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Abstract

Electromagnetic-based cancer ablation represents an innovative approach to selectively target and destroy cancerous tissues while minimizing damage to healthy cells. One promising mechanism under investigation involves using electric fields to induce localized heating within the targeted tissue. A key challenge in such systems is the ability to deliver energy in a controlled and localized manner, particularly as the electromagnetic properties of the target material vary with frequency and composition. This thesis investigates a tunable matching network and parallel plate structure designed to enable efficient energy transfer to materials under test. This method allows localized electric fields to be induced within the material, facilitating controlled heating.

In an effort to maximize energy transfer to a material under test, a tunable impedance matching network design is a key focus of this work. The impedance matching network is characterized by its ability to dynamically adapt to varying complex impedances and frequency-dependent material properties, such as dielectric permittivity, across a frequency range of 20-100 MHz. Commercially available systems are largely limited to fixed-frequency designs, unable to accommodate changing material impedances or optimize energy transfer, making them unsuitable for applications requiring such adaptability. This work addresses these challenges inherent to systems involving changing material properties. This work addresses these limitations by presenting a tunable, system capable of matching to evolving load conditions, thereby enabling efficient energy transfer in scenarios where existing commercially available systems fall short.

Complementing the matching network, a set of parallel plates provides the physical structure through which energy can be delivered to a material under test. The parallel plate design builds on those used in previous work, refining the geometry to better contain the electric field between the plates. This refinement aims to reduce radiative losses and improve the efficiency of energy transfer to the material under test, while maintaining continuity with the earlier system's design principles. These improvements align with the broader objective of increasing the specificity of energy delivery while also enhancing non-radiating behavior.

Beyond system design, this work uses electromagnetic simulations to investigate the RF interactions of materials like carbon nanotubes (CNTs), which are of particular interest for their potential role in future heating experiments using evolutions of this system. While CNT-based heating is not the primary focus of this thesis, understanding their behavior under RF exposure supports the long-term goal of applying this system to materials with complex electromagnetic properties. Simulations from an electromagnetics perspective provide insights into the RF-induced behaviors of CNTs observed in literature and relevant research. The observed charge distributions and dipole behavior, previously leveraged in other studies to achieve localized heating, help to further contextualize the mechanisms underlying CNT-based RF heating applications. This understanding bridges gaps in existing research, where experimental outcomes have been documented but the mechanisms driving these behaviors remain underexplored.

This work proposes and demonstrates a manually tunable, lumped-element matching network integrated with a refined parallel plate structure for controlled RF energy delivery. The system is validated through both simulation and experimental testing, demonstrating the ability to adapt to varying load conditions and maximize energy delivery across a larger frequency range of 20-100 MHz. This system addresses key limitations observed in fixed-frequency commercially-available designs, contributing a practical step towards scalable

systems for targeted radiofrequency (RF) energy delivery and localized heating. Lastly, this work highlights practical design challenges and tradeoffs, offering insights and recommendations to guide future implementations and evolutions of this work.

Chapter 1

Introduction

1.1 Motivation

Cancer remains one of the most challenging diseases to treat, with current treatments such as chemotherapy and radiation therapy often leading to significant side effects and incomplete eradication of tumors [1]–[4]. As such, there is a growing need for targeted, non-invasive therapies that can selectively destroy cancer cells while minimizing harm to surrounding healthy tissues [1]. This challenge has driven research into novel techniques for cancer ablation, with a potentially promising avenue being the use of carbon nanotubes (CNTs) in conjunction with RF fields.

The unique properties of CNTs, particularly their electrical conductivity and ability to resonate when exposed to RF fields make them ideal candidates for mediating thermal ablation in cancer cells [5]. CNTs have the potential to absorb RF energy selectively, convert it into heat, and thereby induce localized hyperthermia within the tumor, discussed in [2]. This method to induce localized heating ideally works to ablate cancer cells without affecting nearby healthy tissue, offering a less invasive alternative to traditional methods.

Research has already demonstrated the potential of CNTs in cancer therapy. For instance, Gannon et al. in [2] has shown that CNTs can be effectively used in non-invasive RF fields to induce thermal cytotoxicity in malignant cells, with promising results in both

in vitro and in vivo models. Moreover, the work led by Dr. Roger Harrison in [1] and others has highlighted the importance of the suspension medium and molecular interactions in optimizing the RF-induced heating behavior of CNTs. These findings underscore the versatility of CNTs as agents for cancer ablation and the need for further research to refine their application.

To enable continued exploration of CNT-based RF heating and related cancer ablation methods, there is a growing need for systems that can deliver energy in a controlled and adaptable manner. A significant challenge lies in the variability of both the materials being tested and their electromagnetic properties, which may change depending on frequency, composition, or suspension medium. This variability makes energy delivery inconsistent when using fixed-frequency or inflexible systems. Achieving consistent and localized heating requires an RF platform capable of adapting to the properties of the material under test. This motivates the development of a flexible, tunable RF system capable of matching to dynamic loads and enabling efficient energy transfer for localized heating.

The work presented in this thesis takes a step toward addressing this need by developing and evaluating such a system, advancing the practical application of RF-based heating techniques, while also supporting future experimental investigations.

1.2 Previous Work

The ability of CNTs to absorb RF energy and convert it into heat has been previously demonstrated in a variety of experiments such as [1], [2], [6], [7]. The use of CNTs in RF fields was investigated in [2], specifically with a 13.56 MHz RF signal generator, to selectively induce thermal cytotoxicity in malignant cells. Their study in [2] showed that CNTs effectively heated under RF exposure, leading to the destruction of cancer cells in vitro and the complete necrosis of tumors in rabbits following intratumoral CNT injections. This re-

sult demonstrated that CNTs, when exposed to RF fields, could generate sufficient localized heat to destroy cancerous tissues. Similarly, the doctoral thesis by Gabriela Nogueira Furtado e Faria in [1] examined the use of RF fields for targeted heating of CNTs conjugated with proteins in [1] to explore their efficacy in cancer therapy. This work demonstrated the importance of material properties, with suspension mediums and molecular interactions determining the RF responsiveness of CNTs. Suspensions of SWCNTs in sodium dodecyl sulfate (SDS) and sodium deoxycholate (SDC) surfactants were studied in [1], showing that certain surfactants could significantly impact the heating efficiency of CNTs. For example, the research done by Faria in [1] found that CNTs suspended in SDS demonstrated a significantly higher heating rate compared to SDC under RF exposure.

The mechanical properties of CNTs, particularly their vibrational behavior when exposed to electromagnetic fields, have been studied for their potential in biomedical applications as well. The study by Hüttel et al. in [8] demonstrated the use of CNTs as high-quality mechanical resonators, with resonance frequencies reaching up to 350 MHz. This property is particularly relevant to the RF-induced heating mechanism, as CNTs' ability to resonate and absorb electromagnetic energy can be harnessed for localized heating of cancerous tissues. Moreover, Garcia-Sanchez et al. [9] employed a novel scanning force microscopy technique to detect bending-mode vibrations in CNT resonators under RF fields. This mechanical behavior plays a crucial role in the hypothesis that the vibrations of CNTs could contribute to their heating when exposed to RF. In addition to studying the inherent properties of CNTs, research has also focused on optimizing their responsiveness to RF fields for medical applications. The work done by [1], in part, identified key challenges in achieving consistent heating rates, especially when examining the interactions between CNTs and biological tissues. This further underscores the importance of understanding the behavior of CNTs undergoing RF-induced heating. The work shown in [2] and others provided a solid foundation for the *practical* implementation of some CNT-based cancer treatments.

The study by Menéndez et al. in [10] dives deeper into the mechanisms and applications discussed above. RF-induced dielectric heating operates through the interaction of electromagnetic fields with materials. Two primary mechanisms are discussed in [10]: dipolar polarization, where polar molecules attempt to align with the alternating electric field, generating heat through friction, and Maxwell-Wagner polarization, where localized electron movement induces currents and dissipates energy as heat. Unlike materials that rely on magnetic dipole flipping for RF-induced heating, carbon nanotubes do not inherently exhibit strong magnetic properties, as stated in [11]. As such, while [11] explains that CNTs can interact with external magnetic fields and exhibit effects such as magnetoresistance, [7] supports that their response to RF energy is governed by dielectric and conductive mechanisms rather than magnetic interactions. This distinction further supports the emphasis on dielectric heating as the dominant mechanism in CNT-based RF applications.

The study from Menéndez in [10] emphasizes the role of dielectric properties in these mechanisms, particularly the dielectric loss tangent, which determines a material's ability to absorb and convert electromagnetic energy into heat. Materials with moderate dielectric constants and high dielectric loss factors are described in [10] to efficiently couple with RF energy. Further discussion of these two properties is carried out in Section 2.1.1 of this thesis. The authors in [10] also note that for highly conductive carbon materials, exposure may even cause ionization may occur, resulting in microplasmas, as observed in [1]. Several applications of RF-induced heating are discussed in [10] that are relevant to this thesis: non-contact heating, selective material heating, volumetric heating, and the ability to induce heating from within a material body. These properties align closely with the objectives of this thesis and particularly with its application in using RF energy to target materials with evolving dielectric properties or geometries. These heating capabilities support the development of systems designed for specific material responses, as is necessary for experiments involving CNT suspensions. The simulations conducted in this thesis ex-

plore these mechanisms at a fundamental level, helping to better understand the heating behaviors discussed by [10] and in turn, positioning this work as a complementary contribution to the understanding of electromagnetic heating processes.

Vashisth et al. in [6] examine the use of carbon nanomaterials as RF susceptors, emphasizing their ability to generate heat under RF fields in the range of 1–200 MHz. This frequency range is noted in [6] for its safety and versatility compared to microwave systems operating at higher frequencies. The study explores materials such as carbon nanotubes and graphene, which, when embedded into various matrices, enable heating for applications such as composite curing and additive manufacturing. Included is a discussion of dielectric materials, containing permanent or induced dipoles, polarized when placed in an electric field, allowing energy to be stored and dissipated as heat. Notably, the setup in [6, Fig. 1b] resembles the parallel plate configuration used in this thesis, where metallic plates are fed with a signal to generate RF fields for non-contact dielectric heating. One of the observations made in [6], that the penetration depth of RF fields decreases with increasing conductivity, making high-conductivity materials prone to surface heating rather than volumetric heating, supports the importance of being able to use a range of RF frequencies as needed to account for the material properties of the medium, as is explored in this thesis. In the same vein, [6] notes that for fixed frequency and power, the reflection coefficient is strongly influenced by the conductivity of the material under test. This relationship underscores the importance of employing a tunable matching network to reduce reflected power and ensure efficient energy transfer, particularly given the variety of materials with differing conductivities expected to be tested in future work. However, while [6] demonstrates the potential of RF heating with different applicator geometries/methods, it does not present a detailed design or experimental validation of a tunable matching network that can adapt to varying load conditions over multiple frequencies. By contrast, this thesis specifically addresses the design, fabrication, and testing of a multi-frequency matching

network intended for use with parallel plate applicators, enabling efficient power delivery to a material under test.

The study by Sweeney et al. in [7] investigates the use of RF fields as a means of heating polymer nanocomposites containing multiwalled carbon nanotubes (MWCNTs). This work focuses on the dielectric and thermal properties of MWCNT composites and explores various applicator configurations, including parallel plate, direct-contact, and interdigitated fringing field geometries, for effective RF energy transfer. The study contributes significant insights into RF heating mechanisms, the design of applicators, and the practical applications of RF heating in industrial contexts. Relevant to this thesis is the discussion of frequency-dependent heating behavior. As would be expected, the authors observed that optimal heating rates were achieved at frequencies where the impedance of the composite matched the source impedance. The importance of impedance matching networks in such designs is highlighted, although notably left out of the designs presented in the paper.

The role of material properties is emphasized in [7], such as CNT loading levels, in determining the impedance characteristics of the composite, noting that higher CNT loadings behave more like non-reactive resistors, while lower loadings exhibit significant capacitive behavior. These observed properties again support the need for a matching network capable of adapting to varying complex impedances through a tunable matching network. Additionally, their work underscores the importance of frequency tuning to accommodate material properties and maximize energy transfer, supporting the conclusions made throughout the available literature.

The parallel plates, or “applicator” configuration depicted in [7, Fig. 2b] shares similarities with the experimental setup employed in this thesis. However, their design employs circular plates oriented vertically in the XZ plane and separated along the Y axis, contrasting with the set of horizontal rectangular plates used in this thesis, which are oriented in the XY plane and separated along the Z axis. The horizontal orientation of the plates allows the

bottom plate to serve as a platform for the material under test. By contrast, the vertically oriented plates require an additional platform to support the material under test between the plates, introducing the potential for interference or unintended effects on the electric field and energy transfer. Additionally, the materials tested in their work are two-dimensional thin sheets whereas this thesis is designed to accommodate more complex and arbitrary shapes, enabling broader applicability.

A potential limitation in the findings presented in [7, Fig. 2] lies in the inconsistency of experimental conditions across the presented configurations. Each topology was tested under differing power levels, CNT concentrations, and exposure times making direct comparisons difficult. For instance, one direct-contact configuration used 1.0 wt% MWCNT at 100 MHz / 315 W for 4 seconds, a parallel plate configuration used 7.5 wt% MWCNT at 100 MHz / 100 W for 6 seconds, and a third employing an interdigitated geometry used 1.0 wt% MWCNT in PLA at 50 MHz / 315 W for 1 second. Nevertheless, the observed electric field shown in [7, Fig. 3a] can serve as a useful point of reference for future comparisons, as the field distribution generated by their plate geometry shows strong similarities to the field patterns produced by the plates in this thesis.

The study by Altarawneh et al. in [12] investigates the dielectric properties of multi-walled carbon nanotubes (MWCNTs) in nanocomposites, providing insights into the real (ϵ') and imaginary (ϵ'') components of the dielectric constant over a frequency range of 0.5 to 20 GHz. While the specific focus of this work is on MWCNTs, the trends and behaviors observed are still relevant and provide a complementary understanding of CNT behavior under RF excitation. The results presented, particularly the frequency dependence of the dielectric properties and their sensitivity to CNT concentration, offer valuable context for interpreting dielectric heating mechanisms and field interactions with CNTs. The study reveals in [12, Fig. 6] that both the real and imaginary components of the dielectric constant tend to decay exponentially as frequency increases. This behavior highlights the

frequency-dependent nature of the dielectric response, which is a critical consideration for RF systems seeking to maximize energy transfer efficiency. The exponential decay indicates that CNT-based materials may exhibit stronger coupling at lower frequencies, where the dielectric constants are higher, potentially leading to enhanced energy absorption.

In [12, Fig. 6], the loss tangent ($\tan \delta$), defined as the ratio of the imaginary component to the real component, is shown to follow a similar exponential decay with increasing frequency. The findings from [12] suggest that with at least some materials, as frequency increases, the efficiency of dielectric heating decreases, reinforcing frequency selection as a critical design aspect and something that is investigated in this body of work. Additionally, both figures demonstrate a strong dependence of the dielectric properties on the concentration of MWCNTs in the medium. Higher concentrations result in significantly larger real and imaginary dielectric constants, indicating enhanced coupling with the electric field. This trend supports the importance of material composition in tuning the dielectric properties for specific applications. While the exact quantitative values of the CNTs may differ from those in [1], the observed trends provide meaningful context for the understanding of the impact of the material properties of CNTs on their behaviors when exposed to an electric field.

Impedance matching networks are traditionally used in RF systems to ensure maximum power transfer by transforming the load impedance to match the source impedance, thereby minimizing reflections and reducing losses [13]. However, as Balanis discusses in [14], such networks do more than transform the impedance— they also redistribute the voltage and current components of the power within the system as part of the impedance transformation. That is to say, a voltage increase can occur due to the reactive elements, which alternate between storing and releasing power, temporarily accumulating energy and distributing it to the load over multiple cycles, ultimately amplifying the voltage at the load [13].

This principle of voltage magnification is particularly relevant in RF energy delivery applications, such as those leveraging dielectric heating, where strong electric fields must be induced in the load to ensure effective energy deposition, as previously mentioned in this section so far. As electric field strength is directly proportional to the voltage across the load, an increase in voltage results in a corresponding increase in electric field- improving energy delivery. Consequently, the role of impedance matching networks in RF systems extends beyond minimizing reflections to include the critical task of voltage amplification at the load. This voltage step-up effect, a result of impedance transformation and reactive energy storage in a matching network, is a critical design consideration in ensuring effective energy transfer. In the context of work, this principle is leveraged to step up the voltage across the parallel plates when matched, thus helping to facilitate improved energy transfer to the sample.

The significance of this effect is further highlighted in RF heating studies involving carbon-based nanomaterials. For example, [10] and [6] both mention how CNTs subjected to strong enough electric fields can, under some conditions, even lead to plasma formation due to the formation of nanoscale field intensities. In some experimental cases, such behavior has even been observed unintentionally, as in the work by [1], where arcing and visible plasma formation occurred during high-power tests using parallel plates. This highlights that such effects can emerge when voltages exceed the dielectric limits of the surrounding medium. While the system in this thesis is focused on controlled RF energy transfer to a material under test rather than plasma generation, which would likely be a detrimental system behavior if used with biological material, it is important to understand that these applications share foundational principles, particularly in the context of the generation of electric fields and the role of impedance matching. The work by Al Bastami in [15] presents a more deliberate use of these effects through a high-power RF system designed for plasma generation. Unlike this thesis, which focuses on a simpler, experimentally validated match-

ing network for energy delivery across a parallel plate structure, the system in [15] targets automated, real-time control of plasma environments. Their design incorporates advanced control logic, microcontroller feedback loops, and high-power RF amplifiers in a fully enclosed system. However, many of the motivating challenges are aligned with those faced in this thesis: enabling impedance matching over a wide frequency range, handling dynamic load conditions, and doing so within practical size, cost, and hardware limitations.

