{in}$, and $\Gamma_S$ are as previously described above and depicted in Fig. 3.9 and Fig. 3.10(a).

There are a number of different adaptive antenna tuning approaches in literature. They can typically be categorized by the way in which adaptive tuning is accomplished.
or by way of sensing (measuring load impedance, mismatch, power gain, etc...) and summarized by [32]:

1. Adaptive tuning that maximizes or minimizes a particular parameter of interest. Typically the VSWR or $S_{11}$ is minimized. These are mainly gradient descent like methods [6,33,34] or trial and error methods [35,36]. The approach used in this research is similar to this method. Directional couplers placed in different locations of the system sample reflected and or transmitted power and tuning goals are implemented minimizing and or maximizing different parameters. The goal of this work is to maximize transducer power gain (TPG). The disadvantage of this method is tuning speed, any properly implemented knowledge based tuning method will outperform a trial and error tuning method.

2. Adaptive tuning that use analytic computation assuming direct measurement of impedance [37]. This results in impedance measurements at a range of frequencies requiring a more computationally expensive control loop. The disadvantage of exact computation of the input impedance of the antenna is that it requires voltage and current measurements placing high demands on detection circuitry.

3. Adaptive tuning using phase difference to tune a reactive impedance to zero [38–40]. These methods require RF phase detectors [38,41] or Foster-Seeley discriminators [39,40]. In this case an optimum solution is a zero-phase difference signal. However the real portion of the impedance is not necessarily matched. If an electrically small antenna is used, it is already known the real portion is typically very low hence the matching network maintains a low real impedance.
Four directional coupler based tuning topologies are illustrated in Fig. 3.10 with associated tuning goals and brief descriptions. As mentioned previously, the topology used by the designer needs to ensure TPG is maximized. Each topology of Fig. 3.10 attempts to maximize TPG utilizing a different number of components, and monitoring different parameters. Items to consider when choosing a topology include:

1. A configuration that properly senses TPG so that it can be maximized by the micro-controller tuning algorithm.

2. Complexity of the topology, i.e. number of components. This can increase cost.

3. Insertion loss due to the directional couplers. With 2 directional couplers more insertion loss will be incurred to the system than if 1 is used. Of course this may become a balance between accurately sensing TPG and component count.

4. One limiting factor of each topology is operating bandwidth. This is a limitation inherent to the selected directional coupler and detection diode combination. This can be troublesome for UWB applications.

The following discussion will describe the operating principal of each presented topology, its tuning goal, and design trade-off’s.

### 3.3.1 Load Pull Configuration Fig. 3.10(a)

Fig. 3.10(a) is essentially the same topology used in traditional load pull measurements as seen in [42–45]. Load pull measurement systems use this topology to accurately measure power gain. As described in [45], the power gain $G_P$ is calculated from:

$$ G_P = \frac{P_{out}}{P_{in,del}} = \frac{(|C| - |D|)}{(|A| - |B|)} = \frac{|C|(1 - |\Gamma_L|^2)}{|A|(1 - |\Gamma_{in}|^2)} \quad (3.14) $$

This can be troublesome for UWB applications.
Maximize Minimize

A & C-D B

Direct method to measure power gain in load-pull characterization of devices. Minimizes reflections from the impedance tuner and antenna while maximizing power out from the tuner.

• High component count
• Incurs more insertion loss with 2 couplers
• Most accurate measure of TG, thus best to ensure TG is maximized

Maximize Minimize

A B

Common $\Gamma_{in}$ minimizing topology used in literature.

• Effectively minimizes $\Gamma_{in}$
• Can potentially maximize circuit losses
• May be suitable for very low loss circuits

Maximize Minimize

C B

Minimizes $\Gamma_{in}$ and maximizes power out of the tuning circuit.

• More accurate than topology (b)
• Not as accurate as topology (a),(d)
• Incurs more insertion loss with 2 couplers compared to (b) or (d)

Maximize Minimize

C D

Maximizes power out of the tuning circuit and minimizes $\Gamma_{L}$.

• Mimics topology (a) performance with half the insertion loss (1 direction coupler)
• Relevant for lossy and loss-less circuits

Figure 3.10: Adaptive tuning topologies.
Where $P_{\text{out}}$ is the output power delivered by the tuner, and $P_{\text{in,del}}$ is the input power delivered to the tuning circuit from the source. In (3.16), “A” represents both the power from the source and the power reflected from the source. “B” denotes the power reflected from the impedance tuner (between Coupler 1 and the tuner), as well as power that was transmitted through the tuner and reflected back either from Coupler 2 and or the antenna through the coupler in the reverse direction. “C” denotes power transferred through the tuner and power reflected from the tuner as observed from a wave traveling from the right to the left (backward direction) of the figure. Finally, “D” denotes power reflected from the antenna. This discussions assumes the antenna is set as a transmitting device. From (3.16) it is observed that the topology of Fig. 3.10(a) should maximize (A) and (C-D), while minimizing (B). Reflections from the tuner and antenna are minimized while maximizing power out from the tuner. As defined by [10], power gain is the ratio of power dissipated in the load to the power delivered to the input of the two port network. I.e. power gain is not dependent on $Z_S$, while transducer power gain is dependent on $Z_S$. Since it is assumed that our antenna is ultimately connected to a $Z_S = 50 \Omega$ source, (a matched source) $\Gamma_S$ reflections are eliminated. In this case, maximizing power gain would be the same as maximizing transducer power gain. While this topology is a direct method to measure TPG and quite accurate, it has a high component count with 2 directional couplers and 4 detection diodes. With 2 directional couplers, more insertion loss is introduced to the system which is undesired.
3.3.2 Minimizing $\Gamma_{in}$ Fig. 3.10(b)

As observed from equation (3.12), maximizing $G_t$ does not only depend on $\Gamma_{in}$. Thus minimizing $\Gamma_{in}$ as in Fig. 3.10(b) will not necessarily guarantee maximum $G_t$. Traditional tuning topologies assume that minimizing $|S_{11}|$ as in equation (3.15) will maximize efficiency. However, equation (3.15) is only true for a lossless system. That is, (3.15) is not suitable for agile matching systems with multiple stages and lossy components.

$$|S_{21}| = \sqrt{1 - |S_{11}|^2}$$  \hspace{1cm} (3.15)

If the tuning circuit is relatively low loss, Fig. 3.10(b) would be suitable to use. If no losses exist, a matched source is used, and $\Gamma_{in}$ is eliminated, then likewise $\Gamma_L$ and $\Gamma_{out}$ would be eliminated for a passive bi-directional device such as a typical tuning circuit, and TPG would be maximized. This topology does use half the components as Fig. 3.10(a), thus insertion loss due to directional couplers would be halved, and cost would be lower.

3.3.3 Minimize $\Gamma_{in}$ and Maximize Power Out Fig. 3.10(c)

Fig. 3.10(c) aims to minimize the reflected power (B) from the tuning circuit while maximizing the power transferred through the tuning circuit (C). If power out of the tuning circuit is maximized, this energy can still reach the antenna, where a mismatch can exist, and power can reflect back ($\Gamma_L$). With this topology it is assumed that reflections from $\Gamma_L$ will travel back through the tuning circuit and be present in monitored power (B). As such, the power delivered to the antenna might be maximized while minimizing reflections. As such Fig. 3.10(a) performance might be achievable with less detection diodes. It will be shown in section 3.4 that this is
not always the case. However, this topology is more accurate than Fig. 3.10(b) at the expense of an extra directional coupler adding to insertion loss.

3.3.4 $|S_{21}|$ Maximizing Fig. 3.10(d)

Fig. 3.10(d) neglects any sampling of reflected energy from the tuning circuit, but rather intends to maximize the power through the tuning circuit (C) and minimize the power reflected from the antenna or load (D). According to equation (3.13), TPG for a matched source is dependent on reflections from the antenna/load, and $|S_{21}|$ of the matching network. Thus if power out of the circuit is minimized, and reflections from the antenna are indeed minimized, TPG will be maximum. Section 3.4 will show that this topology provides similar performance as Fig. 3.10(a). This is accomplished with half the number of components, and is relevant for both lossy and loss-less circuits. As a result, since this topology uses as many components as Fig. 3.10(b) with equal (for a lossless circuit) or better (for a lossy circuit) performance, it is a suitable alternative to any $|S_{11}|$ minimizing topology. In reality a perfectly matched source is difficult to realize, as such this topology’s performance would not be as good as Fig. 3.10(a) as it does not sample all required parameters to maximize TPG. Regardless, the cost benefit of half the number of components (less insertion loss) may surpass the slightly better performance offered by Fig. 3.10(a).

3.4 Tuning Topology Analysis

Three tuning circuits are simulated to analyze the effectiveness of each tuning topology of Fig. 3.10. Triple, double, and single stub varactor loaded impedance tuners depicted in Fig. 3.11 are used for analysis. Each circuit was built and simulated in Advanced Design System (ADS) 2011. The layouts of Fig. 3.11 are implemented
with a 50mil Rogers TMM10 substrate ($\varepsilon_r = 9.56$), and trace widths of 50 Ω. The variable capacitors are realized with Skyworks SMV1801-079LF varactor diodes, the spice model [46] of the varactor was used in simulation. Fig. 3.11 omits illustration of the DC bias circuitry including DC and RF blocks that were included in the simulation model and illustrated in Fig. 3.17.

![Diagram](image)

Figure 3.11: Microstrip impedance tuners to characterize Fig. 3.10 topologies. $C_b$ represents a varactor tuning bank with a tuning range of 11.2-70pF.

Transducer gain heat plots are created to depict the simulated performance for each tuner at a specific frequency. Such plots may be foreign to the reader, the method behind producing the transducer gain Smith Chart characterization depicting performance is described in Fig. 3.12. First a topology is chosen for characterization, such as one from Fig. 3.10 (Step 1). Next, a matching network is implemented in ADS with the required directional coupler configuration, such as the triple stub tuner from Fig. 3.11(a) (Step 2). Then a particular load impedance at a given frequency
is chosen to match i.e. $50 + 50j\Omega$, and set as $Z_L$ for the two port network (Step 3). The built in ADS genetic algorithm is used to optimize each of the $C_b$ tuning banks based on the tuning goal for the directional coupler configuration, i.e. maximize $C$ and minimize $B$ (Step 4). After optimization of each $C_b$ is completed, the transducer gain is recorded in software. This TPG value is represented by the color of the data point on the Smith Chart at the location of the tuned load impedance. In this case, after tuning, TPG was $\approx -1\text{dB}$ (Step 5). This process is then repeated for each load impedance across the Smith Chart at the chosen frequency (Step 6).

Each circuit in Fig. 3.11 has the same design parameters with the specified number of stubs and used for analysis. Simulation results are not full wave, but do include the full spice model of the varactor as well as ohmic, conduction, and bias network losses.

Fig. 3.13(a), Fig. 3.14(a), and Fig. 3.15(a) represent the near ideal case for tuning the varactor banks based on direct calculation of TPG in software (all parameters in equation (3.12) are known). It is “near ideal” because the circuit optimizer may not have chosen the best solution due to its starting condition, an inherent limitation of tuning algorithms. However, the genetic algorithm is generally able to find global maximum / minimum. For the three tuners, we see that the load pull topology, Fig. 3.13(b), yields very similar results to the ideal case as expected.

However, drastic deterioration of circuit efficiency is noticed in Fig. 3.13(c) where only $|S_{11}|$ of the circuit is minimized. Several points on the Smith Chart are matched to circuit loss. The topology used in Fig. 3.13(d) is certainly an improvement over Fig. 3.13(c), though it does not yield ideal performance. Fig. 3.13(e) effectively maximized power transferred through the tuning circuit and minimized reflections
Step 1:
Choose a tuning topology

Step 2:
Implement designed matching network

Step 3:
Choose an impedance at a particular impedance to match, set it as $Z_L$

Step 4:
Optimize tuning banks based on the tuning goal

Step 5:
After tuning, save and plot the Transducer Gain

Step 6:
Repeat for all load impedances on the Smith Chart for that frequency

Figure 3.12: Transducer Gain Smith Chart Characterization.
from the load. This approach is closer to the ideal case and similar to Fig. 3.13(b). This is an attractive topology to use as it provides favorable performance with less components. Analysis for the double stub tuner is similar to that of the triple stub. The circuit has less loss due to fewer stages however, the $|S_{11}|$ minimizing topology still significantly degrades performance.

The single stub tuner simulation results are shown in Fig. 3.15 for each of the tuning topologies. As expected the single stub tuner has less impedance matching agility than the triple stub tuner. The transducer gain is near identical for each topology. There is one location in Fig. 3.15(c) where the $|S_{11}|$ minimizing topology again matches to circuit loss as opposed to the ideal case in Fig. 3.15(a). Overall, any
Figure 3.14: Double stub (from Fig. 3.11(b)) simulation analysis for topologies at 300MHz from: (a) ideal $|S_{21}|$ optimization, (b) Fig. 3.10(a), (c) Fig. 3.10(b), (d) Fig. 3.10(c), (e) Fig. 3.10(d).

of these tuning topologies for the single stub tuner would suffice, which is in contrast to the triple stub tuner.