Where [15] employs a high-speed digital controller to manage an automated, continuously adjusting matching network, this body of work uses a manually tuned, lumped-element network that emphasizes simplicity, reproducibility, and voltage magnification. In this thesis, inductors and capacitors are used and tested for their ability to step up voltage to the load while remaining stable during low- to moderate-power testing. Although no plasma generation was targeted in this work, the system still requires the ability to match dynamically. Voltage amplification in this system is used to boost the electric field across the test sample, which is critical for exploring mechanisms like dielectric heating or CNT-based energy absorption, without exceeding safety or component limits. Rather than pursuing plasma ignition or breakdown in enclosed environments that [15] does, this thesis presents a simpler and more accessible RF system. This approach enables hands-on experimental validation of impedance transformation and voltage step-up effects using a deliberately accessible system in light of the lack of commercially available solutions that would offer adequate tunability.

While a wide range of studies have explored the use of RF heating application, both with and without the use of carbon nanomaterials, and several others have demonstrated sophisticated approaches to high-power, tunable matching networks, there remains a gap in the literature when it comes to experimentally accessible systems that bridge these two areas. The research in this thesis addresses this gap by developing and validating a manually tunable lumped-element matching network specifically designed to deliver energy to a load

through a parallel plate structure. Unlike the automated and high-power systems seen in prior work such as [15], this system is intentionally designed for control and observation, allowing the underlying mechanics of this work to be observed and understood.

1.2.1 Relevant Research

Work led by Dr. Roger Harrison alongside his student, Gabriela Nogueira Furtado e Faria, in [1] serves as a significant source for the present study, particularly in understanding RF-induced heating effects for targeted cancer treatment. Early tests utilized a commercial Comdel generator, operating at 600 W, to assess baseline heating effects in CNT suspensions. These tests demonstrated that the choice of suspension medium significantly influenced heating efficiency, with CNTs suspended in sodium dodecyl sulfate (SDS) showing higher heating rates than those in sodium deoxycholate (SDC). These findings underscore the importance of material properties, specifically the suspension medium, in optimizing for RF-induced heating.

Further exploration within [1], supported by Dr. Caleb Fulton and his student, Josh Short, at the Advanced Radar Research Center (ARRC), expanded upon this work by aiming to test over a select range of frequencies to investigate the frequency-dependent behavior of CNTs as described in [1], [16], and [17]. In these experiments, an electric field was created between horizontal plates, enabling testing across various frequencies. This setup, as briefly described in [1], included an ‘RF device’ that generated the field between parallel plates to observe frequency-dependent heating behavior.

However, [1] leaves significant gaps in technical documentation regarding the RF device and system configuration used in these experiments. For instance, the term RF device does not convey information in this context. While it can be inferred that this device likely included a signal generator feeding into the aforementioned plates, the specific nature and

setup of the RF device remains unclear. This lack of clarity extends to the plates themselves, where a single width measurement is provided in the caption for [1, Fig. 58]. Other parameters such as the length of the plates and the separation distance between the plates are unreported, leaving details of the experimental setup undocumented. Additionally, while a matching network is later mentioned, there is no description of its design, components, or configuration. Without detailed documentation of the matching network, it is very difficult to replicate or fully understand the system's behavior, particularly regarding the interaction between the plates, the field distribution, and any impedance characteristics.

An additional limitation identified was the restricted power output of the RF device used during this phase of testing. The initial experiments using the RF system were conducted at an estimated 15 W output, which, as acknowledged in [1], is likely to have been insufficient to generate measurable differences across frequencies. Later work increased the power supply to 100 W, a level hypothesized in [1] to be sufficient based on observed responses at a single frequency of 13.56 MHz. However, this higher power test was restricted to a single frequency, and thus the results leave room for further assessment across a broader frequency range that this work studies.

1.3 Outline

This thesis is organized into six chapters, each focusing on specific aspects of this body of work.

Chapter 1 introduces the motivation for the work, outlining the potential of RF-based systems and CNTs for energy transfer and localized heating applications. The chapter identifies key challenges and gaps in existing literature, including the lack of commercially available tunable matching networks capable of addressing diverse material properties and the need for further understanding of CNT behavior under RF excitation. The objectives

and scope of the thesis more generally are then presented.

Chapter 2 provides an overview of the fundamental principles of RF systems. It begins with a discussion of dielectric heating and relaxation mechanisms to establish the theoretical basis for energy transfer at varying frequencies. The chapter also explores the interaction of RF energy with biological tissues, emphasizing thermal damage thresholds and regulatory considerations such as SAR limits. Additionally, the chapter introduces key RF concepts used throughout the rest of the work. Concepts such as complex impedance, reflection coefficients, S-parameters, and the Smith Chart are discussed, each characterizing how signals behave in RF systems. These tools provide the foundation for understanding how impedance matching networks are analyzed, tuned, and validated. Collectively, the concepts introduced in this chapter serve as a backbone for simulations, calculations, and measurements performed in later chapters.

Chapter 3 describes the modeling and simulation efforts undertaken to design the RF system. Simulations of CNT behavior under RF excitation are detailed, confirming their dipole-like behavior and localized charge distributions. Then, an evaluation of the parallel plates used previously in the Comdel system was performed, identifying its limitations. Refinements to the plate geometry and material selection are then presented, culminating in a new parallel plate design aimed at improving field intensity between the plates while minimizing radiative behaviors. The chapter then introduces the design of the tunable matching network, including simulations in ADS to assess its impedance-matching capabilities across a wide frequency range. Component selection and testing for the network are discussed, followed by simulations involving a human tissue analog to validate the system's ability to operate with materials under test.

Chapter 4 describes the use of additional subsystems, such as an RF amplifier and power supplies, incorporated to ensure consistent power delivery, while the creation of a power check circuit provided a means of monitoring the power supplies. The chapter concludes

with the integration of these components into a cohesive system, designed to transition from simulation to practical experimentation.

Chapter 5 presents the experimental validation of the RF system. The matching network is evaluated through matching tests, highlighting its ability to adapt to a variety of load conditions at a variety of frequencies, demonstrating consistency with simulation results. Single-frequency tests are conducted to assess system performance at fixed frequencies. The chapter provides detailed descriptions of the experimental setup, including the use of oscilloscopes, directional couplers, and other monitoring tools, as well as the results and their analysis. A series of calculations using S-parameter data experimentally gathered from the realized network and plates is employed to estimate voltage amplification and power transfer, providing further insight into the system's behavior. These calculations are first validated against simulated data, showing strong agreement with results extracted directly from simulation, before being applied to experimentally measured S-parameter data. The results reinforce the network's effectiveness in improving power transfer while highlighting the impact of practical component constraints on performance.

Chapter 6 summarizes the work presented in the thesis, emphasizing its contributions to RF system design and future CNT-based heating applications. The chapter outlines potential areas for improvement, including further refinements to the matching network, exploration of the spacer's role in field containment, and the use of advanced components for greater system efficiency. Directions for future research are identified, such as scaling the system for diverse applications and conducting high-power experiments with CNT suspensions to validate heating capabilities.

In summary, the main product of this thesis is an RF system designed to deliver energy to a material under test. A foundation of relevant RF theory and background is first established. This is followed by the design of a refined parallel plate structure before exploring the development of a tunable impedance matching network. Together, these elements form

a cohesive system to ensure energy transfer to a material under test across a wide range of frequencies. Evaluation is performed through both simulation and experimental testing, with results and conclusions presented in the final chapters.

1.4 Summary

This chapter has established the motivation and foundational context for the research presented in this thesis. The use of CNTs in RF-based cancer ablation has shown significant promise, as highlighted in previous studies, offering the potential for localized heating and targeted destruction of cancerous tissues. However, existing work often lacks detailed exploration into the mechanisms of RF-induced heating and efficient system designs that accommodate dynamic material properties over a wide frequency band. Additionally, the absence of commercially available tunable matching networks designed to meet the specific needs of such systems further underscores the need for this work.

To address these gaps, this thesis introduces an RF system design with a tunable matching network and a refined parallel plate configuration. These advancements aim to enhance energy transfer and provide a flexible framework for future experimental setups involving diverse material properties with a range of frequencies in use. By building on earlier studies, this research contributes to a growing body of work focused on improving the application of RF-based systems, ultimately paving the way for more effective applications in cancer treatment and other fields.

This sets the stage for the technical details and experimental results explored in the subsequent chapters. The next chapter will provide an in-depth background on the mechanisms of RF heating, material interactions, and critical considerations for system design.

Chapter 2

Fundamentals

2.1 Radiofrequency Heating

Radiofrequency (RF) heating occurs when a material absorbs energy from an applied alternating electric field, converting it into heat through molecular and charge-related interactions. A core mechanism responsible for this heating is dielectric heating. Dielectric heating arises from the reorientation of polar molecules within the material as they attempt to align with the oscillating electric field, resulting in energy loss due to molecular friction and relaxation effects. As discussed in Chapter 1, the effectiveness of RF heating depends strongly on the properties of the material, including its permittivity, conductivity, and dielectric loss tangent. These properties, along with the frequency of the applied field and the ability to efficiently deliver energy through impedance matching, influence key outcomes such as penetration depth, power absorption, and localization of heating within the material [6].

This chapter first explores the time-dependent mechanisms of dielectric heating before introducing relevant concepts in RF engineering. These concepts such as impedance matching, power reflection, and S-parameter analysis are critical for quantifying how power is transferred from the source to the material under test. These foundational concepts are essential for interpreting results, understanding system behavior, and reasoning through the

practical design tradeoffs discussed throughout this work- all providing important context for the tunable system presented later in this thesis.

2.1.1 Dielectric Heating

As discussed previously in Section 1.2, dielectric heating is a form of RF heating induced in materials exposed to an alternating electric field [6]. The molecules of a material can experience dipole rotation and try to align themselves with the oscillating field, in doing so producing friction and thus heat [10]. The strength of the interaction between the molecules acting as a dipole and electric field is given by

$$\tau = p \times \vec{E} \quad (\text{Eq. 2.1})$$

where the torque on the dipole, τ ($\text{N} \cdot \text{m}$), results from the cross product of the dipole moment, p ($\text{C} \cdot \text{m}$), and the electric field, $\vec{E}$ (V/m) [18]. This equation describes the torque experienced by a molecular dipole in an electric field. Again from [10], as the dipoles rotate to align with the field, their collisions with surrounding molecules result in energy being dissipated as heat, making this interaction fundamental to the dielectric heating process. The magnitude of Eq. 2.1 can be taken and then simplified to

$$\tau = pE \sin(\theta) \quad (\text{Eq. 2.2})$$

where θ is the angle between the dipole moment and the electric field. This equation shows that the torque is maximized when the dipole moment is perpendicular to the electric field ($\theta = 90^\circ$) and decreases as the dipole aligns with the field ($\theta \rightarrow 0^\circ$). It's worth noting the relationship between charge magnitudes on a dipole and the torque experienced by the

dipole, described by [18] as

$$p = q \cdot d \quad (\text{Eq. 2.3})$$

where the charge, q (C), and the distance between the charges, d (m) factor into the dipole moment. Thus, the larger the charges or their separation, the greater the torque exerted on the dipole, resulting in stronger interactions with the electric field.

Another primary interaction between the field and a material, leading to the power lost in said material, is governed in part by the dielectric constant which can be expressed as

$$\epsilon = \epsilon' - j\epsilon'' \quad (\text{Eq. 2.4})$$

where the dielectric constant, ϵ , is expressed as a complex quantity comprising firstly of the real component, ϵ' , which represents the materials ability to store energy and secondarily, the imaginary component, ϵ'' , which represents the material's energy dissipation [13]. A high ϵ'' indicates that a larger portion of the energy from the electric field is being converted and lost as heat and thus, materials with higher loss factors are better suited for applications aimed at heating under RF exposure.

Furthermore, Eq. 2.5 suggests the angular frequency can amplify these effects, as faster oscillations of the electric field lead to increased dipole interactions and consequently, greater energy dissipation in the form of heat. The power dissipated in part due to this dielectric property of ϵ'' of the material can be described by

$$P = \omega \epsilon'' \epsilon_0 E^2 \quad (\text{Eq. 2.5})$$

where the P is the power dissipation per unit volume (W/m^3), ω is the angular frequency of the field (rad/s), ϵ'' is the dielectric loss factor, and E is the electric field (V/m) [13]. Eq. 2.5 underscores how the dielectric loss factor, which quantifies a material's energy

dissipation capability, plays a critical role in RF heating efficiency.

It's pointed out in [14] that the components of a complex dielectric constant of a material will vary with frequency, with maximums in the imaginary component occurring at some resonant frequencies. From [14], the frequency-dependent form of Eq. 2.4 can be written as

$$\epsilon(\omega) = \epsilon'(\omega) - j\epsilon''(\omega) \quad (\text{Eq. 2.6})$$

The findings in [12] help to further contextualize the role of the dielectric loss factor and frequency in RF heating applications. Their study highlights the frequency-dependent behavior of the imaginary component of the dielectric constant for CNTs, showing an exponential decay for both components of the dielectric constant as frequency increases. As such, these findings suggest reductions in the heating ability of CNTs with increases in frequency. However, Eq. 2.5 illustrates that power dissipation is also proportional to the angular frequency. The interplay between a decreasing dielectric constant and increasing angular frequency suggests that the overall power dissipation depends on striking the right balance between the two.

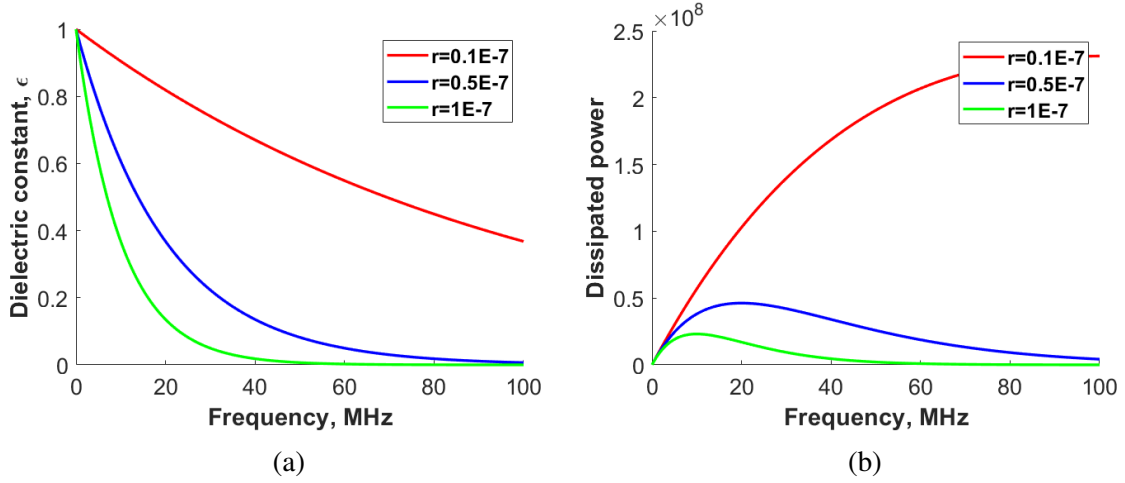


Figure 2.1: (a) Modeled exponential decay for a set of arbitrary dielectric constants vs frequency, each modeled as $\epsilon'' = a(1 - r)^f$ with $a = 1$, for three different values of the rate of decay, r . (b) The corresponding power dissipation, $P = 2\pi f\epsilon''\epsilon_0 E^2$ with $\epsilon_0, E = 1$, as a function of frequency for the three dielectric constant decays. Peaks in power dissipation are observed for each curve, with the peak shifting in frequency corresponding to the rate of decay.

Results from MATLAB simulation above in Fig. 2.1 further demonstrate the interplay between the dielectric loss factor and frequency in determining power dissipation within an RF system. In Fig. 2.1a, the exponential decay of the dielectric loss factor vs frequency is shown for three different decay rates, chosen arbitrarily. The result of these calculations give varying values for the dielectric for a given frequency. Fig. 2.1b illustrates the impact of these decay profiles on power dissipation. It can be seen that each curve exhibits a peak in power dissipation corresponding to a specific frequency. The presence of a peak for each curve highlights the balance between the frequency and the dielectric constant. For each decay profile, the peak represents the optimal frequency where the combined contributions of both of these variables result in maximum power dissipation. These observed peaks in power dissipation also suggest that for any given material or configuration, there exists an optimal frequency range where energy transfer is maximized. This behavior is especially relevant when working with CNT-based systems, as their dielectric properties are highly

sensitive to frequency, as noted in [12]. This requirement to balance the effects of frequency and the dielectric loss factor further justifies the desire expressed in [1] to expand the usable frequency range of the test setup. By extending the operational frequency range, the system gains the flexibility to potentially explore and identify the optimal frequency for maximizing heating efficiency for various material configurations.

Dielectric heating is a critical phenomenon for energy transfer in RF systems, driven by the interaction of electromagnetic fields with the dielectric properties of materials. However, the efficiency of this process is also influenced by how quickly dipoles can reorient in response to changes in the electric field, a behavior governed by dielectric relaxation, discussed in the following section.

2.1.2 Dielectric Relaxation And Debye Model

Dielectric relaxation time, or the dipole relaxation time, is the delayed response of dipoles in response to an applied electric field. As mentioned in Section 2.1.1, the dipoles will tend to align with the field. However, the dipoles do not realign instantaneously to an oscillating electric field due to their inertia and collisions with any surrounding material, leading to losses as heat. As mentioned in [14], the Debye model provides a straightforward means of expressing both the real and imaginary components as functions of frequency, making for a practical understanding and approximation of dielectric behavior. From the Debye model, [14] expands on Eq. 2.6 as

$$\epsilon(\omega) = \epsilon'(\omega) - j\epsilon''(\omega) = \epsilon'_\infty + \frac{\epsilon'_s - \epsilon'_\infty}{1 + j\omega\tau_r} \quad (\text{Eq. 2.7})$$

where the ϵ_s is the static dielectric constant at zero frequency (where $f \rightarrow 0$), ϵ_∞ is the high frequency dielectric constant (where $f \rightarrow \infty$), and τ_r is the relaxation time. From [19], the relaxation time in the above equation is the result of the combination of the physical

properties of the dipole as is described by

$$\tau_r = \left(\frac{4\pi\eta r^3}{k_B T} \right) \quad (\text{Eq. 2.8})$$

where the relaxation time, τ_r , is given by the viscosity of the medium, η (Pa · s), the effective radius of the dipole, r (m), the Boltzmann constant, k_B (1.38×10^{-23} J/K), and the temperature, T (K). Ultimately, the imaginary component of $\epsilon(\omega)$ from Eq. 2.7 can then be isolated as

$$\epsilon''(\omega) = \frac{(\epsilon_s - \epsilon_\infty)\omega \left(\frac{4\pi\eta r^3}{k_B T} \right)}{1 + \left(\omega \left(\frac{4\pi\eta r^3}{k_B T} \right) \right)^2} \quad (\text{Eq. 2.9})$$

Adapted from [14], Eq. 2.9 emerges from the Debye model and describes dipole polarization lag under an oscillating electric field. The imaginary term helps to signify the energy dissipated due to the inability of the dipoles to perfectly follow (polarization lag) the alternating field. This straightforward model illustrates how various materials may exhibit a peak in loss varying frequencies. Based on Eq. 2.9, the maximum value for ϵ'' , where the most loss will theoretically occur, will be found when

$$\omega_r = \frac{1}{\tau_r} \quad (\text{Eq. 2.10})$$

where the Debye frequency, ω_r (rad/s), is the inverse of τ_r . From Eq. 2.10, the theoretical frequency for maximum power absorption, f_r (Hz), can be found to be

$$f_r = \frac{1}{2\pi\tau_r} = \frac{1}{2\pi \left(\frac{4\pi\eta r^3}{k_B T} \right)} \quad (\text{Eq. 2.11})$$

While other, more complex models (e.g., Kramers–Kronig relations and the Cole–Cole model in [20],[21]) can capture multi-distribution relaxation behavior, the Debye model provides a simple approach that is sufficient for illustrating the relationship between a theoretical relaxation time and frequency. This connection again underscores the importance

of understanding how variances in material properties and frequency relate to energy transfer. Furthermore, the frequency influences the depth of field penetration into a material penetration, essential for the effective ablation of cancerous tissues.

The skin depth, δ (m), is the depth at which an EM field's intensity within an exposed material drops to $1/e$ ($\sim 37\%$) of its original intensity at the surface [13]. This measure of penetration is described in [14] as

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \mu_r \sigma}} \quad (\text{Eq. 2.12})$$

where μ_r (H/m) is the relative permeability of the material and μ_0 (H/m) is the permeability of free space, and σ (S/m) is the conductivity of the material. Previously highlighted by Eq. 2.9 and Eq. 2.11, the choice of frequency also plays a key role in the dielectric response of a material. However, optimizing the frequency for maximum penetration through skin depth may conflict with optimizing the dielectric response of the tissue or material. Therefore, in practice, there may be situations where optimizing for skin depth and favorable dielectric response cannot be achieved simultaneously. For example, selecting a low frequency for deeper tissue penetration could lead to sub-optimal energy absorption by the material. Conversely, choosing a frequency that maximizes dielectric loss in order to maximize heating may, in turn, limit the penetration depth.

This design trade-off can be observed in the design of microwave ovens, described in [22], with the goal being to heat food as efficiently as possible while aiming to optimize both energy absorption and penetration depth. Using the value presented by Vollmer for the relaxation time of water molecules (8 ps), Eq. 2.11 and Eq. 2.12 can be used to find the corresponding Debye frequency and skin depth to be 19.89 GHz and 5.05 mm, respectively. Instead of this frequency, microwave ovens typically opt to use 2.45 GHz, resulting in a larger skin depth of around 1.7 cm albeit at the cost of energy absorbed. This design choice represents a deliberate compromise, balancing the need for sufficient

penetration to uniformly heat food while still achieving reasonable energy absorption by water molecules.

Principles discussed in this section such as dielectric heating, dipole relaxation, and how frequency selection plays a role in these two form a foundation for understanding how RF energy interacts with materials to produce the heating effects observed in the literature. The heating effects and energy transfer are intrinsically linked to the frequency in use and could be varied to achieve desired effects induced by RF exposure. These concepts become particularly important when considering the interaction of RF fields with biological tissues, where the effects of RF exposure must be carefully evaluated in light of the material properties, operating frequencies, and field behavior, as explored in the next section.

2.2 RF Effects on Tissue

The interaction between RF energy and biological tissues is primarily seen through induced thermal effects, which has been leveraged in [2] for cancer ablation techniques. The destruction of cancerous tissues comes from the heating described in Section 2.1. However, due to the potential for excessive heating caused by RF exposure, such maximum limits set on the Specific Absorption Rate (SAR) in [23], providing a standard for safe exposure levels and ensuring that healthy tissues remain unaffected. Furthermore, understanding the thermal damage thresholds of both malignant and healthy tissues is critical to optimizing the efficacy of RF treatments while minimizing risk. While falling out of the scope of this work, ensuring a safe environment would require the system to be designed with real-time monitoring and feedback mechanisms to regulate tissue temperature and energy absorption during treatment. The discussion presented in this section is also intended to guide future engineers following up on this work, highlighting the importance of integrating safety considerations and compliance with regulatory standards into the design of RF systems. As

future iterations of this work evolve, the understanding of RF interactions with tissues will remain critical for developing effective and safe systems.

2.2.1 Regulatory Limits on RF Exposure

Based on recommendations from organizations such as IEEE, the National Council on Radiation Protections and Measurements (NCRP), and the American National Standards Institute (ANSI), the Federal Communications Commission (FCC) began setting regulations in 1996 on RF exposure [24]. In order to ensure exposure does not result in adverse effects such as tissue damage, these regulations include limits placed on the specific absorption rate (SAR), a commonly used measure of energy absorption by the body from RF emitting devices [23]. This limit is set as 1.6 W/kg averaged over 1 g of tissue for common RF emitting devices such as mobile phones.

The power density, S (W/m²), can be described by

$$S = \frac{P}{4\pi r^2} \quad (\text{Eq. 2.13})$$

where P (W) is the total transmitted power and r (m) is the distance from the source [25]. This expression from [25] assume an isotropic radiator with uniform radiation in all directions, but can provide a useful approximation for assesing exposure in far-field conditions. The SAR can then be related to the electric field strength through power density as

$$SAR = \frac{\sigma(S\eta)}{\rho} = \frac{\sigma E^2}{\rho} \quad (\text{Eq. 2.14})$$

where η (Ω) is the impedance of free space and ρ (kg/m³) is the density of the material, such as tissue [13]. The SAR can be a useful measure for ensuring that the heating of healthy tissues does not reach an unsafe level. In this work, the operation of the RF system

during testing will be guided by these exposure standards to minimize risks during testing especially when scaling the system to higher power levels.

2.3 Impedance Matching Networks

In virtually all RF systems, an impedance matching network works to ensure maximum power transfer between the source and the load [13]. The primary function of these networks is to minimize reflections by matching the impedance of the source (such as a signal generator or RF amplifier) to that of the load (in this case, the plates and material under study) at a single frequency or across a frequency band of interest [13]. When the impedances are mismatched, a portion of the transmitted power is reflected back to the source, reducing the efficiency of the system [13]. This inefficiency is quantified by parameters such as the reflection coefficient, return loss, and the voltage standing wave ratio (VSWR). By using an impedance matching network, the complex load impedance, which includes both resistive and reactive components, is transformed to match the source impedance [13]. As described in Section 1.2 and throughout this work as a while, this process is important for maximizing absorbed RF energy.

2.3.1 VSWR and Complex Impedance

The VSWR is a key parameter in assessing the efficiency of power transfer in an RF system, defined in [13] as

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (\text{Eq. 2.15})$$

where the reflection coefficient, Γ , represents the fraction of the signal reflected back due to an impedance mismatch. The reflection coefficient itself is

$$\Gamma = \frac{Z_L - Z_S}{Z_L + Z_S} \quad (\text{Eq. 2.16})$$

where Z_L is the complex load impedance and Z_S is the complex source impedance [13]. A complex impedance consists of both a real and imaginary component, given by

$$Z_n = R + jX \quad (\text{Eq. 2.17})$$

where the resistance, R (Ω), is the real part of the impedance and the reactance, X (Ω), is the imaginary part of the impedance [13]. Employing combinations of inductors and capacitors arranged in a variety of configurations can be used to cancel out the reactive components of the load impedance, bringing the overall impedance to a purely resistive value, equal to the source impedance. The component value needed to produce a certain reactance at a given frequency is given as

$$X_L = 2\pi fL \quad (\text{Eq. 2.18})$$

$$X_C = -\frac{1}{2\pi fC} \quad (\text{Eq. 2.19})$$

where X_L (Ω) is the inductive reactance, X_C (Ω) is the capacitive reactance, and f (Hz) is frequency [13]. For a well-matched network, where the matched impedance and source impedance are equal, $|\Gamma| = 0$ and $VSWR = 1$, indicating maximum power transfer with no reflections.

By minimizing reflections and ensuring efficient power delivery, impedance matching should enhance the effectiveness of the goal of improved power delivery as outlined in prior

sections. The Smith Chart, introduced in the following section, serves as a fundamental tool for representing impedance and reflection coefficients, aiding in the design and analysis of matching networks.

2.3.2 S-parameters and The Smith Chart

The Smith Chart is a graphical tool widely used in RF engineering to aid in solving problems related to transmission lines and matching networks. It is particularly useful for visualizing how the complex impedance changes with frequency. The reflections that occur without a proper impedance match are often expressed in terms of S-parameters (scattering parameters), or more specifically, S_{11} [13]. This complex value is a measure of the reflected voltage at port 1 relative to the incident voltage at port 1- thus it is the reflection coefficient at the input [13]. Typically, S_{11} is plotted vs frequency on the Smith Chart. While the Smith Chart is technically a plot of the complex reflection, it is labeled and interpreted in terms of normalized impedance [13] (typically $50\ \Omega$, an industry standard). The Smith Chart provides a visualization of both reflection and impedance matching where the position of plotted data indicates the quality of the match- the center of the chart, where $Z_n = 1 + j0$ and $S_{11} = 0$, indicate a perfect match [13]. However, if the real part of the impedance is greater or less than the normalized system impedance, the point will move to the right or left of center along the horizontal axis, respectively. Alternatively, the imaginary part of the impedance determines the vertical position on the chart— inductive reactance places the point above the horizontal axis while capacitive reactance places it below. The further that the point for S_{11} (and by extension, $|\Gamma|$) moves from the center of the Smith Chart, the larger the reflection [13]. Thus the Smith Chart communicates both the magnitude and type of mismatch, aiding in the design and tuning of matching networks.

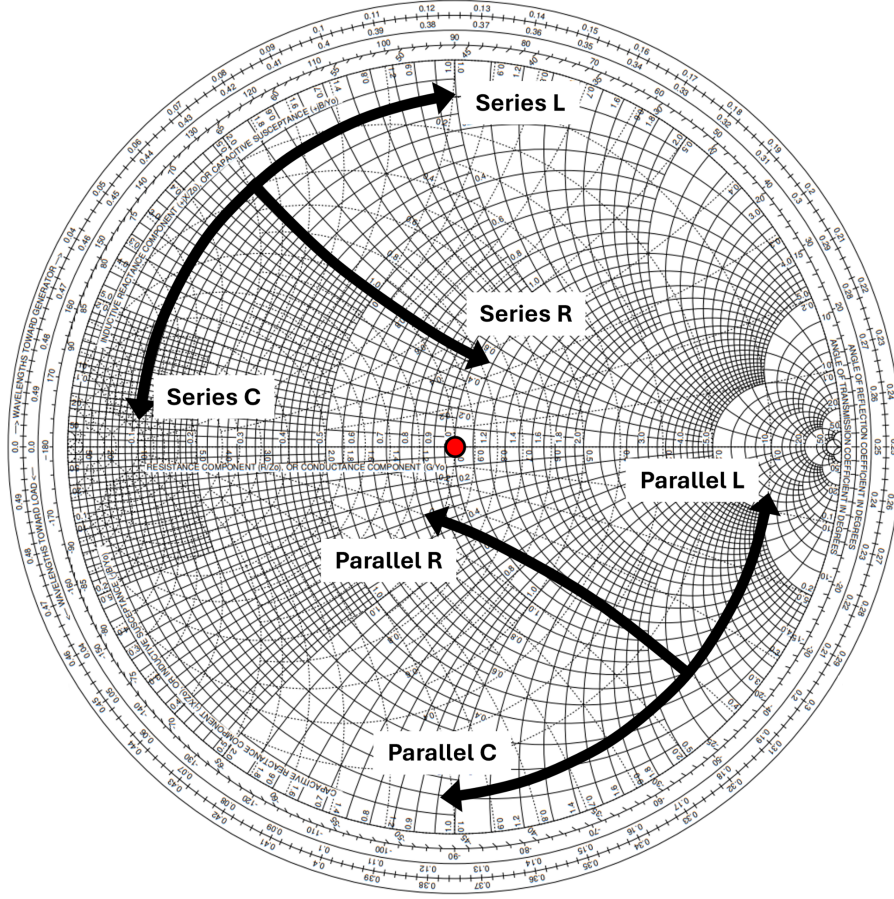


Figure 2.2: The Smith Chart (adapted from [13]) annotated with component effects. Arrows illustrate how normalized impedance or admittance moves on the Smith Chart when reactive or resistive components are added in either series or parallel (shunt) configurations. Inductance denoted by L, capacitance denoted by C, and resistance denoted by R.

In Fig. 2.2, the effects of adding series and parallel components are visualized through directional arrows on the Smith Chart. A series inductor moves the impedance clockwise along arcs of constant resistance and conversely, a series capacitor moves it counterclockwise along arcs of constant resistance. Parallel (shunt) components affect the system admittance. A parallel inductor causes a clockwise movement along arcs of constant conductance on the admittance chart, while a parallel capacitor moves counterclockwise. A series resistor increases the real part of the impedance and moves the point horizontally

along a constant reactance arc. A parallel resistor decreases the real part of impedance and shifts the point horizontally along a constant susceptance arc. Understanding these transformations is essential when using the Smith Chart to visualize and understand impedance matching within this work.

In the context of this work, the full set of S-parameters for a two-port matching network becomes relevant. In addition to S_{11} , a two-port network is also characterized by S_{21} , S_{12} , and S_{22} [13]. S_{21} represents forward transmission, or how much of the signal entering port 1 is seen at port 2 [13]. Conversely, S_{12} represents the reverse transmission from port 2 to port 1 [13]. Lastly, S_{22} describes the reflection seen at port 2 relative to the incident signal at port 2 [13]. Collectively, these parameters can be used to evaluate the amount of reflection, transmission, and loss going in either direction. A visualization of these parameters is provided in Fig. 2.3 below.

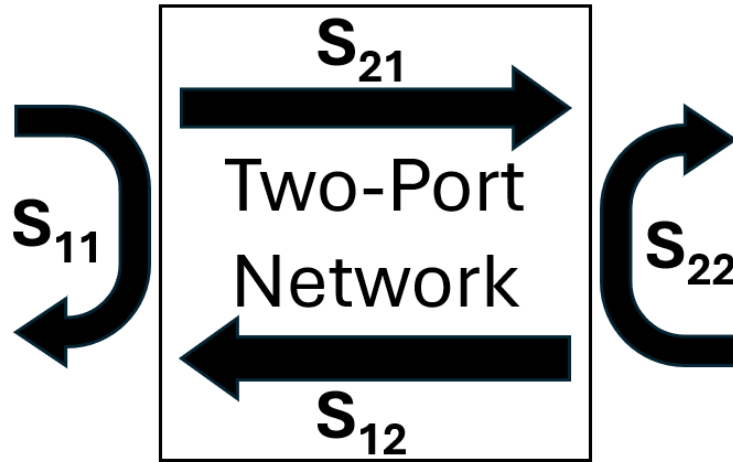


Figure 2.3: Visualization of S-parameters for a two-port network

In the case of a two-port network terminated in a load, the behavior of the system can largely be described by the S-parameters of both the network and the load. Beginning with the terminating load, where $S_{11,load}$ can be used to determine the voltage and power reaching the load when driven directly from the source, without any impedance matching,

as

$$V_{pk,unmatched} = (1 + |S_{11,load}|) \times V_{source} \quad (\text{Eq. 2.20})$$

$$P_{unmatched} = (1 - |S_{11,load}|^2) \times P_{source} \quad (\text{Eq. 2.21})$$

where V_{source} and P_{source} are the voltage and power from the source [13]. Again, Eq. 2.20 and Eq. 2.21 describe the voltage and power when no matching is taking place and thus can provide a point of reference to highlight the impact of a matching network on power transmission to the load.