It is evident for the triple stub tuner that an $|S_{11}|$ only minimizing topology can maximize circuit losses which are dominated by the small signal resistance $R_s$ of the varactor diodes. The loss problem is much more of an issue for the triple stub tuner due to the multi-stage matching network. Because of multiple stages, there are increased reflections and standing waves in the tuning circuit which repeatedly dissipate in $R_s$ as heat, which does not occur for the single stub tuner. While this ADS simulation was not full wave, circuit radiation is also a concern and would contribute to increased losses.
Figure 3.15: Single stub (from Fig. 3.11(c)) simulation analysis for topologies at 250MHz from: (a) ideal $|S_{21}|$ optimization, (b) Fig. 3.10(a), (c) Fig. 3.10(b), (d) Fig. 3.10(c), (e) Fig. 3.10(d).

Lastly, in application, each tuning topology should be studied across the operating frequency band. Based on previous discussion, it is evident that Fig. 3.10(a) will always be the best topology to use. However, as component count becomes a concern it is viable that Fig. 3.10(d), or Fig. 3.10(b) (for less aggressive low loss systems) are options. Regardless, during design it would be best to study the chosen topology’s performance vs frequency. Fig. 3.10 summarizes the performance of each topology and should be consulted during the design process.

3.4.1 Topology Performance vs Frequency

In this section the topologies of Fig. 3.10(a), Fig. 3.10(b), and Fig. 3.10(d) are studied across a 2:1 bandwidth for the triple stub tuner in Fig. 3.11 and compared
to the ideal case. Transducer gain results are shown for these 4 cases in Fig. 3.16. Due to the analysis results at a single frequency from section 3.4, one would expect each topology to perform similarly across an operating bandwidth. Fig. 3.16(a) shows the ideal performance where TPG is directly calculated and is the optimization goal. Fig. 3.16(b) represents the load-pull topology, and Fig. 3.16(d) the $|S_{21}|$ maximizing topology where only one coupler on the output of the tuner is used. As expected, and consistent with earlier results, these three topologies perform quite similar. The $|S_{11}|$ optimized topology is shown in Fig. 3.16(c) and as expected, has worse performance at every frequency compared to the other topologies. These results are consistent with section 3.4 and verifies that each topology does indeed perform consistently across an operating band.
3.4.2 Prototype Measurements

The agile impedance tuner of Fig. 3.11(a) was fabricated to validate the tuning performance of Fig. 3.13(c) and Fig. 3.13(e) and is depicted in Fig. 3.17. The prototype was comprised of the same short circuited stub length, inter-stub spacing, and $C_b$ varactor tuning bank, implemented using Skyworks SMV1801-079LF varactor diodes.
as in Fig. 3.11. It was fabricated on a 50 mil Rogers TMM10 substrate ($\epsilon_r = 9.56$) and tuned to 340MHz.

Figure 3.17: Fabricated triple stub microstrip tuner from Fig. 3.11(a) design, C1-C3 values are in Fig. 3.18, minimum varactor Q = 230 at 340MHz.

The tuning network in Fig. 3.17 was tested for Fig. 3.10(b) (minimizing $\Gamma_{in}$) and Fig. 3.10(d) (maximizing $|S_{21}|$) topologies and optimized by a microcontroller with a brute-force tuning algorithm. Mini-Circuits ZFBDC20-62HP+ bi-directional couplers were used to implement the tuning topologies. The topology of Fig. 3.10(a) (load pull configuration) was also tested and yielded identical results to that of Fig. 3.10(c) and thus the results are omitted. For the measurements, the antenna in Fig. 3.10 was replaced with a load impedance $Z_L = 8.5 - 20.5j\Omega$ at 340MHz. Each topology is tuned based on the associated goals in Fig. 3.10 and the transducer gain is measured in a load-pull configuration (more details in section 3.6) and compared to the un-tuned mismatch loss. The un-tuned mismatch loss is representative of power that is not received by the load due to impedance mismatch, not dissipated losses.
Figure 3.18: Prototype measurements (a) $|S_{11}|$ and (b) normalized $|S_{21}|$ with and without tuning at 340MHz, where C is maximized and D minimized: $C_1 = 11.5$ pF, $C_2 = 70.8$ pF, $C_3 = 11.7$ pF, and where B is minimized and A is maximized: $C_1 = 14.8$ pF, $C_2 = 20.2$ pF, $C_3 = 19.7$ pF at 340MHz.
A comparison of $|S_{11}|$ data for $Z_L$ under different tuning conditions is shown in Fig. 3.18(a). As expected, without tuning, the circuit is severely mismatched with a $|S_{11}|$ of -2.4 dB. It is also seen that the traditional method of minimizing $|S_{11}|$ (Fig. 3.10(b) topology) yields a slightly better match than the proposed Fig. 3.10(d) topology maximizing $|S_{21}|$ of the system. However, while both tuning methods provide acceptable $|S_{11}| (< -10\text{dB})$ results, the power delivered to $Z_L$ differs drastically. Where the topology of Fig. 3.10(d) improves the untuned case by 5dB while the topology of Fig. 3.10(b) degrades performance by 8dB.

The prototype circuit in Fig. 3.17 was tested using a load-pull measurement. The low-band tuning circuit ($\ell_{1stub} = 60^\circ$ and $\ell_{1series} = 45^\circ$ at 250MHz) from Fig. 3.2 is fabricated to demonstrate Fig. 3.10(b) (minimizing $\Gamma_{in}$) topology compared to Fig. 3.10(c) in an antenna tuning configuration. This prototype is also fabricated on a 50mil Rogers TMM10 substrate ($\epsilon = 9.56$) illustrated in Fig. 3.19. The circuit was fabricated on two separate boards with vias connecting the DC bias circuit to the varactors on the RF circuit. A single Skyworks SMV1801-079LF varactor diode was used to produce variable capacitance at the end of each tuning stub with a tuning range of 2.8 to 6 pF where $Q = 200$ at 6pF at 400MHz. Note that this prototype lacks the phase shifter and full capacitive tuning range previously described for this design and will thus have limited frequency and impedance agility.

Two topologies are implemented and tested, the first is that seen in Fig. 3.10(b) where the directional coupler is placed between the network analyzer and the tuning circuit and $|S_{11}|$ is minimized. The second is that of Fig. 3.10(c) where an additional directional coupler is placed between the tuning circuit and radiating antenna. Here $|S_{21}|$ through the circuit is maximized while $|S_{11}|$ is minimized. The measurement
setup is illustrated in the inset of Fig. 3.20(a), where a resonant 1GHz dipole is tuned and used for transmit. A receiving antenna on port 2 of the analyzer is used to measure radiated power data. Measured results are presented in Fig. 3.20 and Fig. 3.21. Fig. 3.20(a) compares $|S_{11}|$ of the 1 GHz resonant dipole for different tuning conditions. As expected without tuning the antenna is severely mismatched with a $|S_{11}|$ of -1.2 dB (VSWR = 14:1). When comparing $|S_{11}|$ of the two tuning methods, we see that the traditional method of minimizing $|S_{11}|$ yields a slightly better match than minimizing $|S_{11}|$ and maximizing $|S_{21}|$, however both effectively accomplish tuning this severely mismatched antenna at 292MHz. It is observed that while both methods of tuning result in an acceptable $|S_{11}|$, the radiated power yields drastically different results. When minimizing $|S_{11}|$ and maximizing $|S_{21}|$, we see that there is a 12.8 dB improvement in the radiated power when compared to the no tuning case. However, where $|S_{11}|$ is minimized, the radiated power is decreased 4 dB relative to the no tuning case. While this extreme difference in radiated power does
Figure 3.20: Prototype measurements from Fig. 3.19, (a) $S_{11}$ and (b) normalized transducer gain with and without tuning at 292MHz.
Figure 3.21: Measured normalized radiated power frequency tuning range from Fig. 3.19 based on Fig. 3.10(c) topology.

not occur for all tuning cases, it does illustrate the potential of an $|S_{11}|$ minimized circuit to maximize loss thus minimizing radiated power. This example illustrates the serious potential consequences of only minimizing $|S_{11}|$ in a realistic matching circuit. Fig. 3.21 demonstrates the frequency range in which this circuit can tune this resonant length 1 GHz antenna. A 20% tuning range from 268 to 327MHz is realized before improvements begin to degrade.

In summary, when only minimizing $|S_{11}|$, troublesome $R_s$ from the varactor diodes can cause a loss in power due to multiple reflections / standing waves in the multi-stage network. Circuit radiation from resonances also contribute to system loss. This validates the Smith Chart simulation comparison of Fig. 3.13. The antenna designer should always analyze their system and choose the tuning topology best suited for the application.
3.5 Impedance Tuning Algorithms

For adaptive tuning systems, an iterative tuning method that attempts to converge on a solution goal is used. An iterative method is avoided when either: A direct measurement of input impedance can be obtained and variable component values set according to prior tuner characterization. Or when a look-up table can be used to set tuning states assuming the antenna is characterized vs frequency ahead of time. These approaches are also the fastest methods of tuning. Of course the latter of the two assumes a static environment and thus this solution is not useful when the antenna can be subjected to dynamic coupling effects. Furthermore, as described in [47,48], the tuning process can cause signal phase and amplitude modulation as matching components are varied. Thus transmitted data can be corrupt if tuning occurs during vital transmission times. Tuning speed can become of concern, it would be best if tuning was limited to idle information transmit periods to reduce the possibility of data corruption. Another concern with adaptive iterative tuning algorithms is how robust they are. There are many solution spaces that have several local minima or maxima. Algorithms may find tuning solutions that are at a local min or max as opposed to a global min or max. As such, the optimum solution may not be found. A suitable algorithm must be chosen based on the solution space of the particular tuner and feed-back system in use. There are a myriad of different optimization algorithms that are suited for impedance tuning. A few noted options are reviewed below.
1. Brute-Force Tuning: This is an exhaustive search method as opposed to an adaptive optimization process. Brute force tuning involves testing every combination of tuning states. Once a tuning state is set, the feedback loop provides data based on the tuning goal (i.e. VSWR, power gain, etc), this quantity is then compared to the previous best solution. The tuning state corresponding to the best feedback solution is saved in memory. After every state has exhaustively been tested, the best solution parameters are refreshed in the system. This approach will always guarantee a global max or min solution at the expense of time. If there is a large search space this may not be feasible, however for a relatively small number of combinations, this may be a suitable approach. For the author’s work, unless otherwise stated, a brute force approach was used for any tuning solutions in measurement.

2. Hooke and Jeeves’s Algorithm [6,49]: An arbitrary starting condition is chosen as a base point (i.e. combination of tuning parameters) and moves along each of the coordinate directions in the search space. Periphery points relative to the starting condition are explored and tested to find the direction that leads to a better solution. A new best solution is found and sets the basis as a new starting position to repeat the search process. The algorithm is repeated until a threshold is satisfied or a certain number of iterations are completed. This algorithm will not necessarily find a global optimum solution. An illustration of how the algorithm iterates is in Fig. 3.22 modified from [50].

3. Simplex Method [6,49]: This method geometrically discretizes the solution space into triangles in 2-dimensions or tetrahedron in 3-dimensions. The geometric
Figure 3.22: Example of the Hooke and Jeeves optimization algorithm. “I” represents number of iterations.

The figure is composed of n+1 points in an n-dimensional space. With a random starting condition the simplex method moves from one feasible extreme point to another, each time improving the performance criteria. This is repeated across the search space until an optimum is found. The principle of the method is to enumerate only a small fraction of a very large space before iteratively reaching an optimal solution. An illustration of how the algorithm iterates is in Fig. 3.23 modified from [51].

4. Genetic Algorithms [36,47,52,53]: Often referred to as an “evolutionary” algorithm, resembles the concept of natural selection and reproduction. Rules are set that ensure the survival of the best solutions in a large population. Random mutations are added and best solutions are mated with each other in
new iterative solution spaces. With continued operation, the genetic algorithm (GA) adapts to the system, changing tuning parameters, and achieves matching without requiring explicit rules. This algorithm is quite good at avoiding local solutions and finding global solutions. The scope of the author’s work uses GA’s in tuning simulations where ADS was used to characterize transducer gain.

5. Single-Step Algorithm: This method is noted for requiring little memory. An initial value is chosen, a short “step” in one neighboring direction of the search space is taken, tested, and compared to the previous value. If the new solution is better than the previous solution, another “step” is taken in that direction of the search space. If the solution is worse, a step in a different neighboring direction is explored. The algorithm repeats until no search direction results in a more optimum value, a threshold is reached, or a predetermined number of iterations is completed. This algorithm is susceptible to local solutions. The author’s
work will use this method in measurements in later sections. An example of the performance of the algorithm is in Fig. 3.24. A double stub tuner is simulated where at 250MHz the inter-stub spacing is 45°, short circuited stub lengths are 40°, each stub is loaded with a variable capacitor that varies between 0 to 15pF, and all transmission lines are 50Ω. The solution space represents the transducer gain when a load impedance $Z_L = 10 - 100j$Ω is tuned through the matching network for every permutation of capacitance for the two stubs. Fig. 3.24 shows four different initial values for the tunable capacitors. Three of the starting conditions yields a global maximum solution for the transducer gain. One of the initial values finds a local solution at capacitance values of 0pF for both stubs. This demonstrates the potential for this algorithm to find a local maximum for transducer gain. However, the other three starting conditions converge on a global solution after sampling a very small amount of the total solution space, hence a computational and time savings over a brute-force attempt.
Figure 3.24: Example of the Single-Step optimization algorithm. Solution space is for a double stub variable capacitor loaded tuning circuit optimizing transducer gain for $Z_L = 10 - j100\Omega$ at 250MHz.

### 3.6 VHF Impedance Synthesizer For Load Pull Measurements

Transducer gain (TPG) analysis was previously shown for different tuning topologies in simulation. While this is straight forward in software, fully characterizing Smith Chart TPG vs frequency in measurement is more cumbersome. Fig. 3.25 depicts the proper measurement setup to characterize an antenna impedance tuner (the device under test) for different topologies. While two directional couplers are present, any directional coupler topology previously discussed can be tested by only optimizing the antenna tuner based on the desired sampled voltages (a, b, c, or d). Of interest is how to create all load impedances on the Smith Chart in a measurement
setup. The solution is a robust Load Impedance Tuner which replaces the antenna for each of the previously discussed tuning topologies. The following outlines a novel load impedance tuner for TPG characterization in measurement.

Load pull measurements are the standard for characterizing output power, gain, efficiency, and stability [54,55]. Characterization of the device under test (DUT) in Fig. 3.25 (microwave transistors, amplifiers, or antenna impedance tuners (AITs)) is essential for design and verification. This work presents a robust load impedance tuner (Fig. 3.25) that synthesizes load impedances ($Z'_L$) presented to the DUT during characterization. While load pull measurements use the term “Load Impedance Tuner,” the device is better described as a “Variable Impedance Synthesizer” (VIS). Therefore, it will be referred to as such throughout this Letter. The application of this VIS is to assist the characterization of power gain for AITs at different $Z_L$ vs frequency. While a specific antenna might have one set of impedance values vs frequency in free space, its input impedance may change in a dynamic environment. As such, it is of interest to characterize the AIT across all impedances on the Smith Chart at a given frequency. The VIS makes such characterization possible.

Most AITs are excited with a 50Ω source impedance. Thus the “Source Impedance Tuner” inherent to load pull measurements is un-necessary. For the DUT in Fig. 3.25, the power gain $G_P$ can be calculated from [10,45]:

$$G_P = \frac{P_{out}}{P_{in,del}} = \frac{|c|^2 - |d|^2}{|a|^2 - |b|^2} = \frac{|c|^2(1 - |\Gamma_L|^2)}{|a|^2(1 - |\Gamma_{in}|^2)} \quad (3.16)$$

Here, $P_{out}$ is the output power delivered by the tuner, and $P_{in,del}$ is the input power delivered to the tuning circuit from the source.

As depicted, the voltages at a, b, c, and d are sampled by low-loss directional couplers and measured by a network analyzer.
Figure 3.25: Load-pull measurement configuration (omitting source pull tuner), suitable for characterizing power gain of the DUT for a 50Ω source.

Commercial solid state electronic VIS operate as low as 400MHz providing impedances inside the VSWR 10:1 circle with up to 5W of power handling in a 5.7x7.2x2in package including control circuitry [56]. Reconfiguration is typically achieved using varactors or PIN diodes [57]. To achieve broader impedance coverage and lower operating frequency, electromechanical VIS use slide screws, line stretchers, slug stubbed tuners, stepper motors, and other mechanical means to produce a broad range of load impedances. Commercial products such as [58] operate down to 250MHz, but are physically large (12x10x37in), producing loads inside a VSWR 25:1 circle, and handling as much as 250W of power.

Still, automatic VIS for large Smith Chart coverage below 250MHz are not available, hence the pursuit of this work. A triple stub varactor topology is employed to design a VIS operating from 125 to 300MHz to characterize the DUT using a 1.75x1.75in unit (≈ 0.03λ at 125MHz, excluding control circuitry). The VIS can achieve $Z_L$’s inside the VSWR 30:1 circle over the presented 2.4:1 bandwidth. A
novel approach of exploiting varactor diode losses and using different terminations (short circuit, open circuit, and 50Ω) leads to a robust and low-cost VIS. This VIS was tested at 0dBm power levels. However using back-to-back or anti-series varactor configurations can improve linearity and handle up to 2W of power [26, 59].

3.6.1 Triple Stub Varactor Loaded Tuner

A variable capacitor placed on a fixed length short or open circuited stub has been shown to provide excellent tunability [9]. To achieve broad Smith Chart coverage across a wide bandwidth, a short circuitted triple stub configuration is used as depicted in Fig. 3.26(a). When combined with a variable capacitor, these short circuitted stubs can realize both variable inductance and capacitance [60]. The spacing between stubs, and stub lengths are fixed at 21.2° and 18.3° at 275MHz respectively. A bank of four Skyworks SMV1801-079LF varactor diodes are used to vary the susceptance of each stub. Each individual varactor is able to produce capacitances from 2.65pF to 87.66pF using a reverse bias voltage range of 30 to 0V respectively. Each capacitor bank \( C_{bank} \) from Fig. 3.26(a) varies from 10.6pF to 350pF for the four parallel varactors.

While this may seem like a significantly wide tuning range, this particular varactor becomes quite lossy beyond 8pF, at 300MHz the small signal resistance \( R_S \) reaches 4.25Ω at 87 pF. As such, the impedance of the varactor has a significant real part (loss) between 8pF and 87pF. However, this is not an issue for the VIS as the goal is to synthesize a wide range of real and imaginary impedances for the DUT. Fig. 3.27 shows the spice model simulated varactor loss \( R_S \) vs frequency for different varactor capacitances (bias voltages). At 100MHz, the loss varies between 0.005 and 17Ω as
$C_{bank}$ is varied from 2.65 to 87pF respectively. Clearly, this provides a wide range of reactance and resistance values for the triple stub configuration.

![Diagram](image)

Figure 3.26: Short circuited triple stub varactor load impedance tuner, (a) schematic where $C_{bank}$ has a tuning range of 10.6 to 350 pF, and the two port device can have different terminations, (b) tuner implementation in microstrip.

It is acceptable for the load impedance created by the tuner to come as a result of resistance (loss) in the matching network. As summarized in [61], for multi-stage matching networks, inter-stage mismatches produce reflections / standing waves which repeatedly dissipate in the small signal resistance of the varactor as heat. In
some cases the varactors appear resistive rather than capacitive to the matching network. While this is undesirable for an AIT, as an impedance synthesizer, losses are helpful in creating more impedance values. To produce as many impedance values as possible, the VIS’s second port is terminated in a open circuit, short circuit, or 50Ω load (Fig. 3.25, Fig. 3.26(a)) as demonstrated in the following section.

### 3.6.2 Load Impedance Tuner Coverage vs Frequency

To evaluate the proposed tuner, the circuit in Fig. 3.26(a) was fabricated on a 50-mil thick Rogers TMM 10 substrate ($\epsilon_r = 9.56$) using 50Ω trace widths, as depicted in Fig. 3.26(b). RF blocks of 4.4 μH and DC blocks of 2.4 nF were also added for proper DC biasing. The resulting circuit is only 1.75x1.75in area ($\approx 0.03\lambda$ at 125MHz) and fabricated on standard PCB making for easy, low cost fabrication. An exhaustive
set of measurements were conducted as the bias voltage of \( C_{bank} \) was increased from 0 to 30V in 1V increments, and the input impedance was measured.

To measure every permutation of \( C_{bank} \) for the three stubs, 27,000 states must be tested. This comprehensive set of tests was repeated for the open circuit, short circuit, and 50Ω load cases illustrated in Fig. 3.25 and Fig. 3.26(a) from 100 to 375MHz. Each load termination produces its own set of synthesized load impedances that does not necessarily sample the entire Smith Chart. Fig. 3.28 shows the achieved measured impedances at 175MHz where the open circuit load samples the right outer perimeter of the Smith Chart best. The short circuit and 50Ω loads sufficiently sample the rest of the Smith Chart. Combining all 3 measured solution sets as in Fig. 3.29, Smith Chart coverage to a VSWR 30:1 is achieved from 125 to 300MHz. A finer resolution in voltage step (.25V instead of 1V) would allow for a more finely sampled Smith Chart. Below 125MHz the inter-stub lengths are too electrically short for uniform impedance coverage. Above 300MHz the tuner is still useful for smaller VSWR circles (15:1, 10:1, and 5:1 at 325, 350, and 375MHz respectively). These results demonstrate the usefulness of this low cost VIS as a viable means of producing a wide range of impedances for device power gain characterization.

The method of operation for the tuner is depicted in Fig. 3.25. With the tuner fully characterized vs frequency for each of the termination types (open, short, 50Ω), a lookup table can be used to automatically synthesize a desired impedance by setting the proper bias voltage on \( C_{bank} \) and switching to the proper termination. Once a load impedance is set, the network analyzer would conduct a measurement from voltages a, b, c, and d. From this information the computer will compute the power gain, and repeat this process to characterize the desired region on the Smith Chart.
It is noted that the specific measurements in Fig. 3.29 were taken at 0 dBm power level. However, it is well known that varactors are non-linear devices producing intermodulations and it is therefore important to address this issue. To do so, two varactors in a back-to-back or anti-series configuration can be used for higher power applications [26,59].

A low cost, small (1.75x1.75in) VHF impedance synthesizer was presented. The synthesizer operates from 125 to 300MHz. This impedance synthesizer exploits normally un-desired losses in varactor diodes to produce a wide range of load impedances bounding a VSWR 30:1 circle on the Smith Chart. It was shown that by using three different terminations for the circuit, three sets of $Z'_L$s are created. Together they broadly sample the Smith Chart over a 2.4:1 bandwidth creating a robust low-cost tuner for power gain measurements. The concept of exploiting component losses combined with different terminations for tunability is a novel method to design low cost simultaneously frequency and impedance agile tuner for load-pull device characterization.
Figure 3.29: Measured impedance coverage vs frequency for Fig. 3.26 triple stub VIS. Open, Short, and 50Ω termination solutions are combined for each frequency, VSWR 30:1 Smith Chart coverage achieved from 125 to 300MHz.
Chapter 4: FAR-FIELD SENSOR FOR RECONFIGURABLE ANTENNA FEED-BACK LOOP

In Chapter 3, several directional coupler based tuning topologies were studied as a means to ensure transducer gain (TPG) is maximized in an adaptive tuning system. This approach has the advantage of being well integrated with the impedance tuner, and when designed properly, is indicative of far field radiated power. However, it can be costly as one or two directional couplers must be used with a number of detection diodes. The circuit footprint is increased as proper biasing for diodes and space for couplers must be accounted for. Finally, the directional couplers add unavoidable insertion loss to the network, and have their own band-limited operation.

To overcome some of these limitations, an alternative tuning topology is presented in Fig. 4.1(a). As the tuner becomes better matched to the antenna, the antenna will radiate more energy which will couple to the sensor as depicted in Fig. 4.1(b). This topology is advantageous as it only requires one RF to analog detector and avoid incurring the insertion loss associated with a directional coupler. Flexibility in the probe design also allows specific bands of operation and the ability to choose exact coupling coefficients based on system design constraints such as transmit power. With the directional coupler topologies, one is limited by whatever is commercially available. Also, the tuning algorithm has the simple goal of maximizing the analog
Figure 4.1: (a) Alternative tuning topology with near field radiated power detector and (b) Near field probe to antenna coupling, as the antenna is better matched more power is radiated creating a stronger magnetic field inducing more current on the probe.

voltage (D) from the detection diode connected to the sensor. This is in contrast to weighting a number of monitored voltages in the directional coupler topologies. While this concept sounds favorable in theory, care must be taken when designing the near field probe. Since the probe is in the near field, it samples both the near field reactive energy inherent to small antennas, as well as the far field radiated power. As such it is possible for the probe’s response to be dominated by the near field reactive energy, hence the probe may not properly track peak realized gain.