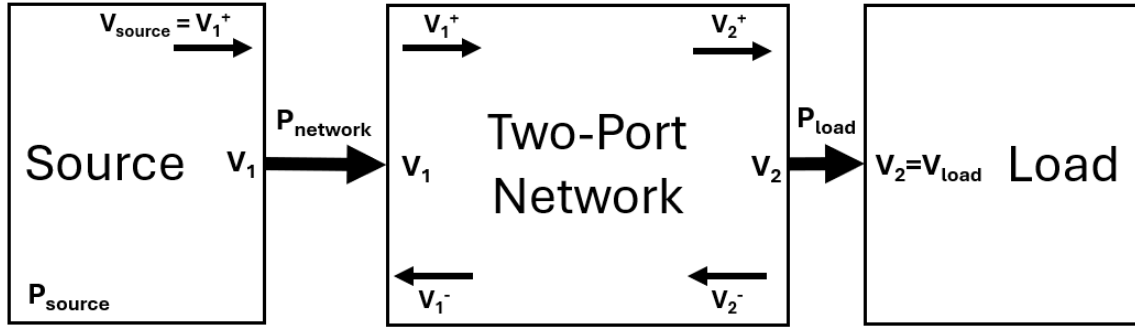


Figure 2.4: Visualization of a two-port network between a source and load, depicting incident and reflected voltage waves. Also shown is the power entering the network, $P_{network}$, and the power entering the load, P_{load} .

In order to find the voltage and power at the load when a matching network is added between the source and the terminated load as pictured above in Fig. 2.4, the following equations from [13] can be used.

$$V_1^- = S_{11,network} V_1^+ + S_{12,network} V_2^+ \quad (\text{Eq. 2.22})$$

$$V_{2-} = S_{21,network} \cdot V_{1+} + S_{22,network} \cdot V_{2+} \quad (\text{Eq. 2.23})$$

where V_1^+ and V_1^- are the forward and reverse (reflected) voltages at port 1 while V_2^+ and V_2^- are the forward and reverse (reflected) voltages at port 2 as pictured in Fig. 2.4.

However, the total voltages at both ports as shown in 2.4 are represented by V_1 and V_2 . The equations above can then be used to derive the voltage across the load as

$$V_{load} = V_2 = \left[\frac{S_{21, network} (1 + S_{11, load})}{1 - S_{11, load} S_{22, network}} \right] V_1^+ \quad (\text{Eq. 2.24})$$

Similarly to Eq. 2.21, the total amount of power entering the system through the network can be found as

$$P_{network} = (1 - |S_{11, system}|^2) \times P_{source} \quad (\text{Eq. 2.25})$$

where $S_{11, system}$ is the measured input response from the network and load connected together. However, this is the power entering the network, not the power to the load which can instead be found by

$$P_{load} = P_{network} \times |S_{21, network}|^2 \times \frac{(1 - |S_{11, load}|^2)}{|1 - S_{22, network} S_{11, load}|^2} \quad (\text{Eq. 2.26})$$

where $P_{network}$ is the amount of power feeding into the network, $|S_{21, network}|^2$ determines the portion of the power continuing through the network to the load, $(1 - |S_{11, load}|^2)$ represents the power going to the load (see again Eq. 2.21), and $\frac{1}{|1 - S_{22, network} S_{11, load}|^2}$ represents the power being reflected at the interface between the network and the load [13]. This section has laid out a process for using S-parameters to find the voltage and power for both an unmatched and matched load impedance. This quantitative approach will later be applied in this work to evaluate the effectiveness of a realized matching network and load.

The concepts discussed in this section, including VSWR, reflection coefficients, impedance transformations, and S-parameters, are critical for understanding the role of impedance matching networks. The Smith Chart provides an intuitive way to visualize impedance and reflection behavior, helping guide the selection of inductive and capacitive components needed to achieve a match. These principles are integral to the simulations

and modeling efforts discussed in the following section, employed to evaluate and refine the performance of not only the matching network but also the design of the parallel plates.

2.4 Numerical Methods

When designing RF systems, simulation tools play a vital role in accurately modeling the complex interactions between electromagnetic fields and materials. Two primary software tools utilized in this research are Ansys' HFSS (High Frequency Structural Simulator) and Keysight's ADS (Advanced Design Suite). These two programs allow for detailed analysis, simulation, and design of RF components and systems.

2.4.1 HFSS

HFSS is a comprehensive 3D electromagnetic simulation software primarily used for high-frequency designs. It employs the Finite Element Method (FEM) to model and solve electromagnetic fields, making it ideal for simulating complex structures such as antennas, waveguides, and microwave circuits. HFSS is typically used for computing key parameters like S-parameters, impedance, radiation patterns, and field distribution based on 3D models. The ability to model electromagnetic fields in and around conductive and dielectric materials is crucial in understanding the interaction of RF fields with tissue material.

2.4.2 ADS

ADS is another simulation tool widely used in RF and microwave circuit design. While HFSS excels in 3D electromagnetic field simulation, ADS is specialized in the design and analysis of RF circuits and systems at the schematic and layout levels. ADS provides advanced features for designing and simulating with matching networks, amplifiers, and filters, making ADS particularly useful within the context of this work.

2.5 Summary

This chapter has provided a review of the theoretical and practical principles underlying RF systems and their interactions with materials, particularly in the context of dielectric heating and relaxation. The discussion emphasized the role of frequency in optimizing energy transfer and heating efficiency, highlighting the importance of understanding material-specific dielectric properties. Also explored was the applications of these principles specific to the potential of CNTs RF-induced heating applications. These principles were further contextualized through a review of RF effects on biological tissues, including thermal damage thresholds and regulatory considerations, which are critical for the safe and effective application of RF energy. Additionally, the concept of impedance matching and use of the Smith Chart was introduced as crucial elements in meeting the objective of enabling power transfer and minimizing losses, providing needed context for the design and implementation of the tunable RF system presented in this thesis. Furthermore, this included an expanded discussion on S-parameters and how they can be used to assess the behavior of a network, quantifying reflection, transmission, and loss.

The insights from this chapter provide the necessary foundation for subsequent investigations through simulation and experimentation. The next chapter will focus on the modeling, design, and simulation of key components, including the matching network and revised parallel plate designs, to achieve the research objectives.

Chapter 3

Design and Trade-offs

This chapter describes both the design and modeling processes used to develop the RF system used in this work, beginning with a set of design objectives that guide both the system layout and component-level choices. This is followed by simulations focused on the electromagnetic behavior of carbon nanotubes (CNTs), helping to establish a foundation for future material testing. The next section reviews past work related to parallel plate structures and presents updates made to improve field confinement and energy delivery. The final sections walk through the design and simulation of a tunable impedance matching network, including component selection. Together, these efforts lead to the practical setup described and measured later in the thesis.

3.1 Design Objectives

The primary objective of this research is to develop an RF system that transfers power to a medium between parallel plates by employing an impedance matching network. Complementing the matching network is an update in the design of a parallel plates used previously, intended to help generate a uniform electric field through a medium. Using simulation tools, this design will help to optimize the geometry and spacing of the plates to deliver a specific electric field while ensuring uniform exposure across the medium.

While the primary focus of this thesis lies in achieving targeted power transfer, an additional objective of this research involves the simulation of CNT behavior when exposed to RF fields. These simulations explore the underlying mechanisms of RF heating at a fundamental level. By modeling these behaviors, this work helps to confirm and support the understanding of heating processes.

Measurements taken from both simulation and experimental setups will be used to characterize the new system and validate its effectiveness in delivering power to the load. The results from both simulation and experimentation will be compared to evaluate how well the physical system aligns with the intended design goals.

3.2 CNT Modeling and Analysis in HFSS

This section delves into the electromagnetic modeling of CNTs using HFSS, particularly focusing on their dipole-like behavior and charge distribution under exposure to RF fields. By leveraging detailed simulations of CNT properties, this analysis addresses gaps in existing literature.

3.2.1 Material Properties and Simulation Setup

The behavior of CNTs in RF fields heavily rely on accurate modeling of their physical and electromagnetic properties. To develop and perform precise simulations, key material properties such as electrical conductivity, thermal conductivity, relative permittivity, and loss tangent were recorded from sources in literature, summarized below in Table 3.1. Additionally, dimensional values specific to the CNTs used in this study, such as length and diameter, were obtained from [1], which provided dimensions of 1500 nm in length and a 25 nm radius.

Material Property	Value	Source
Length	1500 nm	[1]
Radius	25 nm	[1]
Electrical Conductivity	10^5 S/m	[26]
Thermal Conductivity	3000 W/m · K	[27]
Relative Permittivity	10	[28]
Loss Tangent	0.01	[29]
Mass Density	1500 kg/m ³	[30]
Specific Heat	700 J/kg · C	[31]

Table 3.1: Material Properties of CNT(s) used in HFSS Simulations.

A 3D model of a single CNT was constructed in HFSS and placed within a $10\text{ }\mu\text{m}^3$ volume of air, shown in Fig. 3.1. The CNT was oriented along the Z-axis to simulate an idealized alignment for RF excitation.

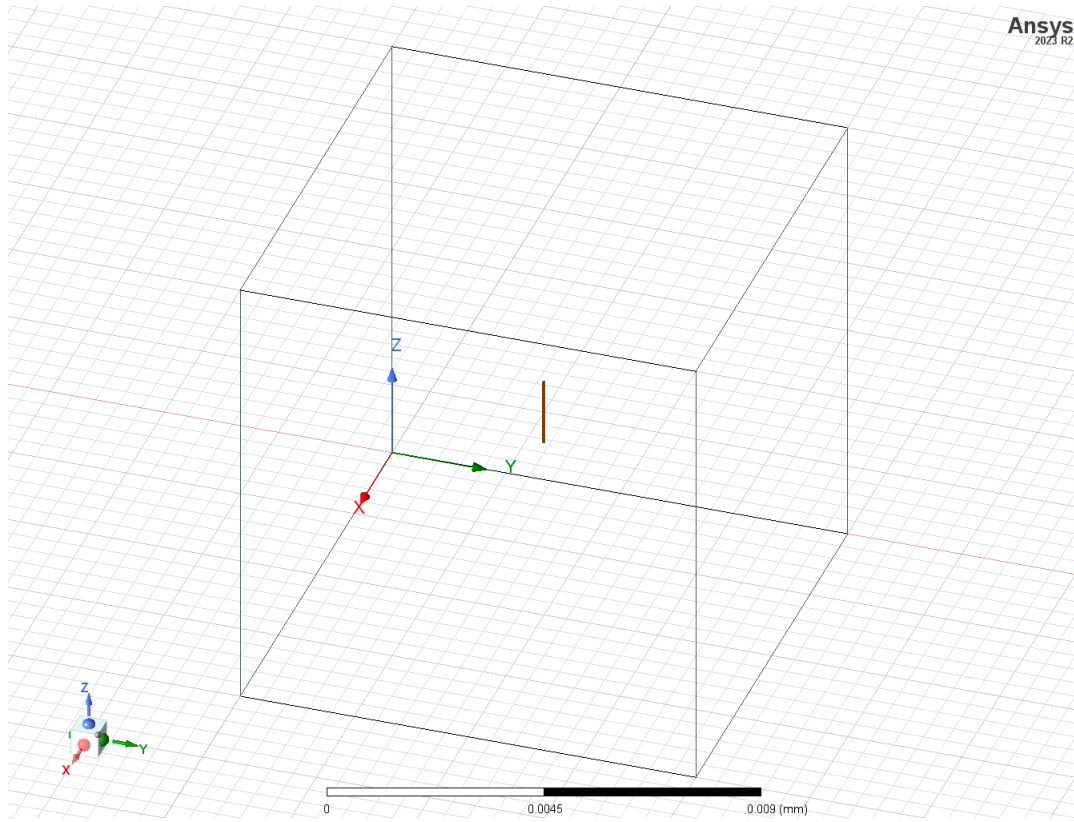


Figure 3.1: 3D model of a single carbon nanotube (CNT) oriented along Z-axis within an air-filled simulation volume in HFSS. This simplified model is used to investigate electromagnetic behavior and field interactions under high-frequency excitation.

To excite the CNT to a maximum extent, two waveports were applied at opposite faces of the volume in the XZ plane shown in Fig. 3.2, and the setup was excited with a signal of 100 MHz at 100 Watts.

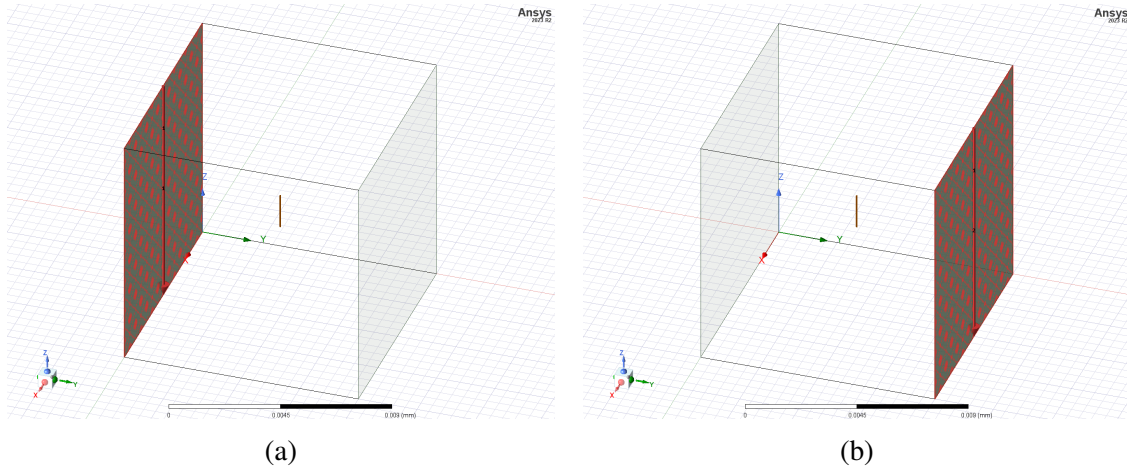


Figure 3.2: HFSS simulation setup showing the waveport excitations used for CNT modeling. (a) Port 1 and (b) Port 2 are placed in the XZ plane on opposite sides of the simulation volume to create an electric field in the air-filled volume and excite the CNT.

To help facilitate accurate field behavior within the volume at such a small scale, PECs (perfect electric conductor) and PMCs (perfect magnetic conductor) were applied as shown in Fig. 3.3. The PMCs were applied to the top and bottom faces of the airbox in the XY plane while the PECs were assigned to the remaining sides of the airbox in the YZ plane. These boundary conditions were applied based on the properties of EM waves to ensure that the quasi-static RF field was confined and uniform within the simulated volume.

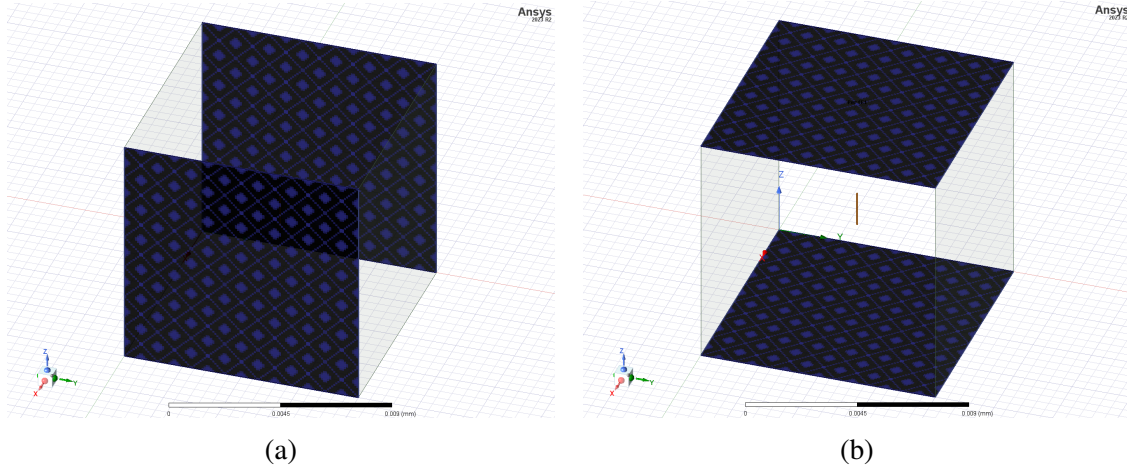


Figure 3.3: HFSS simulation setup showing boundary assignments on air-filled volume containing the CNT. (a) PEC boundaries are applied to restrict the electric field component normal to the boundary, stabilizing wave behavior. (b) PMC boundaries are used to control the magnetic field distribution, promoting a uniform excitation field while emulating a magnetically isolated environment.

The PEC in Fig. 3.3a ensure that the electric field component perpendicular to the boundary (in this case, the Z-axis running along the length of the CNT) is zero. This confines the electric field within the airbox and stabilizes wave behavior, especially along the CNT's orientation. At the same time, the PMC in Fig. 3.3b enforces a magnetic field condition perpendicular to the boundary, creating a uniform background excitation field. This helps to maintain a stable quasi-static field and mirrors real-world behavior where an isolated CNT would have negligible influence from the exterior magnetic field. The combined effect of including these boundaries result in promoting an isolated and focused RF environment. The following section discusses the results from this setup.

3.2.2 Simulation Results

The HFSS simulation results demonstrate significant electric field concentrations at the opposing ends of the CNT, suggesting that it behaves as a dipole within the applied RF field. The field concentrations, as shown below in Fig. 3.4, align with expected dipole

behavior discussed in Section 2.1.