This chapter begins with a discussion on how to decide whether a E field or H field probe should be used for integration in a tuning system. Design guides are provided for several different antennas including the monopole, loop, slot, and patch antenna. Considerations for system transmit power, probe sensitivity, and detection diode selection are also covered.
4.1 Sensing E-Field or H-Field

E and H near field probes have been extensively used in the areas of electromagnetic compatibility (EMC) / electromagnetic interference (EMI) diagnostic testing [62–69]. It is generally understood that small pin or monopole probes are E field selective while small loops are H field selective. Commercial EMI probes such as [70] have been shown to provide up to 30 to 40dB of E/H or H/E rejection for E and H field probes respectively. To properly design a near field probe to act as a far field radiated power sensor, it is necessary to understand why a small pin is E field selective and why a small loop is H field selective. Consider Fig. 4.2 reproduced from [71]. Here the wave impedance, $Z_W$, is compared for an electric and a magnetic source.

![Figure 4.2: Wave impedance ($Z_W$) as a function of distance from electric and magnetic sources.](image-url)
The wave impedance is the ratio of the electric and magnetic fields:

\[ Z_W = \frac{E}{H} \]  \hspace{1cm} (4.1)

From inspection of equation 4.1, it is easy to conclude that a pure electric field (E) has a high wave impedance and a pure magnetic field (H) has a low wave impedance. However, the distance from a particular EM source also factors into the wave impedance equation. From [71], the formula for a magnetic source is:

\[ Z_W = 120\pi \frac{r}{\sqrt{r^2 + 1}} \]  \hspace{1cm} (4.2)

And for an electric source:

\[ Z_W = 120\pi \frac{\sqrt{r^2 + 1}}{r} \]  \hspace{1cm} (4.3)

where \( r \) is the distance from the source. Observing Fig. 4.2, it is seen that as the distance from the source increases, \( Z_W \) changes, and each field begins to produce its complementary field. The H field wave impedance increases with the distance from the source (the near field), and the E field wave impedance decreases. In the transition region, fields emitted from both sources end up with the same wave impedance, and remains the same through the far field. Here \( Z_W \) becomes equivalent to the characteristic impedance of the propagation media, 377\( \Omega \) for air. Thus, beyond the transitional region, the ratio of E and H fields becomes the same for all signals regardless of how they began.

Since the wave impedance in the near field is a consequence of the source it comes from, Ohms law \( (R = V/I) \) also characterizes the field. For the electric source, the voltage (V) is strong and the current (I) is weak. Conversely, for the magnetic source,
there is a strong current and a weak voltage. With this understanding, it is of interest to observe the ratio of E/H in the near field for each source. Fig. 4.3 plots the HFSS full wave simulated ratio of the complex total |E field| to the complex total |H field| in the plane of a $\lambda/100$ monopole and loop on an infinite PEC ground plane. A solid green color relates to a ratio of 377, where the ratio of E and H are the same. In Fig. 4.3(a) the region encompassing the small monopole is a solid red color, a high E/H ratio, indicating E field selectivity (strong V and weak I). In Fig. 4.3(b) the region encompassing the small loop is dark blue, a low E/H ratio, indicating H field selectivity (weak V and strong I). The E/H ratio plot of the small loop is non-symmetric because the ideal source excitation is on the left side of the loop. Plots are limited to a rectangular plane of $0.5x0.5\lambda$ in area. Of course, if the E/H plots in Fig. 4.3 extended to the far field, the same 377 ratio would be observed for both antennas as shown in Fig. 4.2.

In practice a purely electric or purely magnetic field cannot be realized. As outlined in [71], a very small monopole will have a high voltage excitation generating a strong E field, but will always carry some current to ground via parasitic capacitance creating a weak H field. Likewise, a small loop will carry a substantial current causing an H field, yet will still carry some small voltage that will create an E field. As such, these sources generate quasi-static waves in the near field because they approximate the electromagnetic fields produced by static current and charge distributions. Therefore, the far field E and H field equations for a Hertzian dipole must simplify to a function that is similar to the static case in the near field. From [72], for a Hertzian dipole antenna of current $I = \hat{a}_z I$ and length $\Delta L$ located at the origin of the coordinate system:
Figure 4.3: Full wave simulated $\frac{\text{Complex } |E_{\text{total}}|}{\text{Complex } |H_{\text{total}}|}$ ratio for $\lambda/100$ (a) monopole and (b) loop. High E/H ratio implies E field selectivity, low E/H ratio implies H field selectivity.
\[ \overline{E} (\tau) = \hat{a}_r E_r + \hat{a}_\theta E_\theta \ [V/m] \] (4.4)

where

\[ E_r = -\beta^2 \eta \frac{2I \Delta L \cos \theta}{4\pi} \left[ \frac{1}{(j\beta r)^2} + \frac{1}{(j\beta r)^3} \right] e^{-j\beta r} \] (4.5)

\[ E_\theta = -\beta^2 \eta \frac{I \Delta L \sin \theta}{4\pi} \left[ \frac{1}{(j\beta r)} + \frac{1}{(j\beta r)^2} + \frac{1}{(j\beta r)^3} \right] e^{-j\beta r} \] (4.6)

and

\[ \overline{H} (\tau) = -\hat{a}_\phi \beta^2 \eta \frac{I \Delta L \sin \theta}{4\pi} \left[ \frac{1}{(j\beta r)} + \frac{1}{(j\beta r)^2} \right] e^{-j\beta r} \ [A/m] \] (4.7)

As the electric distance from the Hertzian dipole source goes to zero (i.e. the near field: \( \beta r \to 0 \)) the \( \frac{1}{(j\beta r)^2} \) term in equation (4.5) can be neglected with respect to the \( \frac{1}{(j\beta r)^3} \) term. Likewise in equation (4.6), both the \( \frac{1}{(j\beta r)} \) and \( \frac{1}{(j\beta r)^2} \) terms can be neglected with respect to the dominant \( \frac{1}{(j\beta r)^3} \) term. After applying the following expansion:

\[ e^{-j\beta r} = 1 - j\beta r - \frac{(\beta r)^2}{2} - \ldots \] (4.8)

Only the first term is kept as \( \beta r \to 0 \). Applying these operations to equation (4.5):

\[ E_r \approx -\beta^2 \eta \frac{2I \Delta L \cos \theta}{4\pi} \frac{1}{(j\beta r)^3} = \frac{2I \Delta L \cos \theta}{j4\pi r^3 \beta/\eta} \] (4.9)

applying the relationship:

\[ \frac{\beta}{\eta} = \frac{\omega \sqrt{\mu/\epsilon}}{\sqrt{\mu/\epsilon}} = \omega \epsilon \] (4.10)
and equation (4.9) simplifies to:

\[
E_r \approx \frac{2I\Delta L\cos\theta}{j\omega\epsilon 4\pi r^3}
\] (4.11)

Applying the substitution \( Q = I/j\omega \) equation (4.11) becomes:

\[
E_r \approx \frac{2Q\Delta L\cos\theta}{\epsilon 4\pi r^3} \text{ [V/m] (for } \beta r << 1) \] (4.12)

Applying the same process to \( E_\theta \), equation (4.6) simplifies to:

\[
E_\theta \approx \frac{Q\Delta L\sin\theta}{\epsilon 4\pi r^3} \text{ [V/m] (for } \beta r << 1) \] (4.13)

The \( E_\theta \) and \( E_r \) electric field components for the near fields of the Hertzian dipole has the same form for an electric dipole with \( \vec{p} = \hat{a}_z Q\Delta L[C - m] \) from [73]:

\[
\overline{E}_F = \frac{p}{4\pi \epsilon r^3} (\hat{a}_r 2\cos\theta + \hat{a}_\theta \sin\theta) \text{ [V/m]} \] (4.14)

After substituting equation (4.12) and equation (4.13) back into equation (4.4) for the near field case, equation (4.4) has the same form as equation (4.14) where equation (4.4) contains phasors and equation (4.14) is a static field.

Following the same procedure for equation (4.7), keeping the dominant terms as \( \beta r \to 0 \) gives:

\[
H_\phi \approx \frac{I\Delta L\sin\theta}{\epsilon 4\pi r^2} \text{ [A/m] (for } \beta r << 1) \] (4.15)

Which is the same form as the static magnetic field produced by a current element \( I\Delta L \) applying the Biot-Savart law for magnetostatics (not shown here). The same
exercise, simplifying the E and H field equations for a small loop as $\beta r \to 0$, would also produce the associated electrostatic and magnetostatic functions.

In summary, the E and H fields electrically close ($\beta r \to 0$) to the Hertzian dipole have the same form as the associated static case with the distinction that the fields oscillate with time. Again, this demonstrates the quasi-static nature and field selectivity of the small antenna in the near field. Since the probe is a very electrically small antenna, it of course is a reciprocal device. By integrating an appropriate probe in a tunable antenna, it will be necessary to ensure an associated quasi-static radiating E or H field is sampled. The design guides presented in the following sections will study the E/H ratio of the near field for each antenna to appropriately choose the proper E or H field probe. Each antenna / probe is designed to be tuned between 200 to 400MHz.

4.2 Outline of Sensor Design Guidelines

There are several guidelines and concepts that will be used in the following sections to properly integrate a small probe in the near field of the antenna that properly tracks far field radiated power. They are outlined here and useful for all antennas, beyond the four that are explored in this chapter.

- Observe $\text{Complex}|E_{\text{Total}}|/\text{Complex}|H_{\text{Total}}|$ in the near field of the antenna. By observing the ratio of E/H, over the region of the antenna where a probe would be integrated, the designer gains an understanding of which field should be probed. However, the E/H ratio is not constant across frequencies. Thus E/H should be observed over the tunable band. Furthermore, an impedance tuning unit may not always be applicable. As will be discussed, some antennas have
real / imaginary impedances outside of a matchable tuning range, especially when radiation resistances are very small. In these cases the antenna aperture can be modified to integrate tunable components (variable L’s and C’s). As such the electrical length of the antenna is directly tuned creating different current distributions on the antenna for a given frequency. In these cases, E/H must be observed vs frequency and the tunable range of the reconfigurable components.

• Choose to probe E or H (small pin or loop). After observing the E/H ratio of the antenna, the designer should know if the E or H field should be probed. An appropriate small probe is then selected for integration with the antenna.

• Pick probe location and ensure proper polarization. Once an E or H probe is chosen, its location on the antenna must be chosen. If an E probe is chosen, the region selected for integration should be E dominant for the tunable band. Furthermore the probe will have its own radiation pattern / polarization. As such, it must be properly oriented co-polarized with the chosen E or H field. Obviously if this is neglected, a properly chosen probe may not track far field radiated power as very low coupling between the probe and antenna occurs due to polarization mismatch. Depending on how wide the tunable band is, more than one probe, probe type, and location may be necessary to properly sense radiated power.

• Ensure antenna and probe coupling is below -15dB. At -15dB coupling, the probe will reduce antenna gain by 0.14dB. However, the probe also adds a geometric change to the antenna. As such, the more dramatic the change (more coupling) the more the input impedance of the antenna will be affected by the probe. A
change in probe size will more dramatically affect an electrically small antenna than a resonant sized antenna. As a general rule -15dB of coupling is followed to minimize the affect on the antenna $|S - Parameters|$ and gain. While this is a general rule to follow when starting a design, if more coupling is required, the affects of the probe on the antenna should be studied. The affect of the probe on the antenna may shift the antenna input impedance outside the tunable range if coupling and geometric changes to the antenna are not considered.

- Choose detection diode based on antenna / probe coupling. The coupling coefficient between the antenna and probe will set the dynamic range and sensitivity required by a detection diode connected to the near field probe to provide feedback to the closed loop tuner. For example, at -15dB of coupling and a 1 Watt (30dBm) transmitter, a detection diode must be able to sense a 15dBm signal. As the antenna is tuned there will be a range of coupling coefficients at a given frequency, for example -15dB down to -60dB for a matched to mismatched condition respectively. This sets the dynamic range requirement for the detection diode of 45dB. Thus the selected detection diode must have the correct sensitivity and dynamic range for a given transmit power for the designed probe across the tunable frequency band.

### 4.3 Monopole Antenna Probe Design Guide

In this section, a near field probe that samples far field radiated power will be designed and integrated with an electrically small meandered monopole. This same monopole will be implemented in the next chapter with an impedance tuner and fabricated for measurements and validation. The small antenna to be tuned is shown
in Fig. 4.4 with dimensions in Table 4.1. The antenna is 3in x 5in, electrically $\approx \lambda/10$ at the lowest design tunable frequency of 200MHz. The space for the meandered monopole is 3in x 1.5in, $\approx \lambda/17$ at 200MHz. Ground plane exists on both the top and bottom of the PCB as the tuner will be implemented in grounded coplanar waveguide (GCPW). The antenna un-tuned simulated impedance at $Z_S$ is shown in Fig. 4.5, the loss tangent of the dielectric is included in the model ($\tan\delta = 0.0022$).

![Figure 4.4: (a) Electrically small meandered monopole (b) PCB side view.](image)

Over the 2:1 band, $Z_{\text{Real}}$ varies from 0.63 to 971$\Omega$ and $Z_{\text{Imag}}$ varies from $-1.6k$ to $1.8k\Omega$. This creates a worst case VSWR of 725:1 for an impedance tuner to handle. These values are within reasonable range for a realistic impedance tuner. Although, with a radiation resistance of 0.63$\Omega$, losses in the matching network may become a concern.

The following steps will be followed as a design guide for each antenna in this and subsequent sections:
Table 4.1: Fig. 4.4 Meandered Monopole Dimensions

<table>
<thead>
<tr>
<th>Label</th>
<th>Dimensions</th>
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<tr>
<td>GPW</td>
<td>3in</td>
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<tr>
<td>GPL</td>
<td>3.5in</td>
</tr>
<tr>
<td>FeedD</td>
<td>7.1mm</td>
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<tr>
<td>FeedL</td>
<td>6.5mm</td>
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<tr>
<td>ATW</td>
<td>1.5mm</td>
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<tr>
<td>D</td>
<td>9.4mm</td>
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<tr>
<td>ATL</td>
<td>1.5in</td>
</tr>
<tr>
<td>SubH</td>
<td>50mil</td>
</tr>
</tbody>
</table>

1. Observe $|E_{Total}|/|H_{Total}|$ in the near field of the antenna over the operating band

2. Choose to probe E or H (small pin or loop)

3. Pick probe location (ensure probe is polarized correctly with desired field)

As shown in Fig. 4.4(b), the sensor will be integrated on the bottom ground plane, underneath the plane of the antenna. As such, fields will be plotted 5mil beneath the probe’s ground plane. Fig. 4.6 shows the simulated E and H fields at 250MHz. For integration with the system, the probe will protrude from the ground plane and attached to a detection diode.

Either E or H should be sampled between the ground plane and the first turn of the meandered monopole. For both E and H in Fig. 4.6(a) and (b) respectively, strong components exist in this region of the antenna. Based on the scales for each plot, the H field appears to be more dominant. Regardless, it is difficult to determine if E or H should be probed without trial and error probe placement. After computing the E/H ratio in Fig. 4.6(c), probe type becomes evident. The edge of the antenna opposite
Figure 4.5: (a) Small monopole un-tuned imag / real impedance at $Z_S$ in Fig. 4.4 and (b) Associated Smith Chart from 200 to 400MHz.

the excitation is E dominant while the space between the middle of the antenna and the feed is more H dominant. However, we cannot choose a probe location quite yet as the E/H distribution will change at different frequencies, we have only seen the solution at 250MHz.

The E/H ratio is plotted from 200 to 400MHz for the meandered portion of the monopole antenna in Fig. 4.7. Three positions are marked in the 200MHz plot and studied across the operating band. At Pos. 1, from 200 to 250MHz the region starts off E dominant (200MHz), becomes E and H dominant (ratio of 377, 225MHz), and ends H dominant (250MHz). From 275 to 325MHz, Pos. 1 remains H dominant. Then from 350 to 400MHz, Pos. 1 is E dominant. This would be a poor location as neither a small monopole or loop would be sensitive to the proper field over the entire band. Pos. 2 looks more promising, it is E dominant for most frequencies, however
Figure 4.6: Simulated monopole E and H fields at 250MHz.
at mid band (255 to 325MHz) Pos. 2 is H dominant. Across the band, Pos. 3 is E field dominant everywhere except at 325MHz. However, as reviewed in Section 4.1, there will still be a quasi-static E field at Pos. 3 for 325MHz. Therefore, the selection of a small monopole may work over this 2:1 bandwidth. While not plotted here for brevity, for Pos. 3 at all frequencies, the E field is vertically polarized with respect to the figure between the ground plane and the first meandering for the wire. This validates that a small monopole protruding from the ground plane will be co-polarized with the field it is to sample.

A small near field probe is integrated in the bottom ground plane (opposite the monopole) in Fig. 4.8 near Pos. 3 from Fig. 4.7. The small monopole extends only 1mm from the ground plane and has an impedance of 100Ω as this will be the impedance of the detection diode integrated in the next chapter. A via fence is placed around the probe to shield it from any standing waves that might exist in the dielectric between the ground planes due to the stubbed impedance tuner (see next chapter). Before analyzing the effectiveness of this probe as a far field radiated power detector, a parametric study of the probe length is conducted to understand its affect on the input impedance on the antenna.

4.3.1 Antenna / Probe Coupling Parametric Study

Fig. 4.9 shows the parametric study where the probe length is varied from 0 to 9mm. \(|S_{11}|, |S_{21}|, \text{Peak Realized Gain, and the Smith Chart are shown for comparison as the probe length changes. For this study, the monopole is tuned to } \approx 299\text{MHz in simulation. When the probe is 0mm in length, i.e. does not protrude past the ground plane, there is -35.5dB of coupling (|S_{21}|) from the tuned monopole to the probe.}

From
<table>
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<th>Frequency (MHz)</th>
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**Figure 4.7:** E/H ratio vs frequency for meandered portion of monopole.

**Figure 4.8:** Radiated power sensor integrated with monopole antenna.
Figure 4.9: E field probe length parametric study for (a) $|S_{11}|$, (b) $|S_{21}|$, (c) Peak Realized Gain, and (d) Smith Chart. $|S_{21}|$ coupling of -15dB or more (probe length of 4mm) perturbs the input impedance and gain of the antenna. Final design uses 1mm length, -28dB of coupling.
0 to 3mm, the coupling gradually increases to -19dB, and the antenna parameters are consistently un-changed as observed in the peak realized gain, $|S_{11}|$, and Smith Chart plots (blue dashed traces lie on top of the 0mm solid red trace). At a probe length of 4mm (solid green trace), the probe begins to perturb the performance of the meandered monopole. This occurs at about -15dB of coupling between the monopole and probe. Realized gain decreases and the input impedance of the monopole changes. As expected, from 4 to 9mm probe lengths, antenna parameters continue to change as the coupling increases between -15dB and -5dB.

This study provides a general “rule of thumb” when designing near field probes for this application. To ensure the radiating antenna performance is not perturbed by the probe, coupling between the Tx antenna and the Rx probe should be less than -15dB. Of course, as a general rule, coupling should be kept low as the goal is for the antenna to radiate into the far field. A significant percentage of energy being received by the probe would hinder far field gain. Fig. 4.9 marks where the 4mm length (-15dB coupling limit) exists physically, as well as the 1mm design length used in the final design. At 1mm, the $|S_{21}|$ coupling is -28dB.

4.3.2 Free Space Probe Performance

The probe design in Fig. 4.8, is simulated for tuning conditions in free space to assess performance. Fig. 4.10 shows the simulated Peak Realized Gain and $|S_{21}|$ vs frequency from 50 to 600MHz. It is important that Peak Realized Gain is plotted, the direction of peak gain will change as higher order modes are excited by the antenna at higher frequencies. As the probe was designed for the 200 to 400MHz band, Peak Realized Gain vs Frequency and $|S_{-Parameters}|$ vs Frequency are shown in
Figure 4.10: Peak Realized Gain and $|S_{21}|$ vs frequency. Probe $|S_{21}|$ tracks peak gain from 50 to 400MHz.
Fig. 4.10(a) and (b) respectively. To simulate tuning, the source impedance \( Z_S \) of the meandered monopole is swept from 1 to 5\( k\Omega \). The monopole’s \( S_{11} \) and gain show optimal tuning at 300MHz however, across the 2:1 band, the gain varies as \( Z_S \) is swept. Likewise, the \( |S_{21}| \) coupling between the antenna and probe also varies. If the probe properly tracks the gain, then as the gain increases, so should the \( |S_{21}| \). For all the tuning cases in Fig. 4.10(a)(b), it is difficult to determine if this is the case. As such, the Peak Realized Gain is plotted versus \( |S_{21}| \) from 200 to 400MHz in Fig. 4.10(e). This makes it easier to determine if the probe tracks far field radiated power. In studying Fig. 4.10(e), it is seen that as the gain increases for a particular frequency, so does subsequent \( |S_{21}| \). Indeed, for the 2:1 band the probe properly samples the peak gain.

It is noted that at 325MHz the coupling between the antenna and probe is the lowest (dip in the graph at 325MHz in Fig. 4.10(b)). This is explained by examining Fig. 4.7 where at 325MHz the probe is positioned in an H field dominant location, at all other locations, the E field probe is at a E field dominant location. As stated earlier, despite a dominant H field at 325MHz, there still exists a quasi-static changing E field, albeit weaker in magnitude, hence the probe is still able to track far field gain at this frequency. An alternative would be to use a small loop at this frequency, however this may not be suitable at other frequencies.

Considering Fig. 4.10(c)-(e), the probe tracks gain linearly from 50 to 400MHz. Above 400MHz there are several dips in the graph at some frequencies where a subsequent \( |S_{21}| \) does not necessarily correlate to higher gain. This is because the probe is located in an H field dominant location, the probe is dominated by near field stored
energy. However, if the system tuning algorithm can ensure a global maximum is found, the probe is still usable up to 500MHz.

These simulations also give an idea of the dynamic range and sensitivity required by the detection diode. As this tuner is designed to work from 200 to 400MHz, Fig. 4.10(e) shows us that the detection diode should provide at least 35dB of dynamic range ($|S_{21}|$ from -63 to -28dB). Assuming a 5 Watt (37dBm) transmitter, sensitivity from 9dBm down to -26dBm is required. Of course a different transmit power may require more or less coupling to the probe depending on available diode power sensitivity. However, as stated in Section 4.3.1, the coupling should be less than -15dB.

### 4.3.3 Probe Performance Near Infinite Ground Plane

The previous section showed that the probe was able to track far field gain for the designed 2:1 band in free space. However, applications dictate using antennas in different environments unknown to the antenna designer. As such it would be of interest to verify if the probe still tracks gain for different loading affects on the antenna. While there are an endless number of different coupling affects that could be studied near different dielectric medium and metallic loading etc..., it is unrealistic to study them all. The author has chosen to study a few worst case scenarios, the antenna placed close to an infinite perfect electric conductor (PEC). Four configurations are outlined in Fig. 4.11. The first in Fig. 4.11(a) places the infinite PEC sheet 1” ($\lambda/39$ at $f_c = 300MHz$) parallel to the bottom side of the antenna (Config. 1). Config. 2 and 3 place the sheet 1” away from the antenna excitation side and probe side of the aperture respectively (perpendicular to the plane of the antenna). Lastly, Config. 4
Figure 4.11: Outline of four environmental configurations to test probe performance as antenna is tuned. Antenna is placed 1" away from infinite PEC sheet ($\lambda/39$ at $f_c$ of 300MHz).

places the PEC sheet perpendicular to the plane of the antenna, 1" away from the top of the radiating edge of the meandered monopole.

The simulation criteria from Fig. 4.10 in free space are repeated for the configurations in Fig. 4.11. $Z_S$ of the meandered monopole is swept from 1 to 5kΩ. Peak Realized Gain vs $|S_{21}|$ are plotted from 50 to 600MHz. As stated previously, it is important that peak gain is plotted, the direction of peak gain will change vs frequency and ground plane placement. Simulation results for the four configurations in Fig. 4.11 are reported in Fig. 4.12 - Fig. 4.15. Despite drastic changes in environment from the free space configuration, with a ground plane placed $\lambda/39$ at the center frequency ($f_c$), the probe still linearly tracks far field radiated power from 50 to 400MHz. Like the free space configuration, if the system tuning algorithm can...
Figure 4.12: $|S_{21}|$ vs Peak Realized Gain for Config. 1 from Fig. 4.11.

To ensure a global maximum is found, the probe is still usable up to 500MHz for all four ground plane configurations.

The ground plane study confirms the usefulness of the probe in dynamic environments. Despite being in the near field, the probe can be properly designed to track far field gain. While this section went into depth concerning antenna / probe coupling and the performance electrically close to a ground plane, such studies will not be repeated for the subsequent antennas for brevity.
Figure 4.13: $|S_{21}|$ vs Peak Realized Gain for Config. 2 from Fig. 4.11.
Figure 4.14: $|S_{21}|$ vs Peak Realized Gain for Config. 3 from Fig. 4.11.
Figure 4.15: $|S_{21}|$ vs Peak Realized Gain for Config. 4 from Fig. 4.11.
4.4 Loop Antenna Probe Design Guide

In this section a near field probe is designed for an electrically small loop antenna. The small loop is depicted in Fig. 4.16 with dimensions in Table 4.2. The antenna is 3in x 3in and resonant at 1.2GHz, electrically the antenna is $\approx \lambda/14$ at the lowest design tunable frequency of 200MHz. The space for the loop’s wire occupies a 1in x 2.5in area, $\approx \lambda/22$ at 200MHz. Ground plane exists on the top and bottom of the PCB, where the probe will be integrated on the opposite side of the antenna. The loop antenna un-tuned simulated impedance at $Z_S$ is shown in Fig. 4.17, the loss tangent of the dielectric $\epsilon_r = 10$ is included in the model ($\tan\delta = 0.0022$). Over the 2:1 band (200 to 400MHz), $Z_{\text{Real}}$ varies from 0.08 to 10Ω and $Z_{\text{Imag}}$ varies from 150 to 740, with a worst case VSWR of 6250:1 in a 50Ω system. Due to the extreme mis-match and low radiation resistance, a realizable external impedance tuner would not be suitable for this small antenna. Alternatively, reconfigurable components can be integrated with the antenna for tuning.