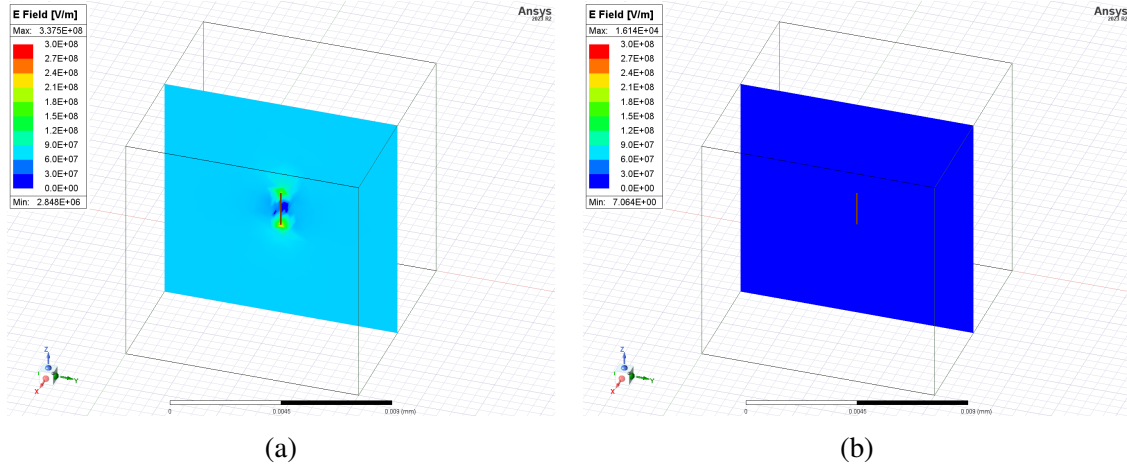


Figure 3.4: Electric field distribution through the CNT and surrounding volume for different excitation phases. (a) Strong electric field concentration at the ends of the CNT, consistent with resonant dipole behavior are seen when phase $\varphi = 0^\circ$. (b) Reduced field intensity shown at opposing ends of the CNT when phase $\varphi = 90^\circ$, highlighting phase-dependent field distribution.

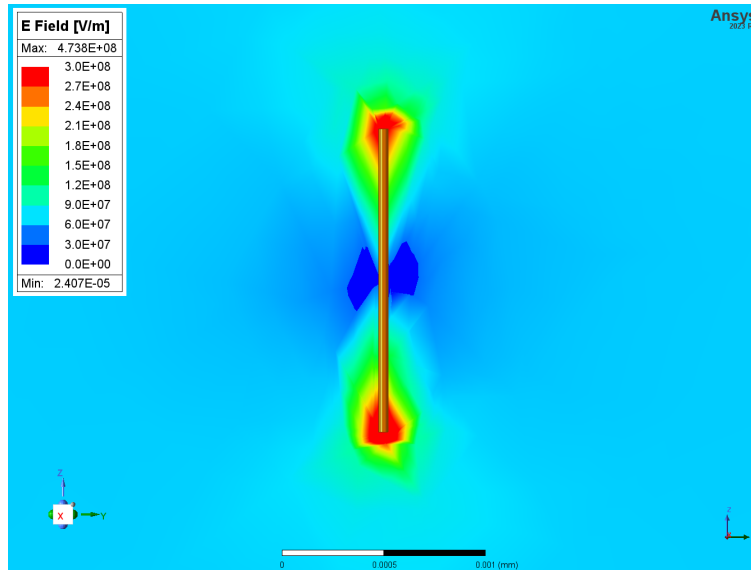


Figure 3.5: Alternative perspective of the electric field distribution for $\varphi = 0^\circ$ from Fig. 3.4a, showing peak field concentrations at the ends of the CNT. This enhanced visualization highlights the dipole-like behavior observed in simulation, where high-intensity electric fields form at the ends of the CNT in response to the RF field.

The field density distribution seen above at $\varphi=0^\circ$ in Fig. 3.4a and $\varphi=90^\circ$ in Fig. 3.4b provides further insight into the dipole-like behavior of the CNT within the RF field. At $\varphi=0^\circ$ in Fig. 3.4a, the field concentration is maximized at the extremes of the CNT, aligning with the behavior of a resonant dipole where peak electric fields form at opposing ends. However, at $\varphi=90^\circ$ in Fig. 3.4b, the field distribution shows reduced concentrations at the ends, indicating a phase-dependent response where peak field concentrations alternate with phase. This phase-based contrast in field intensity underscores the dipole behavior of the CNT, as it cyclically builds and dissipates field concentrations in response to the oscillating RF input. For added precision, the simulation setup within HFSS was refined by increasing the minimum number of adaptive passes, shown in Fig. 3.5.

Looking to the field vectors in Fig. 3.6, where $\varphi=0^\circ$ and $\varphi=180^\circ$, provides additional insight into this behavior. At $\varphi=0^\circ$, the field vectors point in the negative Z direction, whereas at $\varphi=180^\circ$, the reverse can be observed- the field vectors point in the positive Z direction. This reversal of field vectors is typical of dipole responses, where the charge oscillation creates alternating field directions, mirroring the applied RF signal's periodic nature.

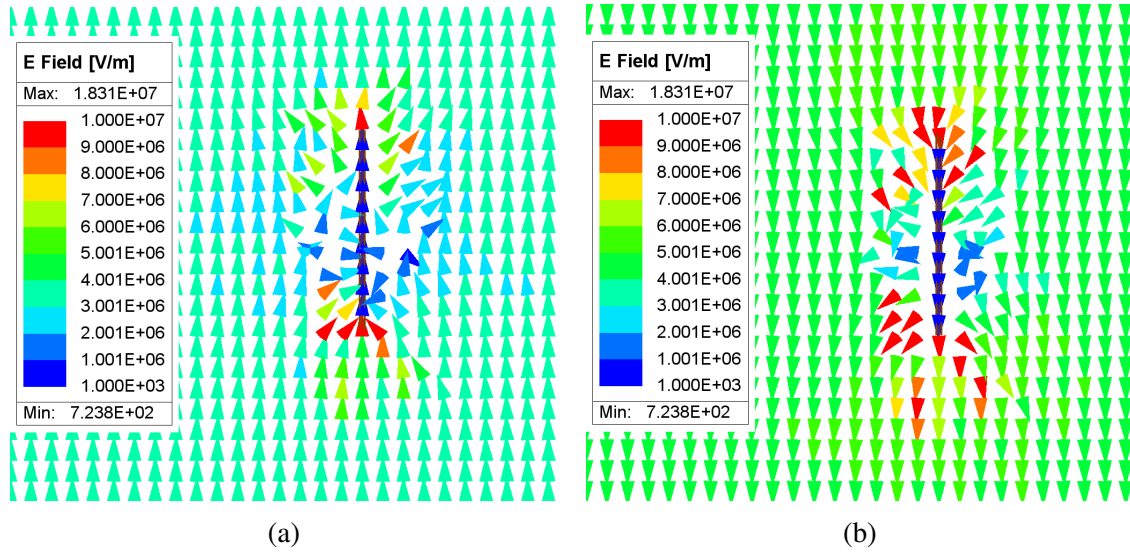


Figure 3.6: Electric field vectors through the CNT and surrounding volume at two phase points. (a) At $\varphi = 0^\circ$, the vectors point predominantly in the positive Z direction. (b) At $\varphi = 180^\circ$, the field vectors reverse and point in the negative Z direction. Result reflects the phase-reversal of the driving RF signal and supports the interpretation of dipole-like charge oscillation across the CNT.

Additionally, the volumetric current density results in Fig. 3.7 indicate that current density is highest at one end of the CNT, providing further confirmation of the dipole-like behavior. The volumetric current is strongest near one end of the CNT, which aligns with the field direction and excitation phase used in the simulation.

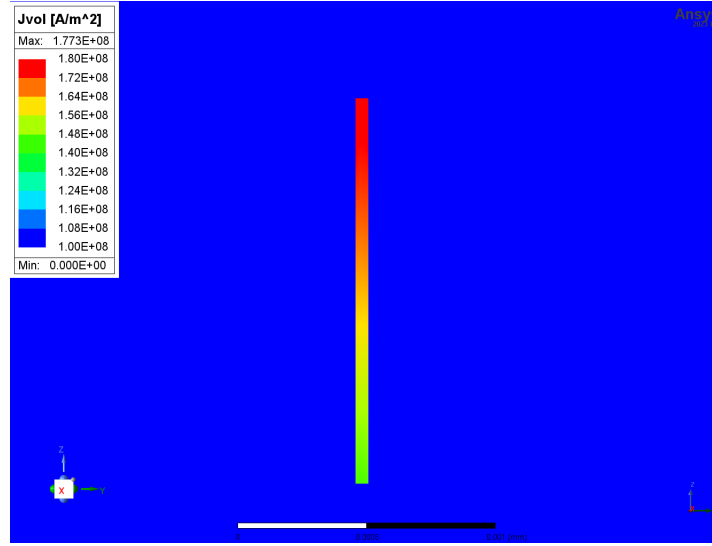


Figure 3.7: Volumetric current density distribution through the CNT under RF excitation. The asymmetric current concentration near one end reflects current behavior along the CNT in response to the applied field.

The current density distribution shown above in Fig. 3.7 further supports the interpretation of dipole-like behavior in the CNT. This localized concentration of current suggests that the CNT responds to the applied RF field by developing an oscillating charge distribution along its length.

The observations presented here in Section 3.2.2 appear to confirm that the CNT behaves similarly to a dipole within the simulated RF environment. This dipole behavior, facilitated by the CNT's conductive and resonant properties, supports potential applications where controlled RF field concentration and enhanced local field intensities are desirable. However, to take full advantage of these properties, it is essential to design a system capable of delivering RF energy with high precision and minimal loss. This necessity underscores the importance of an effective parallel plate design, explored in the following section, playing a crucial role in energy delivery.

3.3 Parallel Plate Simulation and Design

The design and optimization of parallel plates play a critical role in creating controlled and localized electric fields necessary for efficient RF energy transfer. This section outlines the development of a new set of parallel plates, starting from the Comdel plate design utilized in [1]. Limitations observed in the original design, such as inadequate field concentration and broad dispersion, prompted modifications aimed at enhancing field specificity, ultimately reducing unintended exposure. This section examines how simulation insights informed adjustments to the plate geometry. Through simulation and physical testing, these refinements were validated, culminating in a revised plate design that better meets the objectives of localized field production.

3.3.1 Comdel Plate Design

The initial plate design modeled in HFSS was based on the dimensions shown in Fig. 3.2 which were used in the system outlined in [1], made by Comdel. This original design utilized the plates pictured below in Fig. 3.8, with the bottom plate being larger than the top.



Figure 3.8: Original parallel plate structure connected to the Comdel MatchPro CPMX-6000. The RF signal is fed through the housing to the top plate, while the bottom plate is grounded to the chassis, creating an electric field between plates.

In the setup shown above in Fig. 3.8, the parallel plates are connected to the Comdel MatchPro CPMX-6000 system, where the RF signal is fed from the matching network within the housing to the top plate the bottom plate is connected and grounded to the chassis of the device. This configuration creates a potential difference across the plates, in turn generating an electric field between them.

The plates were then modeled in HFSS as shown below in Fig. 3.9 following the measured dimensions from Table 3.2 and surrounded by an air box. A terminal, $50\ \Omega$ waveport was used to excite the top plate while the bottom plate was set as the reference to best replicate the Comdel feed excitation.

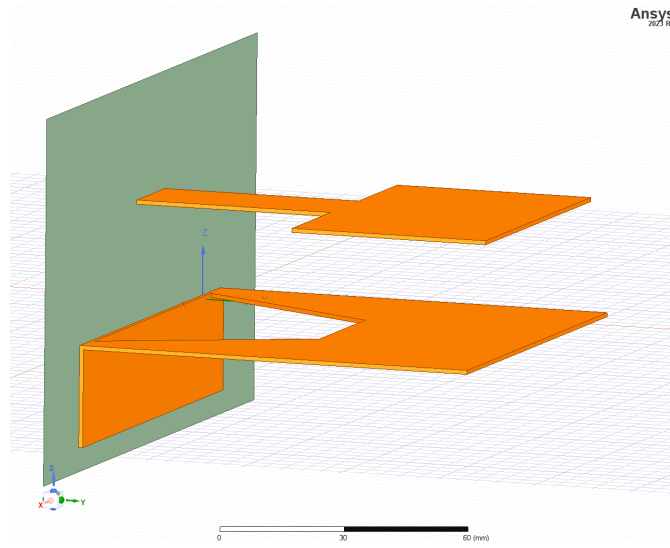


Figure 3.9: HFSS simulation model recreating the original Comdel parallel plate structure. The geometry was recreated based on measured dimensions from the physical system and surrounded by an air volume for accurate field calculation. A terminal, $50\ \Omega$ waveport excitation is shown in green.

Dimension	Value
Length, feed	50 mm
Width, feed	20 mm
Length, top plate	50 mm
Width, top plate	75 mm
Length, bottom plate	75 mm
Width, bottom plate	100 mm
Separation	30 mm

Table 3.2: Dimensions of Comdel plate design.

The total power from the $50\ \Omega$ port, P_{source} , was set to 50 dBm (100 W) based on the estimation from [1], that this power level was predicted to provide sufficient field intensity for observable heating resulting from energy transfer. The RMS voltage at the $50\ \Omega$ port is

determined by

$$V_{rms,source} = \sqrt{P_{source} Z_0} = \sqrt{100 \times 50} = 70.71 \text{ V} \quad (\text{Eq. 3.1})$$

resulting in a total voltage value of $V_{rms,source} = 70.71 \text{ V}$. However, as outlined by Ansys in [32], the voltage used in excitations is expected to be a peak voltage, V_{pk} . Multiplying $V_{rms,source}$ by $\sqrt{2}$ gives $V_{pk,source} = 100 \text{ V}_{pk}$. The source setup window in HFSS is shown below in Fig. 3.10, where the source voltage and power were defined as 100 V_{pk} and 100 W , respectively, to ensure correct excitation at the 50Ω port.

	Source	Type	Magnitude	Unit	Phase	Unit
1	topPlate_T1	Port	100 V		0 deg	

Terminal Excitation Type: ☒ Incident Voltage ☐ Total Voltage

☒ Include Port Post Processing Effects

System power for gain calculations:

☒ Specify System Power:

☐ Use Maximum Available Power

Figure 3.10: Source setup window in HFSS, showing source voltage and power assignments.

Simulations were first conducted at 13.56 MHz, the intended operating frequency of the Comdel plate design. The field intensity depicted in Fig. 3.11 was found to be lower than ideal due to the plate's large surface area. The wider plate configuration not only shows a reduced electric field concentration but also shows a lack in the specificity needed to target smaller, localized tissue regions, discussed below in Section 3.3.2. This contradicts the objective of minimizing field exposure to untargeted areas, which is critical in non-invasive treatments where limiting exposure to healthy tissues is paramount.

In Fig. 3.11 below, the electric field distribution is shown in two different planes to provide a more comprehensive visualization of field behavior between the plates. The first field cut in Fig. 3.11a, taken in the YZ plane, presents a side view of the plates and was extracted along the center of the model's length at $Y = 37.5$ mm. The second field cut in Fig. 3.11b, taken in the XZ plane at $Z = 85$ mm, offers a front-facing perspective looking towards the waveport along the negative Y direction.

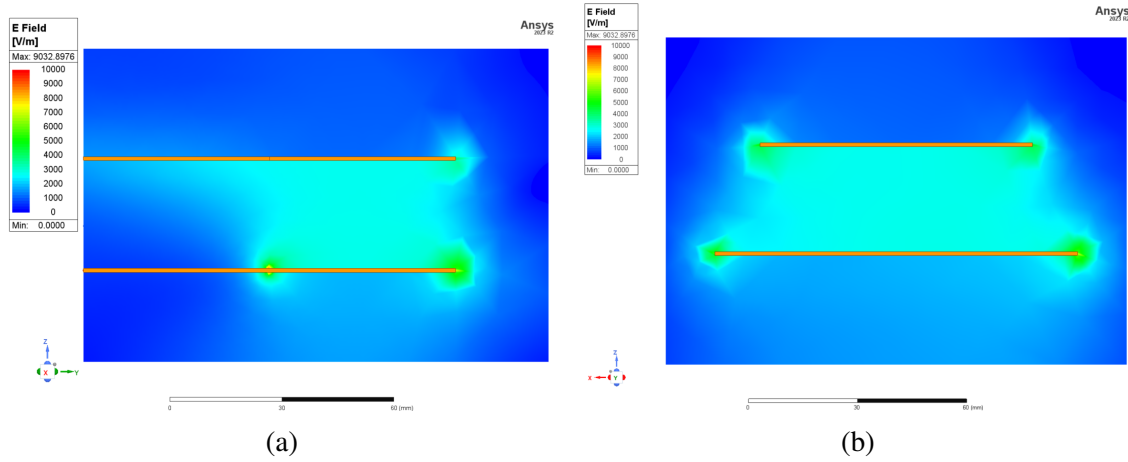


Figure 3.11: Electric field distribution between the original plates when fed with $100 \text{ V}_{\text{pk}}$, 13.56 MHz signal. Cross-sectional view in the YZ plane shown in (a), taken at the midpoint of the plate width. Cross-sectional view in the XZ plane shown in (b), taken at the midpoint of the plate length.