A variable capacitor is placed on the opposite end of the feed of a resonant loop to tune to different frequencies in [74–76]. Alternatively, two variable inductors, or a set of tunable capacitors in series with fixed lumped inductors can be used to tune an electrically small loop. Such a configuration is presented in [77]. The latter of the two configurations are used in this work and depicted in Fig. 4.18 (Fig. 4.18(b) reproduced from [77]). It is noted that while a single inductor could be placed on the middle of the top vertical wire in Fig. 4.18(a), for a resonant condition, this is where the currents cancel. Alternatively, the inductors are placed on either edge of the vertical wires (where currents are maximum at resonance) to produce a phase delay, hence electrically tuning the length of the wire.
Figure 4.16: (a) Electrically small loop antenna (b) PCB side view.

Table 4.2: Fig. 4.16 Loop Antenna Dimensions

<table>
<thead>
<tr>
<th>GPW</th>
<th>GPL</th>
<th>SubL</th>
<th>ATH</th>
<th>ATL</th>
<th>ATW</th>
<th>OS</th>
<th>SubH</th>
</tr>
</thead>
<tbody>
<tr>
<td>3in</td>
<td>1.75in</td>
<td>1.25in</td>
<td>1in</td>
<td>2.5in</td>
<td>1.5mm</td>
<td>7.1mm</td>
<td>50mil</td>
</tr>
</tbody>
</table>

Figure 4.17: (a) Small loop un-tuned imag / real impedance at $Z_s$ in Fig. 4.16 and (b) Associated Smith Chart from 200 to 400MHz.
By using two, as opposed to one inductor, the current distribution inherent to conventional loops is maintained at resonance. Electrically tunable inductance is difficult to produce, thus the proposal of a lumped inductor with a tunable capacitor in Fig. 4.18(a) for a realizable implementation. However, for simplicity, simulations are conducted assuming a variable inductor.

As discussed in Section 4.2, the design approach will be to observe the E/H ratio across the tunable band, choose to probe the E or H field, determine probe placement, and ensure the sensor is co-polarized with the probed field. Fig. 4.19 shows the complex total $|E|$ and complex total $|H|$ as well as their ratio. Like the small monopole, the probe will protrude from the ground plane under the loop. For both E and H in Fig. 4.19(a) and (b) respectively, strong components exist between the antenna and ground plane. Based on the scales for each plot, the H field appears to be more dominant. Regardless, it is difficult to determine if E or H should be probed without trial and error probe placement.
Figure 4.19: Simulated loop E and H fields at 250MHz (L = 225nH).
The E/H ratio in Fig. 4.19(c) helps determine the proper probe type. Most of the region between the ground plane and antenna is E field dominant ($E/H > 377$, solid red), especially at the center of the antenna. Again, this demonstrates the usefulness of examining the E/H ratio when designing the probe. As was the case with the small monopole, the E/H ratio must be studied vs frequency to choose the correct probe location. However, unlike the monopole probe design, the loop has been treated with a variable inductor that changes the current distribution at a given frequency as it is tuned. This adds another dimension when designing the probe, E/H must be observed over the operating band and over the variable inductor tuning range.

Fig. 4.20 shows the near field E/H ratio for selected tuned frequencies (204.5, 294.5, and 398MHz) for the low, middle, and high inductor tuning values (180, 325, and 675nH). The loop is tuned to resonance for the following scenarios: 204.5MHz at 675nH; 294.5MHz at 325nH, and 398MHz at 180nH. Unlike the monopole antenna, where the aperture electrical length remained constant at a given frequency, the tunable loop antenna has electrically small, resonant, and electrically long (higher order mode) conditions. Picking a probe location that will satisfy the low, middle, and high frequencies can be tricky for the following reason. At 398MHz, resonance is tuned at 180nH, for all other inductor values, at 398MHz, the antenna is made “electrically long” by the phase shift introduced by the inductors. Hence a suitable probe location at this frequency must be chosen based on a resonance and above resonance condition. At 294.5MHz the probe must operate for an electrically small case (below 325nH), resonant condition (325nH) and an electrically long case (675nH).
Figure 4.20: E/H ratio vs frequency and tuning L value for small loop. Loop is tuned to resonance at the following conditions: 204.5MHz at L=675nH; 294.5MHz at L=325nH; and 398MHz at 180nH.
Alternatively, at 204.5MHz the probe is always sampling for either an electrically small condition or resonant condition, the inductor tuning range does not allow for an electrically long condition at the lowest tunable frequency.

In the E/H plots of Fig. 4.20, symmetric behavior is observed for the resonant condition at each frequency. At Pos. 1, the antenna is E dominant both below and above resonance, but becomes H dominant at resonance. Also, this position is quite close to the excitation where coupling between the probe and antenna may be too high. Pos. 3 remains mostly H dominant for all tuning conditions, but is quite close to the radiating edge of the antenna, coupling may be too high here. Pos. 2 is E dominant except for the electrically long condition where it starts to become H dominant. Ultimately an E field probe (small monopole) is chosen between Pos. 1 and Pos. 2. At this location the small monopole is co-polarized with the E fields of the radiating loop. The probe is shown in Fig. 4.21 and protrudes from the ground plane 1mm.

![Diagram of loop antenna and E-field probe](image)

**Figure 4.21:** Radiated power sensor integrated with loop antenna.

Free space simulated probe performance, Peak Realized Gain and $|S-Parameters|$ vs frequency for Fig. 4.21 is shown in Fig. 4.22 from 50 to 600MHz. Fig. 4.22(a) and
Figure 4.22: Peak Realized Gain and $|S_{21}|$ vs frequency. Probe $|S_{21}|$ tracks peak gain from 200 to 400MHz.
Figure 4.23: E/H ratio vs Frequency and tuning L value for small loop from 50 to 175MHz.
(b) show the Peak Realized gain and $|S - Parameters|$ from 200 to 400MHz where the inductors (L) are swept from 180 to 675nH to impedance match the loop to different frequencies. As described previously, Fig. 4.22(a) and (b) are used to create the Peak Realized Gain vs $|S_{21}|$ in Fig. 4.22(e). For the designed tuning band the probe/loop $|S_{21}|$ response properly tracks peak gain. The coupling response at each frequency is quite similar hence the curves in Fig. 4.22(e) lie on top of one another. If this design were to be implemented, we see that the maximum coupling is $\approx -25\,dB$, below the -15dB threshold so as not affect antenna performance. For detection diode selection, 60dB of dynamic range would be sufficient sensitivity.

Due to the limited tuning range of the inductors, the antenna cannot be properly tuned from 50 to 175MHz (Fig. 4.22(d)). Furthermore, for most of the lower frequencies (50 to 125MHz), the probe does not track the realized gain. An explanation of this is depicted in the E/H ratio plots vs Frequency and inductor tuning plots for this band in Fig. 4.23. At the lowest frequency of 50MHz and the smallest inductance (180nH), most of the region around the loop is H dominant. As the frequency increases and the inductance increases, the loop becomes more E dominant as it becomes less electrically small. This is why at 175MHz the probe does track peak gain. As such the probe type and location is not suited for this frequency and inductor tuning range. If a wider inductor tuning range was feasible, this problem would be remedied by using two probes, one small loop probe for low frequency low inductance, and the small monopole probe as the inductance increased and began to approach resonance.

Fig. 4.22(c) shows the probe response from 425 to 600MHz. For this entire band, the inductor tuning range has forced an electrically large antenna condition (higher
order modes). An increase in peak gain as the inductance is tuned does not necessarily correlate to an increase in $|S_{21}|$. For some tuning conditions the probe is in an H field dominant location. Again, this problem would be solvable with the use of more than one probe / location. However, if the system tuning algorithm can ensure a global maximum solution is found, the designed probe is still useful in this band.

4.5 Slot Antenna Probe Design Guide

This section will cover the near field probe design for a resonant sized slot antenna. In sections 4.3 and 4.4, the near field sensor was designed for electrically small antennas. However, there are applications where it is desired to tune a resonant sized narrow band antenna to different frequencies. In a static environment, such an antenna would not need a feedback sensor for tuning, a lookup table would be sufficient for tuning. However, in an application where the antenna’s platform might change, a feedback sensor becomes necessary. The slot antenna is depicted in Fig. 4.24 and corresponding dimensions in Table 4.3.

The slot antenna can be tuned by an impedance tuner or by direct reconfiguration of the aperture similar to the loop antenna. One approach to tune the aperture is to place variable capacitors across the slot to tune the antenna’s electrical length as in [78,79]. A compact slotted loop antenna employs the same concept in [80]. Alternatively, a cavity backed slot antenna is loaded with reconfigurable shorting posts with PIN diodes to manipulate the field distribution within the cavity hence tuning the antenna in [81].

The slot is $\lambda/2$ at 325MHz, modeled on a 62mil thick $\epsilon_r = 2.2$ ($\tan\delta = 0.0009$) single side copper cladded PCB.
Figure 4.24: (a) $\lambda/2$ 325MHz resonant sized slot antenna (b) PCB side view.

Table 4.3: Fig. 4.24 Slot Antenna Dimensions

<table>
<thead>
<tr>
<th>GPW</th>
<th>GPH</th>
<th>SlotW</th>
<th>SlotT</th>
<th>SubH</th>
</tr>
</thead>
<tbody>
<tr>
<td>34.4in</td>
<td>25.8in</td>
<td>17.7in</td>
<td>0.86in</td>
<td>62mil</td>
</tr>
</tbody>
</table>

Figure 4.25: (a) Slot antenna un-tuned imag / real impedance at $Z_S$ in Fig. 4.24 and (b) Associated Smith Chart from 200 to 400MHz.
The slot antenna un-tuned simulated impedance at $Z_s$ is shown in Fig. 4.25. Over the 2:1 tuning band (200 to 400MHz), the worst case VSWR is 22:1, and the impedances are well within the matchable range of an impedance tuner. If this antenna were to be realized, the impedance tuner would be integrated on the ground plane of the slot.

The probe design guide, observing the E/H ratio across the tunable band described previously, is used to pick the location and type of probe. Fig. 4.26 shows the complex total $|E|$ and $|H|$ as well as their ratio at 325MHz. The probe will protrude from the ground plane into the slot region of the antenna. Fig. 4.26(a) shows a strong E field component near the middle of the slot and a strong H field component at the ends of the slot in Fig. 4.26(b). This is of course consistent with the fundamental mode field distribution for a slot antenna, the E/H ratio confirms the E and H dominant locations in Fig. 4.26(c). To pick a probe type and location, the E/H ratio is plotted vs frequency in Fig. 4.27. For all frequencies, the edges of the slot remain H dominant as is forced by the boundary condition at the edge of a PEC interface. As the frequency increases from 200 to 400MHz, the strongest E dominant region of the slot continues to move to the left side of the antenna (away from the feed). Beyond 400MHz a higher order mode begins to be excited, hence another E dominant region on the right side of the slot.

Due to the E dominant region discriminated to the left side of the slot, an E field probe location between Pos. 1 and Pos. 2 is chosen, 4.8in from the left side of the slot as depicted in Fig. 4.28. The small monopole is co-polarized with the vertical E field that is excited between the top and bottom of the slot. The probe protrudes from the ground plane 3mm.
Figure 4.26: Simulated loop E and H fields at 325MHz.
Figure 4.27: E/H ratio vs frequency for slot antenna.
Figure 4.28: Radiated power sensor integrated with slot antenna.

Free space simulated probe performance, Peak Realized Gain and $|S - Parameters|$ vs frequency for Fig. 4.28 is shown in Fig. 4.29 from 50MHz to 1GHz. In Fig. 4.29(a) and (b), the peak gain and $|S - Parameters|$ from 200 to 400MHz is plotted for several tuning conditions at $Z_S$. These two plots are used to generate the peak gain vs $|S_{21}|$ in Fig. 4.29(e). For the designed tuning band, the probe and slot $|S_{21}|$ response properly track peak gain. A maximum coupling of $\approx -38dB$ is observed with a 17dB dynamic range. While this amount of coupling is well below the -15dB threshold previously discussed, it may be too low depending on the transmit power. To increase coupling, one would simply increase the length of the probe. The dynamic range observed for this case is low compared to the previous designs. This is because the antenna is not as severely mis-matched across the 2:1 band as was the case with the previous antennas. To gain more insight into the dynamic range a detection diode might need to provide, one would need to study the range of impedance values their matching network would provide.