The results shown in Fig. 3.11 illustrate that the initial design did not achieve an ideal focused intensity as well as point out that the wide plates dispersed the field over too large an area, as already mentioned. A key observation from Fig. 3.11a is that the feed into the top plate, which is relatively large in proportion to the top plate itself, exhibits some field-generating behavior in addition to the field contributions from the top plate. As seen in Fig. 3.8, this feed is positioned over a cut-out portion of the bottom plate, meaning that any radiated energy from the feed is not meaningfully contributing to the intended energy transfer between the plates. This decreased flux density over to the lower plate and more general lack of concentration in the E-field likely translates to a reduction in the system's

ability to target specific regions effectively.

These observations present a clear opportunity to refine the plate dimensions and configuration. Such adjustments would aim to enhance field concentration and improve the overall effectiveness of RF energy transfer within the *intended* target area.

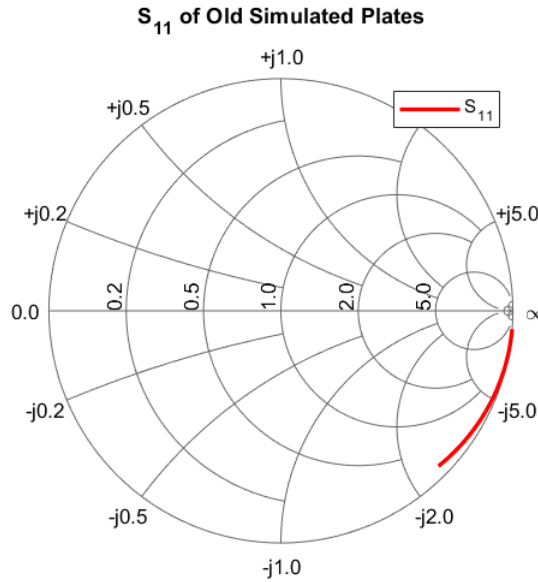


Figure 3.12: Simulated S-parameter response, S_{11} , of the originally used Comdel plates plotted on the Smith Chart. The response indicates a largely capacitive impedance with only a small resistive component, consistent with the plates acting as a weak parallel-plate capacitor in air. This plot serves as a baseline reference for later design changes.

Lastly, the response of the original plates on the Smith Chart can be found above in Fig. 3.12. This plot shows the system's initial response, appearing to be slightly resistive with large capacitive effects. This observation makes sense, seeing as these plates alone are essentially a very weak parallel plate capacitor with air being the dielectric material between. The response shown in Fig. 3.12 provides context further utilized later in Section 3.4.1, acting as a reference point for the improvements made to the design in the following section.

3.3.2 Improvement of Plate Design

To address the limitations observed with the initial design, the width of the parallel plates was reduced, resulting in a more focused E-field with increased intensity over a smaller, controlled volume.

The relationship between charge distribution and electric field intensity can be understood through the surface charge density equation, given in [18] by

$$\sigma = \frac{Q}{A} \quad (\text{Eq. 3.2})$$

where σ (C/m²) is the surface charge density, Q (C) is the total charge on the conductor, and A (m²) is the surface area over which the charge is distributed. When the width of the plates is reduced, the available surface area decreases, thereby increasing the surface charge density for a given applied voltage. This increase in charge density directly enhances the strength of the resulting electric field, as described in [18] by

$$E = \frac{\sigma}{\epsilon} \quad (\text{Eq. 3.3})$$

where E (V/m) is the electric field intensity and ϵ (C/V · m) is the permittivity of the surrounding medium. Since σ increases as A decreases, provided that ϵ remains unchanged, the electric field strength correspondingly increases, given by an expanded form of Eq. 3.3 as

$$E = \frac{Q}{A\epsilon} \quad (\text{Eq. 3.4})$$

This explains why reducing the width of the plates leads to an increase in the field strength and concentration. Additionally, narrowing the plates reduces lateral spreading of the electric field, thereby improving concentration in the intended target area.

Not including the feed, the interior surface area of the top plate is 3750 mm². After accounting for the cutout below the feed, the interior surface area of the bottom plate is 7174 mm². This results in a total inward facing surface area of 10,924 mm² for the original Comdel plates pictured in Fig. 3.9. In the revised plate design from Fig. 3.14, discussed below, each of the plates have an interior surface area of 3650 mm², resulting in a total interior surface area of 7300 mm². Using the surface areas for the two plate designs and arbitrary constants of 1 for Q and ϵ , the expected increase in field intensity between the plates can be approximated by

$$\%_{incr} = \left(\frac{E_{revised}}{E_{original}} - 1 \right) \cdot 100 = \left(\frac{\frac{Q}{\epsilon A_{revised}}}{\frac{Q}{\epsilon A_{original}}} - 1 \right) \cdot 100 = \left(\frac{\frac{1}{1 \cdot 7300}}{\frac{1}{1 \cdot 10924}} - 1 \right) \cdot 100 = 49.64\% \quad (\text{Eq. 3.5})$$

where $E_{original}$ is the electric field between the original plates from Fig. 3.9 and $E_{revised}$ is the electric field between the revised plate design shown later in Fig. 3.14. This prediction for the approximate increase in electric field strength will be referenced later in this section.

It is worth noting that the reduction in interior surface area naturally decreases the overall parallel-plate capacitance which in turn increases the reactance between the plates based on the relationship shown in Eq. 2.19. The parallel plate capacitance can be expressed by

$$C = \frac{\epsilon A}{d} \quad (\text{Eq. 3.6})$$

where the capacitance, C (F), is directly related to the surface area, A [14]. However, the improvement in the revised plate design goes beyond just a change in the plates' nominal capacitance. By reducing the interior footprint and adjusting the plate geometry to better confine the field, the design also mitigates unwanted field effects, such as those surrounding the feed in Fig. 3.11. Thus, the improvement arises not only because the capacitance has changed in a circuit-theoretic sense, but also because the new geometry is more conducive to producing fields with a better potential of contributing to energy transfer into a material.

The plate length from Table 3.3 below was chosen to ensure that sufficient space was available for placing a material under test in a uniform region for RF exposure, preventing edge effects from dominating the response. Meanwhile this chosen length is approximately $\lambda/20$ for the smallest wavelength of 3000 mm at 100 MHz, keeping it well within the range of being electrically small and helping to maintain quasi-static field behavior. If the length was instead electrically large ($> \lambda/10$), the plates would have a greater potential to begin exhibiting behavior akin to a radiating antenna rather than a capacitive system [14].

Changes were made to the feed as well, being reduced to the dimensions listed in Table 3.3 and reflected in Fig. 3.15. The original plate design featured a relatively large feed, which was found to contribute to unintended field generation as discussed above in Section 3.3.1.

Aluminum was chosen as the material for the plates due to its excellent electrical conductivity, low cost, and lightweight properties. Its high conductivity minimizes resistive losses, ensuring efficient energy transfer between the plates while aluminum's non-magnetic nature prevents unwanted interactions with RF fields, which could otherwise distort the electric field or introduce additional losses. The new and improved plate design in Fig. 3.13, whose dimensions are given in Table 3.3, is modeled as shown in Fig. 3.14.

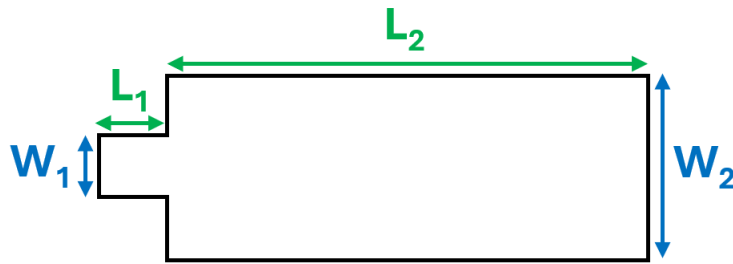


Figure 3.13: Layout of revised plate design.

Dimension	Value
L_1	5 mm
L_2	145 mm
W_1	5 mm
W_2	25 mm
Separation	30 mm
Thickness	1.5 mm

Table 3.3: Dimensions of revised plate design.

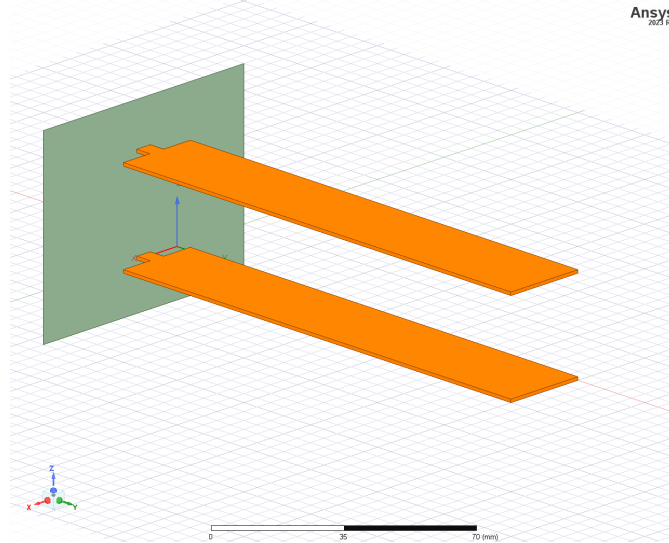


Figure 3.14: 3D HFSS model of the revised parallel plate design. The updated geometry reflects modifications aimed at improving field uniformity and confinement between the plates. This configuration was used in follow-up simulations for direct comparison to the Comdel plates, using the same boundary conditions and excitation setup to isolate the effects of the geometry change.

To maintain consistency in evaluating the effects of plate modifications, the same simulation setup was used for generating the field plots presented in Fig. 3.15 as was used in the analysis of the original plates in Fig. 3.11. Specifically, a terminal, $50\ \Omega$ waveport excitation was applied, with a power of 100 W and a voltage of $100\ V_{pk}$, surrounded by an air

box. Additionally, electric field visualizations were made using the cut planes placed with the same methodology as before in Fig. 3.11- the YZ plane cut was taken at the value along the X axis equal to the midpoint of the plate width, while the XZ plane cut was positioned at the value along the Y axis corresponding to half the total plate length. This described setup ensures that comparisons between different plate designs reflect only the effects of geometric modifications rather than any variations in the simulation setup. The resulting fields seen between the revised plate design are presented below in Fig. 3.15.

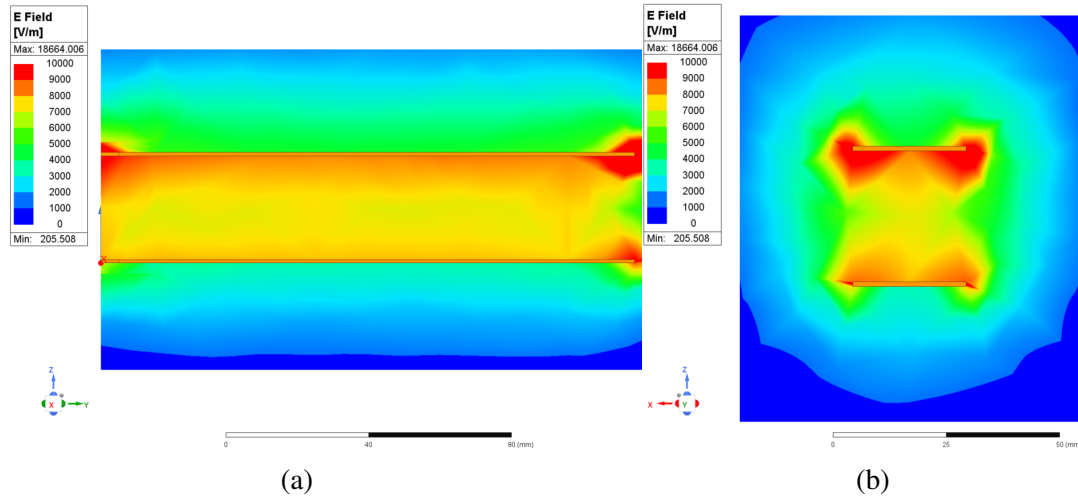


Figure 3.15: Electric field distribution between revised plate design when fed with $100 V_{pk}$, 13.56 MHz signal. Cross-sectional view in the YZ plane shown in (a), taken at the midpoint of the plate width. Cross-sectional view in the XZ plane shown in (b), taken at the midpoint of the plate length.

To quantitatively assess the improvement in electric field concentration between the original and revised plate designs, the average electric field strength was computed within a defined “core-volume” between the plates in both the original and revised design as pictured below in Fig. 3.16.

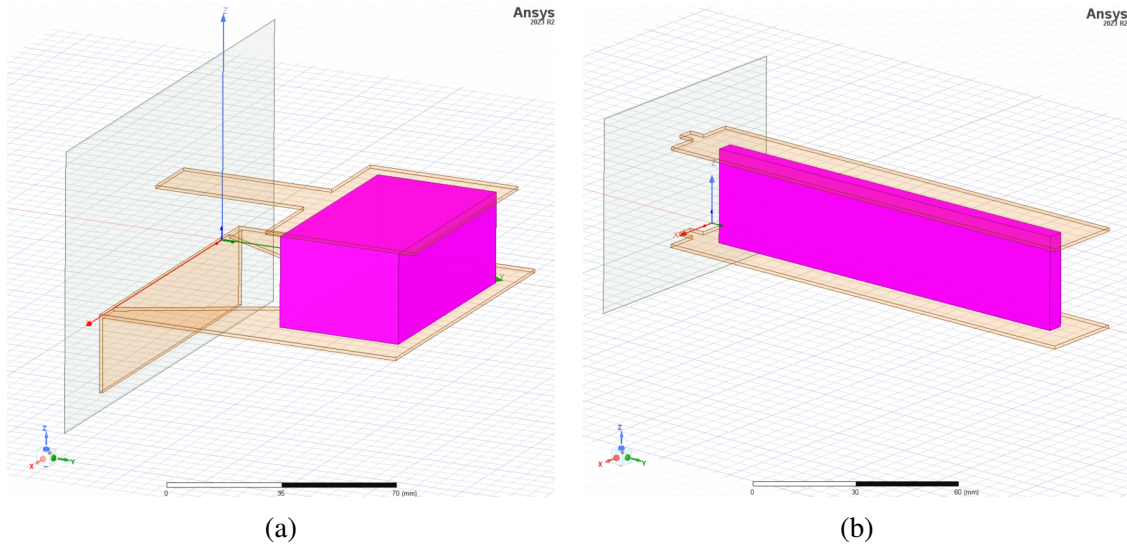


Figure 3.16: Non-model volumes created for field calculation in between the (a) original plates, (b) revised plates. Field calculations within volumes used to quantitatively assess field strength changes from resulting from design modifications.

The non-model volume boundaries were inset 10 mm inward from the edges of the plates in both the X and Y directions, ensuring that the calculation focused on the primary field region in the core-volume relevant to energy transfer and excluded regions near the edges of the plate. The full plate separation distance in the Z-axis was maintained for the core-volume, allowing for the capture of the electric field distribution throughout the intended operational space.

The average electric field strength within this volume was computed using the HFSS field calculator, adapted from [33] as

$$|E|_{avg} = \frac{\int_{Volume} |E| dV}{\int_{Volume} dV} (V/m) \quad (\text{Eq. 3.7})$$

where “Box1” is the volume. In Eq. 3.7 above, the electric field magnitude is integrated over the volume before being normalized by the total volume. As a result, the spatially averaged electric field intensity can be calculated. At 13.56 MHz, the results reveal an average electric field intensity inside the core-volume of 5,457 V/m for the original plate

design and 7,993 V/m for the revised plate design, a 46.47% increase.

To further validate these results, a second, “full-volume” was created between both plate designs without offsets in the X and Y directions, occupying the full region between the plates. Simulations with the full-volume were conducted, finding the average field intensity was 5,547 V/m for the original plate design and 8,132 V/m for the revised plate design, a 46.60% increase. The strong agreement between these two sets of results suggest that the impact of fringing fields and edge-effects appear insignificant and are proportionally similar between the original and revised design. Notably, these simulated results align closely with the 49.64% increase predicted by Eq. 3.5 using the change in interior plate area. Together, these findings help to confirm the improved ability of the revised plate design in increasing field confinement and intensity in the region between the plates.

While the results presented thus far have focused on the intended operating frequency of the original plate design, 13.56 MHz, it is important to recognize that the revised plate design is meant for a system operating over a much wider frequency range. Before concluding the discussion of the original plate design, an additional set of simulations was conducted to evaluate the field intensity of both plate designs over a greater frequency range, from 10 MHz to 100 MHz. In these simulations, no parameters were altered apart from the frequency sweep.

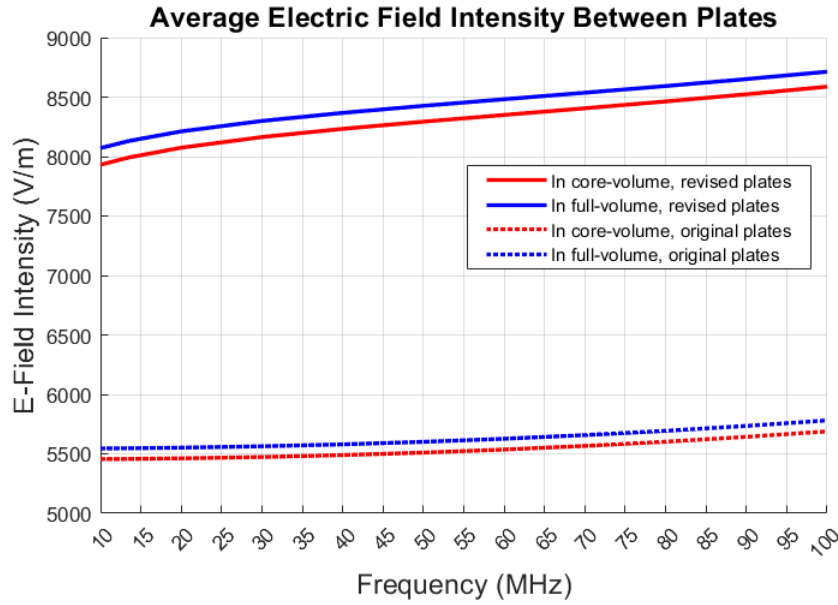


Figure 3.17: Average electric field intensity between original plate design (solid) and the revised plate design (dashed) in both the core-volumes (red) and the full-volumes (blue).