Fig. 4.29(d) shows the lower bound of operation for the probe, from 50 to 100MHz the probe response does not properly track peak gain, thus the lowest operating
Figure 4.29: Peak Realized Gain and $|S_{21}|$ vs frequency. Probe $|S_{21}|$ tracks peak gain from 125 to 950MHz.
frequency is 125MHz. Fig. 4.29(c) and (f) show that the probe continues to track peak gain up to 950MHz, consequently the probe properly tracks peak gain for a ultra-wide 7.6:1 bandwidth. To understand the band limits for the probe, near field E/H is plotted for 50, 175, 800, and 1000MHz in Fig. 4.30. At 50MHz, the entire slot is H dominant, hence the E field probe does not track peak gain. At 175MHz the probe location is approaching the E/H ratio of 377 where each component is of equal strength. The quasi-static E field changes with gain as the antenna is tuned and thus the probe operates as expected in this region. At 800MHz higher order modes are excited, E field dominant regions of the slot straddle the location of the probe, yet the probe is still located in a 377 E/H ratio spot.

![Figure 4.30: E/H ratio vs frequency for slot antenna, H field dominant probe location at 50MHz and 1GHz.](image)

125
At 1GHz the second higher order mode is excited and the probe is located in an H
dominant location, hence it no longer tracks gain. If the antenna were to be tuned
beyond 950MHz, a second E field probe would be necessary.

4.6 Patch Antenna Probe Design Guide

This section covers the near field probe design for a resonant sized patch antenna.
The patch antenna is depicted in Fig. 4.31 and corresponding dimensions in Table 4.4.
The un-tuned patch antenna impedance is shown in Fig. 4.32 where from 275 to
375MHz the real impedance varies from 0.01 to 52Ω and the imaginary impedance
varies from 44 to -8.6Ω, a worst case VSWR of 5288:1. This patch antenna is resonant
at 375MHz with a narrow operating bandwidth similar to most patch antennas. As
such the real impedance drops from 50Ω to less than 0.1Ω very quickly. As such,
the impedance range of the un-tuned patch antenna is outside a matchable range for
an impedance tuner. Thus the patch will be tuned through reconfiguration on the
aperture.

There are several ways to load a patch antenna for tuning. A rectangular patch
is divided into two sections and connected using varactor diodes in [82] to archive
octave bandwidth tuning. A similar concept is used in [83], alternatively, the varactor
diodes are used to break a larger patch into several different smaller patches providing
frequency tuning, polarization agility, and phase shifting. Tuning is achieved over a
1.68:1 bandwidth. Two varactors placed on the radiating patch edge opposite an
inset feed is used to tune the electrical length of the patch antenna in [84] achieving a
10% tuning range. In [85], 2 varactor diode are loaded at the center of both radiating
ends of a pin-fed rectangular patch to tune its electrical length over a wider 1.66:1 bandwidth.

Figure 4.31: (a) λ/2 375MHz resonant sized probe fed patch antenna with integrated tunable capacitor (b) PCB side view, $C_{Tune}$ varies from 0 to 40pF.

Table 4.4: Fig. 4.31 Patch Antenna Dimensions

<table>
<thead>
<tr>
<th>SubL</th>
<th>PatchL</th>
<th>FeedL</th>
<th>SubH</th>
</tr>
</thead>
<tbody>
<tr>
<td>10in</td>
<td>6.2in</td>
<td>2.3in</td>
<td>250mil</td>
</tr>
</tbody>
</table>

In [86] this concept is extended using 8 varactors spread across the two radiating edges (4 on each edge) of the patch in an effort to reduce system losses and achieve
higher gains. A similar concept is adapted to a slotted co-planar patch antenna where a single varactor on one radiating edge of the patch providing a 9% tunable bandwidth in [87]. The concepts in [84–87] is applied in this work, a single variable capacitor on one radiating edge of the patch tunes from 0 to 40pF providing tuning from 275 to 375MHz (1.36:1 BW), the circuit model is depicted in Fig. 4.31(c) (modified from [86]).

Determining probe type and placement will follow the same design outline previously discussed. Near field E and H plots are shown in Fig. 4.33 at 375MHz, where the patch is resonant, with 0pF capacitor loading. Typical E and H fields are observed in Fig. 4.33(a) and (b). The E/H ratio is in Fig. 4.33(c) showing E dominant locations on the top and bottom of the patch radiating edges. This antenna will not be tuned with an impedance tuner, rather the integrated variable capacitor has the effect of changing the patch’s electrical length. As such E/H must be observed vs frequency.
and capacitor tuning state. Fig 4.34 shows the E/H ratio from 275 to 375MHz, over the tunable range of the patch, from 0 to 35pF. The patch is resonant for the following conditions: 275MHz at 35pF; 325MHz at 20pF; and 375MHz at 0pF.

While the resonant condition of the unloaded patch (0pF) in Fig. 4.33 depicts a symmetric field distribution for the top and bottom halves of the patch, the capacitor tuned patch does not as capacitor loading only exists on the top half of the patch.

Figure 4.33: Simulated loop E and H fields at 375MHz at 0pF capacitor loading.
Hence, E or H dominant locations change depending on the tuning state. Through observation of Fig 4.34, it is seen that either side of the non-radiating edges of the patch is usually H dominant, except at 275MHz/0pF and 375MHz at 35pF. The top radiating edge is E or H dominant depending on the tuning state. However, the bottom radiating edge is consistently E dominant and thus an E field sensing small monopole should be used to detect a tuned state.

Two E field sensing probes are proposed in Fig. 4.35. The first in Fig. 4.35(a) is a top loaded monopole 11.75mm away from the bottom radiating edge of the patch. The rectangular strip on the top of the probe provides easy design freedom to ensure the proper amount of coupling is achieved between the antenna and sensor. An alternative sensor is shown in Fig. 4.35(b), here a small probe is placed directly under the center of the bottom radiating edge of the patch. However, the probe only protrudes 150mil into the 250mil thick substrate. This probe takes up less space than that of Fig. 4.35(a) and is directly integrated in the same area of the patch itself. This probe would be advantageous if the patch ground plane were very small. However, it might be difficult to ensure exact probe length in fabrication, hence the top loaded monopole design in Fig. 4.35(a) may be easier to realize.

Free space simulated probe performance, Peak Realized Gain and $|S-Parameters|$ vs frequency for the probe in Fig. 4.35(a) is shown in Fig. 4.36 from 150 to 475MHz. In Fig. 4.36(a) and (b) it is seen that the gain and coupling peak changes as the capacitor is tuned from 0 to 40pF. It is observed that $|S_{11}|$ is not -10dB or better as the antenna is tuned, and the peak gain suffers. Hence, there is room for design improvements, regardless, this performance is suitable to study the effectiveness of the probe design. Fig. 4.36(e) shows that the probe $|S_{21}|$ response properly tracks the
Figure 4.34: E/H ratio vs frequency and tuning C value for patch antenna. Patch is tuned for the following conditions: 275MHz at C=35pF; 325MHz at C=20pF; and 375MHz at 0pF.
Figure 4.35: (a) Radiated power sensor integrated 11.75mm away from the bottom radiating edge and (b) alternative 150mil long probe 1.37mm in diameter integrated under the center bottom edge of patch.

peak gain from 275 to 400MHz. Antenna / probe coupling is -28dB, below the -15dB design threshold. Below 275MHz, the probe no longer tracks the gain. Likewise, above 400MHz, the probe does not properly track peak gain. Similar to the previous analysis, the probe is not necessarily located in an E dominant location outside of the 275 to 400MHz band for each tuning state. Fig. 4.37 shows the probe performance from Fig. 4.35(b). Across the band, performance is almost identical to that of the top loaded monopole probe of Fig. 4.35(a). The small pin integrated under the patch also tracks peak gain from 275 to 400MHz. It is observed that rather than -28dB of coupling between the patch and probe, a lower -38dB of coupling is achieved. For a given transmit power, this probe would require a more sensitive detection diode. Of course coupling could be increased by making the probe longer.
Either probe is effective at properly sampling far field radiated power. Results validate the design guide works for the patch antenna just as it did for the other three antenna designs.
Figure 4.36: Fig. 4.35(a) Peak Realized Gain and $|S_{21}|$ vs frequency. Probe $|S_{21}|$ tracks peak gain from 275 to 400MHz.
Figure 4.37: Fig. 4.35(b) Peak Realized Gain and $|S_{21}|$ vs frequency. Probe $|S_{21}|$ tracks peak gain from 275 to 400MHz.
Chapter 5: DEMONSTRATION OF CLOSED-LOOP RECONFIGURABLE ANTENNA WITH FAR-FIELD SENSOR

This chapter will implement the concepts developed thus far to create a closed loop tuning system. A double stubbed mechanical tuner will be used to impedance match the meandered monopole from Section 4.3. The associated E field probe is integrated and connected to a detection diode so the system micro-controller can sense a matched condition. The micro-controller uses both a brute force tuning algorithm and an iterative algorithm discussed in Section 3.5 to tune the antenna.

This chapter discusses the mechanical tuner, presents a double stub tuner design, and demonstrates a full working system through a fabricated prototype.

5.1 Mechanical VHF Impedance Tuning

Thus far tuning with varactor diodes has been primarily discussed. Other technologies can be used such as PIN diodes, FET switches, and while MEMS are popular, they are costly and not commercially available at VHF. The major drawback of such devices for agile impedance matching over an appreciable bandwidth, are unavoidable losses and lower tuning ratios. To remedy this, it is worth examining the use of mechanical (ME) tuning as opposed to solid state tuning.
Mechanical tuning has its own positive and negative trade off’s when compared to solid state tunable components. The positive aspects include:

- Very low loss
- ME tuning does not suffer from linearity problems
- High RF transmit power can be used once a tuning state is found
- DC control power can be turned off after tuning, thus saving battery power

Some negative aspects include:

- Larger size to implement
- Slower tuning speeds
- Increased DC power consumption during tuning

Until new breakthroughs in material science and the semi-conductor industry occurs, mechanical tuning can yield much more aggressive impedance / frequency tuning agility with lower losses. With the existence of micro-motors and other small movable mechanisms, mechanical tuning certainly is a viable option, and is well suited for a leave behind system that only needs to be tuned once or twice after being left in a static environment. The mechanical tuner concept is depicted in Fig. 5.1 where a shorting arm is mechanically rotated to different positions on a curved grounded co-planar waveguide (GCPW) transmission line.

A nylon arm ≈ 16 x 6 x 3mm in size, fabricated with an SLS 3D printer, is attached to the shaft of the stepper motor as shown in Fig. 5.1(b)(c). A conductive fabric is
Figure 5.1: Mechanical tuner concept with stepper motor (a) GCPW concept (b) GCPW implementation prototype (c) SLS 3D printed nylon arm with conductive fabric (dimensions in mm), (d) Side view of GCPW shorted gaps (e) Front view of shorting arm, fabric hangs and compresses against PCB.

Figure 5.2: S-Parameter comparison of shorting arm and copper tape when the main line of Fig. 5.1(b) is shorted.
fixed to the nylon arm and acts as the shorting mechanism. A commercial VID29-05P stepper motor is used, 7mm thick x 29mm in diameter operating at 2.5V / 20mA (70mW), also shown in Fig. 5.1(b). To assess the usefulness of the shorting arm, its performance is compared to that of copper tape. The shorting arm in Fig. 5.1(b) is rotated to short the three GCPW conductors on the main line. An $|S_{21}|$ and $|S_{11}|$ measurement is recorded, the shorting arm is removed and replaced with 3mm wide copper tape in the same location, and S-Param. recorded again. Measurement results are in Fig. 5.2(a)(b). The shorting arm is quite comparable to copper tape with a 1dB difference in the $S_{21}$ measurement and less than a 0.1dB difference in the $S_{11}$ measurement.

5.2 Small Monopole with Double Stub Tuner

Fig. 5.3 shows the electrically small meandered monopole presented in Section 4.3, $\lambda/11.3$ at 180MHz (the lowest measured tunable frequency), with schematic of components for integration to demonstrate the performance of the closed loop mechanical impedance tuner. Integrated components include: A double stub tuner to tune the small wire antenna, micro-controller for closed loop tuning, and a radiated power sensor. The small wire antenna is simulated in Ansys HFSS where the un-tuned impedance vs frequency (Fig. 5.4) is used to design the double stub matching network depicted in Fig. 5.3(b) where transmission line lengths are given at 300MHz.
Figure 5.3: (a) Small antenna with tuning architecture and (b) double stub tuner schematic.

Figure 5.4: ADS circuit model used to design the double stub tuner. The un-tuned monopole impedance is imported as a load and used to optimize the design.

The un-tuned impedance of the antenna is imported to ADS as a load in a two port network to aid in optimizing the double stub tuner design. The ADS simulated tuned / un-tuned results of the antenna from 200 to 400MHz is shown in Fig. 5.5. Up to a 12dB improvement is observed at the low frequencies, the entire 2:1 band
benefits from tuning. However, this data only shows the tuner’s performance for the impedance curve in Fig. 5.4. To gain a better understanding of its performance for all impedance values, transducer gain for the whole Smith Chart is characterized vs frequency in Fig. 5.6 after tuning in software. From 200 to 300MHz fairly uniform Smith Chart coverage is attained to 3dB of loss. Beyond 300MHz coverage diminishes and the forbidden matching region inherent to double stub tuners is more pronounced. An “X” marker is added to each chart showing the small antenna’s impedance from Fig. 5.4 at each particular frequency. Previously the concept of using a variable capacitor on a fixed length shorted stub was used to gain variability. Alternatively, this design uses a fixed value capacitor and varies the stub length to provide the tunable capacitance and inductance. As such, the lumped capacitor is imperative to the matching performance. Fig. 5.7 shows the simulated Smith Chart coverage for the double stub tuner without the lumped capacitors. As expected, the matching coverage is greatly diminished.

Figure 5.5: ADS simulated $S_{21}$ of the tuned vs un-tuned response of the small antenna Fig. 5.3 impedance.
Figure 5.6: Transducer gain characterization of Fig. 5.3(b) double stub tuner vs frequency. “X” marker denotes Fig. 5.4 monopole un-tuned response at the given frequency.

Figure 5.7: Transducer gain characterization of Fig. 5.3(b) double stub tuner without lumped capacitors vs frequency. “X” marker denotes Fig. 5.4 monopole un-tuned response at the given frequency.
5.3 Fabrication and Measurements

The small monopole and associated components from Fig. 5.3(a) are fabricated as shown in Fig. 5.8. The double stub tuner from Fig. 5.3 is implemented in GCPW with the radial tuning stubs and lumped capacitors on the top side of the PCB. The bottom PCB shows the micro-controller, stepper motors, and near field probe necessary for closed loop tuning. A via wall is placed around the detection probe on the ground plane to shield it from any E fields that might exist due to standing waves in the tuner. A small Arduino Nano micro-controller is fastened to the board to control the stepper motors and to receive input from the probe. A Linear Technology LTC5507 detector is used with the probe to provide an analog voltage to the micro-controller.

Measured results are compared to simulations in Fig. 5.9. Although the tuner was designed to work from 200 - 400MHz, measurements show that tuning was achievable down to 180MHz where the antenna and ground plane (3” x 5”) is $\lambda/11.3$ and the space for the antenna alone (1.5” x 3”) is $\lambda/20$. An 11dB gain improvement is observed at the lower frequencies and a 15dB improvement at 450MHz. Across the entire operating band above 180MHz it is advantageous to use the tuner. The measured tuning taper is quite consistent with the simulated circuit model within 2dB which is good agreement considering the circuit simulator was not full wave. Measurements were conducted with a 5 Watt transmitter through the small monopole antenna. The micro-controller used both a brute force and single-step iterative tuning algorithm (see Section 3.5). One full rotation on each radial stub was discretized into 50 steps. Tuning time for the brute force algorithm to test 2,500 tuning states took 9min. Alternatively, the single-step iterative algorithm took a maximum of 3.5min to find the
Figure 5.8: Fabricated monopole antenna with integrated stepper motors, near field probe, and micro-controller.
best solution. It is noted that none of the tuned realized gain curves reach the simulated directivity. Although the losses for this tuner are quite low, the impedances that need to be tuned are quite extreme with a worst case VSWR of 750:1, the tuner is not able to match them to a VSWR 2:1 or better. Rather, the tuner was designed to provide as much benefit as possible across the band. If a narrower band of frequencies from 250 to 350MHz were to be matched, the double stub tuner could be re-designed to do so with higher tunable gain. This illustrates the difficulty in achieving simultaneous impedance and frequency tuning agility with very low radiation resistance and high reactance. Regardless, the measured data illustrates the benefit of using the tuner, and validates the closed loop tuning topology utilizing the near field probe for feedback, across the 180 to 400MHz (2.2:1) band.
Chapter 6: CONCLUSION AND FUTURE WORK

Adaptive antenna tuning systems are of growing interest as the world of mobile electronics continues to grow. More radios and antennas are being integrated into small devices limiting the space for the antenna. As such, coupling effects on electrically small antennas become an issue requiring adaptive impedance tuning. Thus, there is great interest in antenna systems that can operate over a wide range of frequencies in a multitude of environments placing difficult demands on tuning topologies. Being a system level solution, adaptive tuners consist of the following: Impedance tuner, feedback sensing of a tuned state, micro-controller with tuning algorithm, and the antenna. This dissertation focused on providing novel improvements to adaptive tuning topologies from a system level perspective addressing impedance tuners, sensing techniques, and how they apply to different antennas. While tuning algorithms are of importance, previous work has already provided many robust solutions. The key contributions of this dissertation are:

- New demonstration of how losses are manifested in matching networks is presented. Specifically losses are present due to lumped element resistance and ohmic/conduction losses from the PCB. These component and material losses are made worse through inter-stage impedance mismatches and reflections resulting in standing waves what repeatedly dissipate energy as heat through the
lossy components. Finally, circuit resonances and radiation contribute losses. With said understanding, the designer can take necessary steps to minimize the impact of losses in the system to improve overall efficiency.

- With full consideration of circuit loss mechanisms, a new directional coupler based tuning topology is presented to ensure transducer gain is maximized and validated in simulation / measurements. Traditionally, $|S_{11}|$ minimizing topologies are used, as demonstrated in this work, such a topology can maximize the tuners losses, especially for multi-stage impedance tuners. Hence the importance of the alternative topologies presented in this dissertation. With one directional coupler and two detection diodes placed between the impedance tuner and antenna (as opposed to placement between the source and impedance tuner), maximizing power out from the tuner and minimizing reflection from the antenna effectively maximizes the power delivered to the antenna.

- By exploiting circuit loss mechanisms, a novel VHF impedance synthesizer to characterize impedance tuners for load pull measurements is introduced. By using lossy components in a triple stub configuration with three termination types for the two port network (50Ω, short, and open), VSWR 30:1 impedance synthesize is achieved from 125 to 300MHz (2.4:1 BW). The tuner is small (1.75x17.5in) and low cost, providing robust impedance synthesize at VHF where commercially available synthesizers are lacking.

- A novel method of sensing a tuned state through the use of a near field probe sampling far field radiated power for closed loop tuning is presented. Observing the ratio of $E/H$ in the near field of the antenna is necessary to determine which
dominant field to probe. An associated E or H field sensor is chosen and placed appropriately so as to be co-polarized with the probed field. The probe was shown to effectively track far field radiated power as the antenna is tuned for several classes of antenna.

- Mechanical means of tuning is applied to create an adaptive closed loop tuning system for an electrically small monopole antenna. The integrated near field probe aids in tuning the small antenna over a 2.2:1 bandwidth (180 to 400MHz) where at the lowest tunable frequency the antenna is $\lambda/11.3$ in size. The fabricated prototype validates the low loss mechanical tuner design as well as the novel near field sensor for feedback.

Chapter 2 gives an overview of obtaining reconfiguration for tunable circuits (lumped networks, $\pi$ and T networks, stubbed tuning, and hybrid lumped / transmission line solutions). Impedance agility for single, double, and triple stubbed tuners are investigated. Design concepts and suggestions are made in studying the minimal number of realizable variable parameters needed for stubbed tuners. Finally, Chapter 2 concludes with an overview of realizable component limitations and considerations including DC bias current / voltage, linearity, tuning range, loss, and power handling as it applies to several technologies (MEMS, diodes, etc...).

An understanding of tuner losses in realizable circuits is discussed in Chapter 3. While circuit losses may seem obvious, they are manifested in 3 ways: First, lumped element losses due to $R_S$, ohmic and conduction losses from the PCB (loss as circuit losses approach the radiation resistance). Second, losses due to inter-stage impedance mismatches and reflections. Static component / ohmic / conduction losses are amplified as energy is reflected through the matching network. Third, circuit resonances
and radiation losses. With a full understanding of loss sources, the designer can take steps to minimize said losses as a matching network is developed. $|S_{11}|$ minimizing tuning topologies are typically used in adaptive tuning systems. Such topologies can maximize tuner losses as the topology does not “know” if the energy was dumped into the losses of the matching network or radiated into the far field. Alternative directional coupler based topologies are suggested to ensure the transducer gain is maximized for a given matching network. Concepts are validated through measured prototypes. Finally, Chapter 3 presents a novel low-cost VHF impedance synthesizer to characterize impedance tuners in load pull measurements. Losses in the synthesizer are exploited with three different terminations (open, short, 50Ω loads) to produce impedance synthesis over a 2.4:1 bandwidth bounding VSWR 30:1 circles on the Smith Chart in a compact 1.75x1.75in package at VHF.

In Chapter 4 an alternative novel means of sensing a matched state for adaptive tuners is presented by using near field integrated probes to sample far field radiated power. Such a feedback system has several benefits over directional coupler based topologies. Namely, directional couplers add unavoidable insertion loss, the near field probe avoids these losses. Furthermore, the near field probe requires a single diode for feedback to the micro-controller. A directional coupler based topology would require 2-4 diodes, hence the probe topology is lower cost and simplified. Directional couplers also take up added space, the probe topology is much smaller and provides more design freedom. Lastly, directional coupler have limited coupling and directivity quantities, the probe has no directivity requirement and broad design flexibility for adjusting the coupling coefficient. A general design guide is developed to determine the type of field to probe (E or H), probe placement for integration with the antenna,
and the type of probe to be used (small pin or loop). General rules are developed to aid in design and selection of detection diodes for realizable systems. The design guide is extended to demonstrate the usefulness of the near field probe to track far field gain for a monopole, loop, slot, and patch antennas.

Chapter 5 combines the concepts presented in the dissertation to realize a full closed loop adaptive tuning system for an electrically small monopole. Mechanical means of tuning is revisited to provide a low loss double stub tuner. A near field probe is integrated with the antenna to provide feedback. 2.2:1 bandwidth tuning is demonstrated from 180 to 400MHz where the antenna is $\lambda/11.3$ at the lowest tunable frequency. The fabricated prototype validates the low loss mechanical tuner design as well as the novel near field probe sensor for feedback. Simulations and measurements are within 2dB agreement over the tunable band.

Continued efforts can drastically improve performance and functionality of adaptive impedance tuning systems. Currently, commercially available tunable components do not provide high enough tuning ratios with low enough losses in compact packages. Loss being the biggest hindrance. The biggest contribution to adaptive impedance tuning will be very low loss agile tunable solutions. In the mean time it would be of interest to explore alternative means of tuning. Such solutions would include micro-fluidic means of reconfiguration, which has become a topic of recent interest [88–90] and other tunable materials.

While the near field probe proved useful for detecting far field radiated power, it could be adapted to other antenna feed-back applications. The probe could be extended to MIMO applications where specific phase or weighting conditions must be applied to the antennas. In a dynamic environment a look-up table may not be an
option, as such, the near field probes could be adapted for use. Alternatively, the near field sensors could be extended for polarization reconfigurable antenna systems used in dynamic environments. One application might be sensing a properly generated LHCP or RHCP signal while the antenna is being tuned as the environment changes.

The focus of this dissertation was for VHF, however, the concepts discussed are applicable to any frequency. Still, there may be specific realizable design constraints at different bands like the mm wavelengths. As such, continued efforts would address specific guidelines based on available materials and components at said frequencies.

Lastly, one important aspect of adaptive tuners are power consumption for battery friendly applications. As such, a novel method of obtaining self-powered tuning through power harvesting is presented, especially as it relates to mechanical tuning. Concepts are outlined in Fig 6.1 (modified from Fig 5.3) and Fig 6.2 (Fig 6.2(a) from [91]).

One of the negative aspects of mechanical tuning is higher power consumption. To remedy this, the novel idea of coupling reflected RF energy from the tuner /
antenna to a power harvesting circuit as presented in Fig. 6.2, then converting this energy to DC and re-charging the system battery. For higher transmit powers i.e. above 1-2 watts, enough power from the reflected energy alone can provide power for the mechanical tuner in Chapter 5. Regardless if a mechanical tuner or solid state solution is implemented, this reflected energy will exist, thus harnessing this “dedicated” power source to power the tuner is a great cost benefit to the system battery. This essentially has the potential to become a self-powered tuner.

The proposed idea would use a full wave Dickson rectifier, and RF switches to switch in a directional coupler while tuning takes place. Once a tuned state is found, the directional coupler would be disconnected as to avoid incurring the insertion loss of the coupler. For mechanical tuners, when a tuned state is found, power is no longer necessary to hold a tuned state unlike solid-state solutions. As such, power can be turned off to the mechanical tuner and the antenna would continue to transmit.

Depending on the antenna input impedance and range of the impedance tuner, a perfect match may not be found. If this occurred, the directional coupler could
continue to be attached to the system harvesting any reflected energy and charging the system battery. Again, this reflected power would be wasted anyways, harvesting it would help extend battery life.
Bibliography