The average electric field intensity as a function of frequency for both the original and revised plate designs is illustrated above in Fig. 3.17. Across the frequency range from 10 MHz to 100 MHz, the revised plates consistently exhibit a higher field intensity, confirming the improved ability to generate and concentrate energy in the intended region. In both designs, the observed field intensity increases with frequency, which can be attributed to the increasing size of the plates relative to the wavelength and becoming less electrically small. While an increased electric field at higher frequencies is favorable, considerations such as those discussed in Section 2.1.2 should be taken into account. Additionally, the strong agreement between the core-volume and full-volume results support the previously presented field visualizations in Fig. 3.11 and Fig. 3.15- indicating that while the fringing fields are present, their impact remains minor and similar between designs, ensuring a fair comparison.

Following confirmation of the revised design's effectiveness, the parameters from Table 3.3 were used to fabricate physical versions of the plates from aluminum sheets. One of

these fabricated plates is shown below in Fig. 3.18.



Figure 3.18: Fabricated aluminum plate based on the revised design dimensions from Table 3.3, incorporating changes such as a reduced feed size to improve field confinement and response.

A critical aspect of this construction involved ensuring a stable and consistent separation between the plates. To accomplish this, a spacer was designed using PTFE (polytetrafluoroethylene), commonly known as Teflon. PTFE was selected as it is widely used as a dielectric material, such as in many coaxial cables, and thus is well understood. In fact, PTFE is included in HFSS' material library, with a dielectric constant of $\epsilon_r = 2.1$. Incorporation of the PTFE spacer model between the plates is shown below in Fig. 3.19.

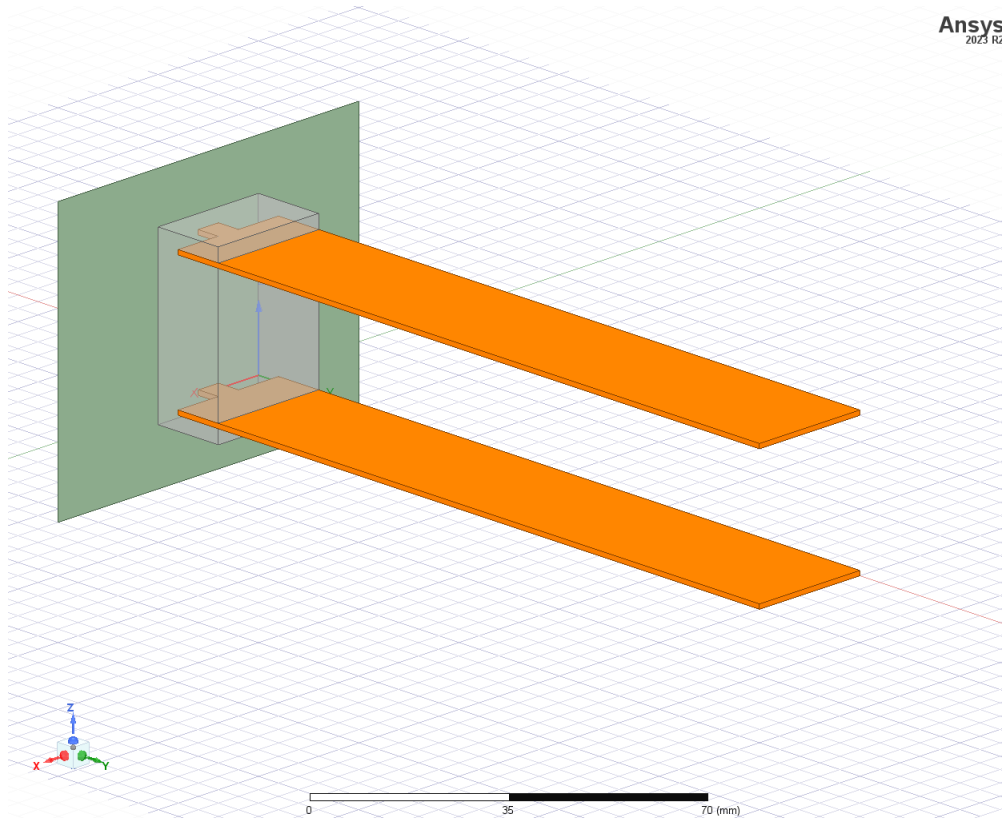


Figure 3.19: Simulation model of the revised plate design, incorporating a dielectric spacer to support plate structure and plate separation when later fabricated.

The simulation setup remained consistent with previous experiments- $50\ \Omega$ port set to have a power of 100 W and a source voltage of $100\ V_{pk}$. However, subsequent field visualizations in this section are shown at 50 MHz, a representative midpoint within the 10 MHz to 100 MHz frequency range. This selection provides a balanced reference point for comparing field behavior across different design modifications. As in previous field visualizations, the scale ranges from 0 V/m to 10 kV/m unless otherwise noted.

While the addition of the spacer led to a decrease in the field intensity between the plates in Fig. 3.15 below, it also significantly reduced the electric field intensity outside the plates. The decrease in field intensity can, in part, be attributed to the spacer. The extent of this loss is related to the dielectric constant of the spacer material, discussed in [13] and

later in this section below.

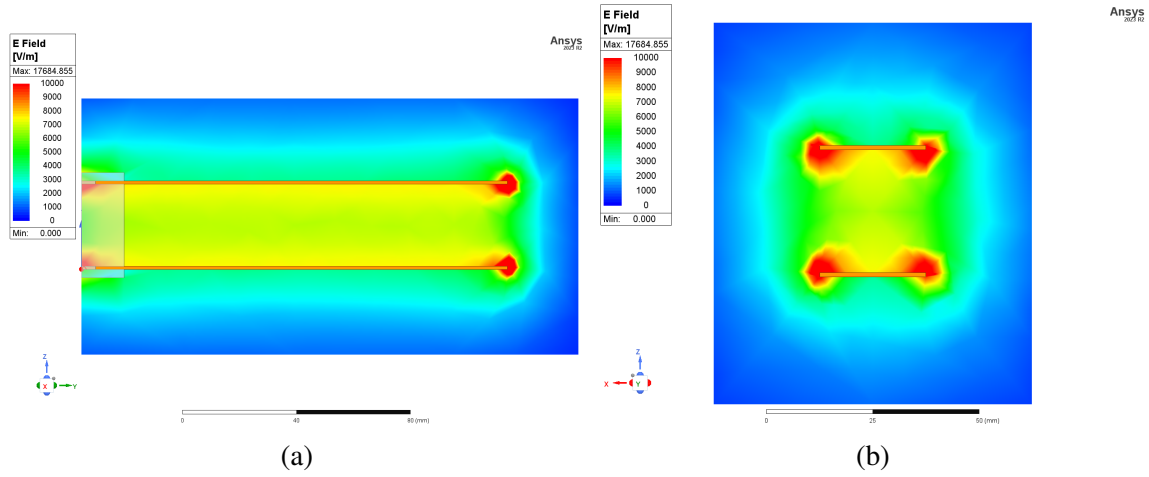


Figure 3.20: Electric field distribution between revised plate design with added spacer of $\epsilon_{r,spacer} = 2.1$, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown in (a), taken at the midpoint of the plate width. Cross-sectional view in the XZ plane shown in (b), taken at the midpoint of the plate length.

To better quantify the impact of the spacer, Eq. 3.7 was again used to find the average electric field intensity in the full-volume between the plates. At 50 MHz, the field intensity from the revised design with the spacer was found to be 7,054 V/m, a 16% decrease compared to the 8,426 V/m found in simulation for the revised plate design without a spacer. However, this still corresponds to a 26% increase relative to the 5,601 V/m measured from the original plate design at 50 MHz. These results demonstrate the spacer's role in defining the field behavior both within and outside of the plates.

Further simulations were run to explore the effect of the spacer's dielectric constant. While the default dielectric constant for PTFE in HFSS is set to $\epsilon_{r,PTFE} = 2.1$, values of $\epsilon_r = 10$ and $\epsilon_r = 25$ were used in subsequent simulations to maximize contrast with the default case. Fig. 3.21 below shows a YZ plane cut of the electric field for a value of $\epsilon_r = 10$ while the following figure, Fig. 3.22, reflects the same simulation but instead with $\epsilon_r = 25$.

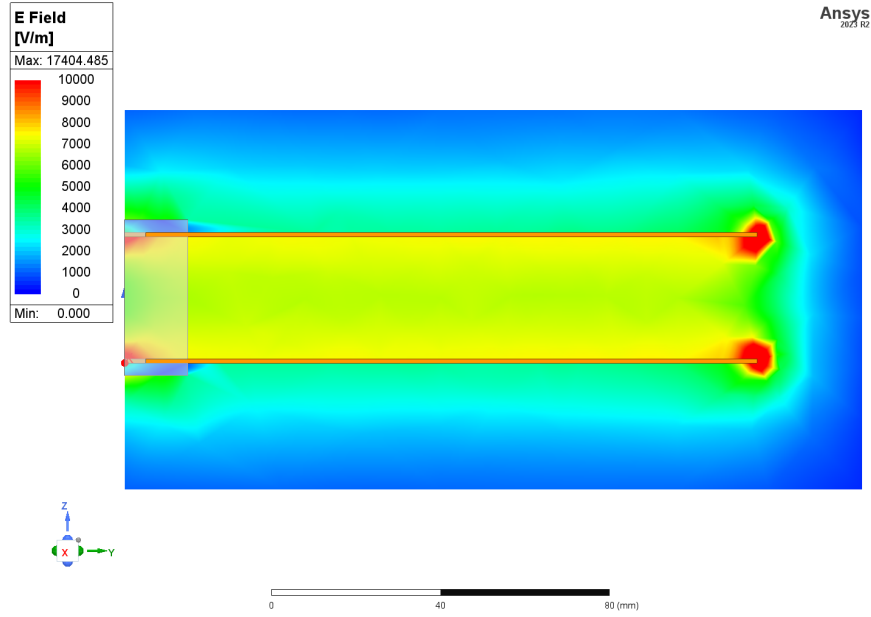


Figure 3.21: Electric field distribution between revised plate design with added spacer of $\epsilon_{r,spacer} = 10$, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown, taken at the midpoint of the plate width.

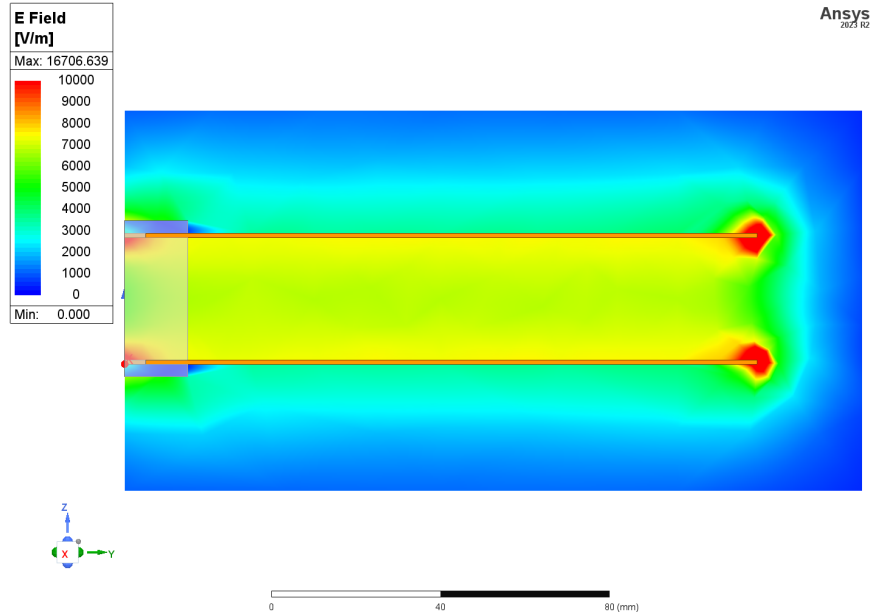


Figure 3.22: Electric field distribution between revised plate design with added spacer of $\epsilon_{r,spacer} = 25$, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown, taken at the midpoint of the plate width.

As the spacer was ultimately intended to be 3D-printed, the dielectric constant for this configuration was also taken into account. A key advantage of using a 3D-printed spacer is the ability to control the infill density, a common practice in 3D printing to minimize material usage while maintaining a degree of structural integrity. 15% infill is a typical choice, balancing material savings with sufficient strength for practical use. Precisely determining the dielectric constant for such a material would require a more complex model, accounting for field interactions within the heterogeneous structure and the effects of different infill patterns. Instead, a weighted average provides a straightforward and reasonable method for approximating the dielectric constant. For a spacer of 15% PTFE by weight, with the remaining internal structure comprised of air, the weighted average can be expressed as

$$\epsilon_{r,spacer} = (\%_{PTFE} \cdot \epsilon_{r,PTFE}) + (\%_{air} \cdot \epsilon_{r,air}) \quad (\text{Eq. 3.8})$$

Substituting $\epsilon_{r,PTFE} = 2.1$ and $\epsilon_{r,air} = 1$, the effective dielectric constant of the spacer determined as

$$\epsilon_{r,spacer} = (0.15 \cdot 2.1) + (0.85 \cdot 1.0) = 1.165 \quad (\text{Eq. 3.9})$$

A YZ plane cut of the electric field for a simulation using this value of $\epsilon_{r,spacer} = 1.165$ is shown below in Fig. 3.23.

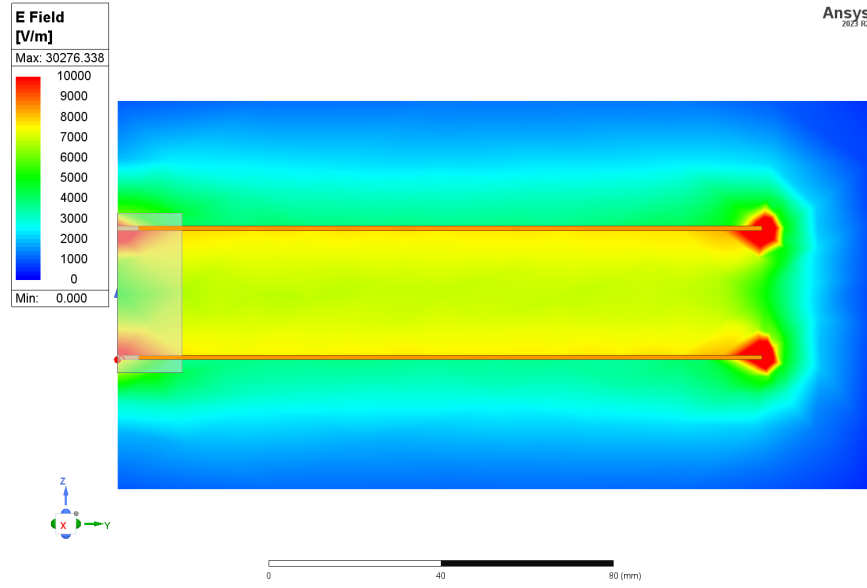


Figure 3.23: Electric field distribution between revised plate design with added spacer of $\epsilon_{r,spacer} = 1.165$, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown, taken at the midpoint of the plate width.

With a dielectric constant of $\epsilon_r = 10$, the field intensity between the plates decreased slightly, as observed in Fig. 3.21 to an average field in the full-volume between the plates of 7,033 V/m. When the dielectric constant of the spacer was further increased to $\epsilon_r = 25$, the average field in Fig. 3.22 over the full-volume between the plates decreased very little, to 7,032 V/m. The estimated dielectric constant for a 3D-printed spacer with 15% infill ($\epsilon_r = 1.165$) shown in Fig. 3.23 resulted in an average field in the full-volume of 7,171 V/m.

A higher dielectric constant is associated with increased dielectric loss, as discussed in [13], leading to greater power dissipation within the spacer. Increased losses in the spacer are likely to lead to unwanted heating and reduce energy transfer to the intended region between the plates. To further investigate this effect, the total loss within the spacer was calculated by performing a volumetric integration of the volume loss density using the HFSS field calculator as

$$P_{loss} = \int_{Volume} \rho_{loss} dV \quad (\text{Eq. 3.10})$$

To summarize the findings of changes to the dielectric constant of the spacer, Fig. 3.24 presents the total losses and field intensities.

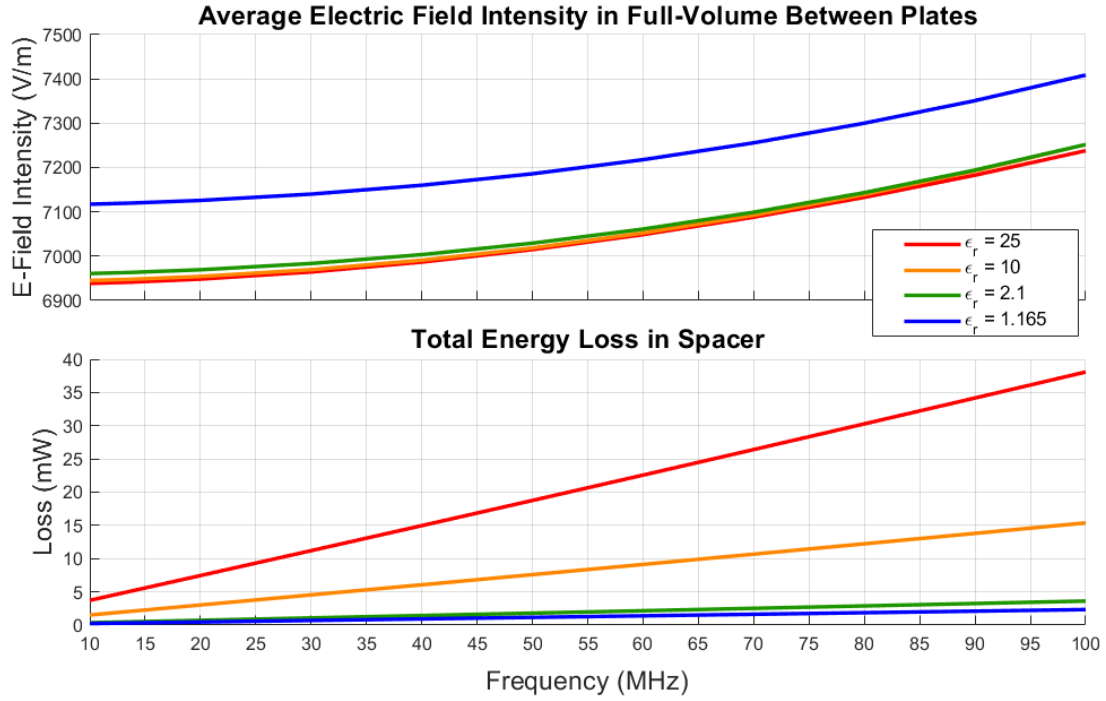


Figure 3.24: Comparison of the average electric field intensity in the full-volume between the plates (top) and the total volumetric energy loss in the spacer (bottom) as functions of frequency for different spacer dielectric constants.

The results from Fig. 3.24 show that the loss in the spacer increases with both dielectric constant and frequency, compounding together for significant increases. However, increases in the field appear to be much more significantly driven by an increase in frequency than changes in the dielectric constant, supporting the field behavior previously observed in Fig. 3.17.

Based on the diminishing effect of ϵ on field intensity alongside its increasing effect on losses in the spacer, the decision was made to proceed with a 3D-printed spacer with

15% infill. While Fig. 3.24 suggests that further reducing the infill and consequently the dielectric constant of the spacer could be beneficial, doing so would risk compromising the spacer's structural integrity.

The spacer design was able to be recreated in Fusion360 before preparing it in OrcaSlicer and producing a physical spacer in Polymaker PTFE on a personally owned Creality Ender-3 V3 KE 3D printer. The fabricated spacer can be seen in Fig. 3.25 below.

Since the RF signal is fed into the top plate while the bottom plate is connected to ground, as depicted in Fig. 3.8 and discussed in Section 3.3.1, the parallel plates inherently form an unbalanced system [13]. This unbalanced configuration enables a well-confined electric field to be developed between the plates. This contrasts to parallel-plate waveguides which, while similar in structure, use a balanced feed where both conductors typically carry signals of equal magnitude but opposite phase [13].

To provide a direct RF feed to the plates, an SMA connector, inherently an unbalanced structure, was selected. The center pin of the SMA connector, carrying the RF signal, was connected to the top plate, while the outer conductor serving as ground was connected to the bottom plate. Given that both the plates and the coaxial feed structure are unbalanced, this feeding method is compatible with the system.



Figure 3.25: Fabricated plate assembly with 3D-printed spacer used to support structure and SMA connector used to feed a signal to the plates.

While the connection from the SMA connector to the plates has the potential to introduce unintended radiation due to the transition from an unbalanced coaxial feed to a balanced plate structure, its impact is expected to be minimal. The radiated power can be quantified by

$$P_{rad} = \frac{1}{2} |I_0|^2 R_r \quad (\text{Eq. 3.11})$$

where P_{rad} (W) is the radiated power, I_0 (A) is the feed current, and R_r (Ω) is the radiation resistance [25]. The radiation resistance, which represents the equivalent resistance accounting for power radiated as electromagnetic waves, is given by

$$R_r = 80\pi^2 \left(\frac{l}{\lambda} \right)^2 \quad (\text{Eq. 3.12})$$

where l (m) is the conductor length and λ (m) is the wavelength [25]. The connection between the center pin of the coax connector, pictured in Fig. 3.25, spanning half the separation of the plates, is 15mm. At the maximum frequency within the sweep of 100 MHz,

the radiation resistance can be found as

$$R_r = 80\pi^2 \left(\frac{15mm}{3m} \right)^2 = 0.02 \Omega \quad (\text{Eq. 3.13})$$

This low radiation resistance suggests that only a negligible portion of the power is lost at the feed. The current through the feed into the plates would be at its maximum with a perfect impedance match of exactly 50Ω . Thus, the expected current for a perfect match is found as

$$I_{pk} = \frac{V_{pk}}{Z_L} = \frac{100}{50} = 2 \text{ A} \quad (\text{Eq. 3.14})$$

where the voltage established at the port is divided by the ideal load impedance, Z_L . The power radiated through the wire feed from a perfect match can be found through Eq.3.11 as

$$P_{rad} = \frac{1}{2} |2|^2 (0.02) = 0.04 \text{ W} \quad (\text{Eq. 3.15})$$

showing only 0.04 W being lost in the feed.

After constructing the new plate assembly shown in Fig. 3.25, the next step involved connecting it to a VNA (Vector Network Analyzer) to observe its response under a low-power feed. Although the plates will ultimately be connected to a high-power source, initial testing with a low-power feed reduces the need to account for stringent safety precautions. For this measurement, the frequency was swept from 10 MHz to 100 MHz while the 50Ω port was set to output 10 dBm (0.01 W), which, based on Eq. 3.1 used to find expected voltage given power and impedance, corresponds to a source voltage at the port of $0.7071 V_{rms}$, or $1.0 V_{pk}$. The VNA used was an Agilent Technologies N5225A. After first calibrating the VNA using an N4691-60006 E-cal kit, the plates were connected and the S-parameter data was exported to MATLAB to create plots. A layout of this test setup is shown below.

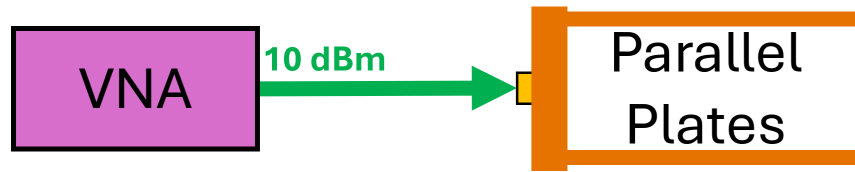


Figure 3.26: Layout of setup used to measure S-parameters of the fabricated plates.

The S-parameters from the physical plates, pictured in Fig. 3.28, and the S-parameters from the simulated plates, pictured in Fig. 3.27, can be compared below.

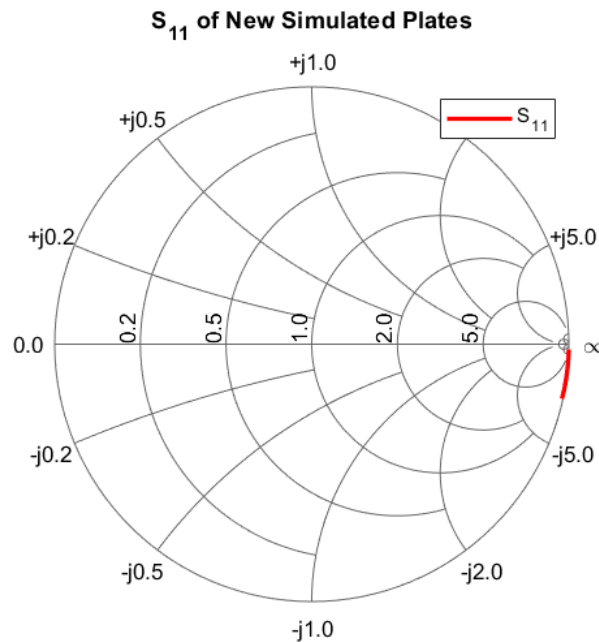


Figure 3.27: S-parameter response of revised simulated plates with spacer. Frequency swept from 20 MHz to 100 MHz. Response shows plate impedance as having a small resistive component and large capacitive reactance.

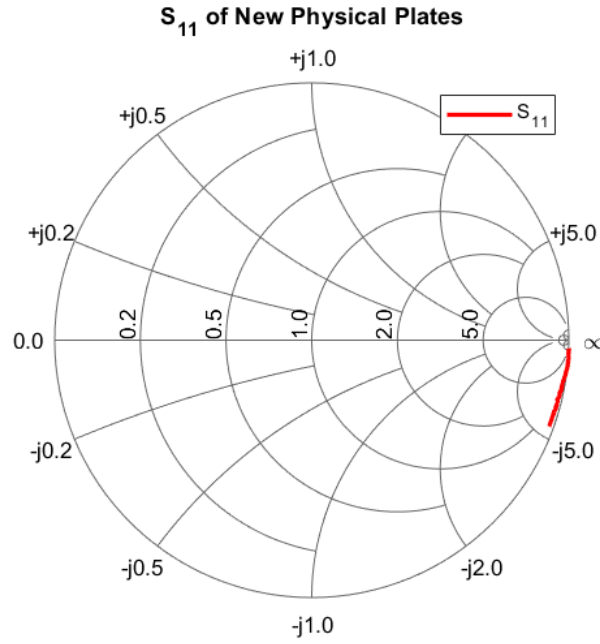


Figure 3.28: S-parameter response of new fabricated plates with spacer. Frequency swept from 20 MHz to 100 MHz. Response shows plate impedance as having a small resistive component and large capacitive reactance. However, the fabricated plates show a slightly larger resistive component than simulated design.

A notable observation is that the fabricated plates are both more resistive and less reactive than the plates used in simulation. Particularly as a result of the decreased reactance, the response of the physical plates in Fig. 3.28 appears to exhibit greater variance across frequency when compared to the simulated plates in Fig. 3.27. This variation can likely be attributed to several factors specific to the construction and measurement environment of the physical model. The connector presence likely causes additional reflections and losses not accounted for in the simulation as well as any minor fabrication tolerances such as slight deviations from the dimensions listed in Table 3.3. Additionally, the measurement setup when testing with the VNA despite calibration, such as any nearby objects interacting with the fields, has the possibility to change the response.

Regardless, the fabricated plates represent an improvement compared to the Comdel

plates shown in Fig. 3.12 . As discussed previously, the driving factors behind this are the reduced interior surface area of the plates and the feed. Additionally, this highlights the changes in reactance seen between the original Comdel design and the new plate design- a reduction in unnecessary capacitive effects. The refined plate design demonstrates enhanced field concentration, aligning more closely with the design objectives. These results highlight the effectiveness of the updated plate geometry in addressing the limitations observed in the original design.

These improvements are particularly relevant when considering practical applications where the interaction between the electric field and various materials is key. The next step involves running simulations with a human tissue analog between the plates, providing insights into how the electric field behaves and assessing the system's ability to transfer energy into a medium.

3.3.3 Simulations with Human Tissue Analog

In designing a parallel plate system for cancer ablation, a primary step involves properly simulating the dielectric properties of the material expected to be between the plates- in this case, human tissue. For accurate HFSS simulations, the material properties relevant to RF interactions must represent biological tissue. The material properties of human tissues can vary widely, described in [34], [35], [36], [37], and [38], depending on the specific tissue type and location within the body. Thus, the values presented in Table 3.4 serve as a general approximation. Again, while this tissue model isn't exact, the tissue-specific material parameters highlighted below in Table 3.4 serve to provide an estimate for accurate simulations and field interactions.

Material Property	Value
Permittivity	80
Electrical Conductivity	0.5 S/m
Loss Tangent	2
Thermal Conductivity	0.5 W/m · K
Mass Density	1100 kg/m ³
Specific Heat	2900 J/kg · C

Table 3.4: Material Properties of tissue used in HFSS Simulations.

Including a mass of tissue between the plates is bound to directly impact the E-field distribution, penetration, and S-parameter response. The dielectric properties of tissue, particularly the high permittivity, will modify the impedance characteristics of the medium between the plates, affecting the E-field intensity at the surface and altering the response observed on the Smith Chart. To better understand these effects, a tissue sample was modeled and placed between the plates from Fig. 3.19.

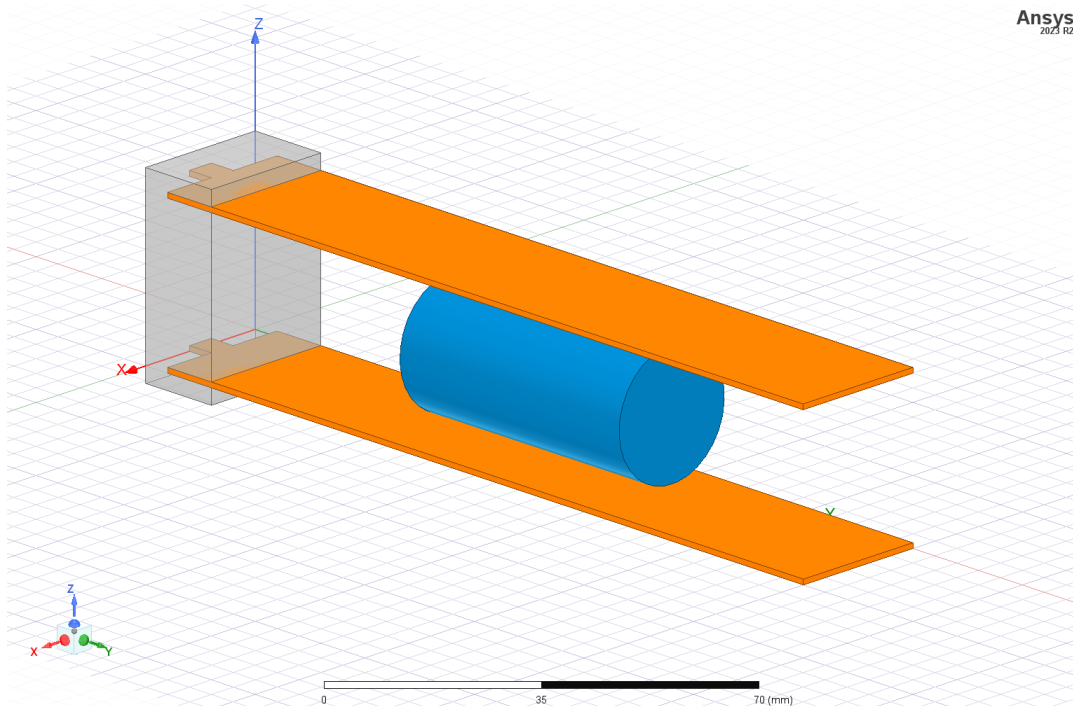


Figure 3.29: 3D model from HFSS of the plates with a tissue sample placed between.

The tissue sample in Fig. 3.29 above was designed as a cylinder with a radius of 12 mm and a length of 50 mm. Based on the mass density values provided in Table 3.4, this model approximates a 25 g sample of tissue, helping to provide meaningful analysis of field interactions.

The cylindrical shape of the tissue sample is well-suited for modeling as it better approximates the rounded contours of many real-world tissue structures, such as muscle fibers, bones, or organs. While biological tissue is not inherently cylindrical, it lacks the sharp edges and corners that a rectangular shape would introduce. From a modeling perspective, a cylindrical geometry more closely aligns with the smooth, continuous nature of biological materials, making it a more representative approximation for simulation.

The following simulation presented in Fig. 3.30 maintains the same setup, excitation parameters, frequency, and source power as in previous analyses such as Fig. 3.23, ensuring that any observed differences in field distribution result solely from the introduction of the

tissue sample. These fields in the YZ plane are presented in Fig. 3.30a while the fields in the XZ plane are presented in Fig. 3.30b.

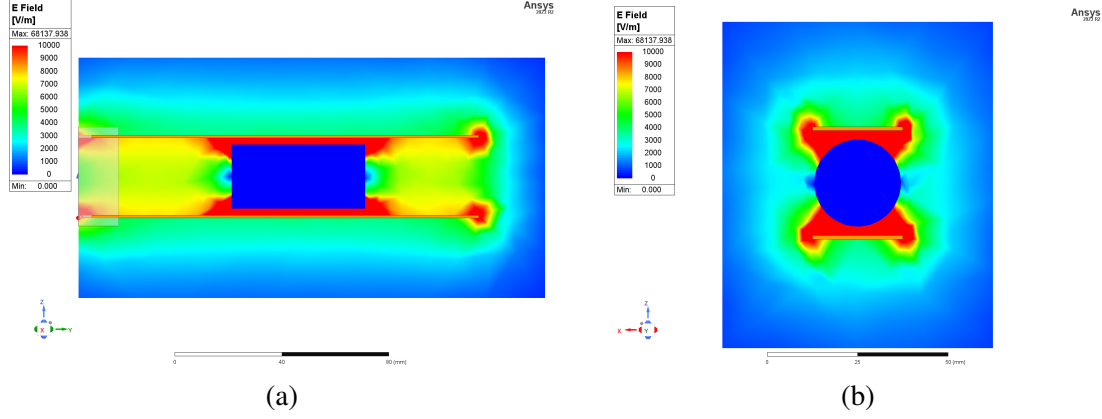


Figure 3.30: Electric field distribution between revised plate design with added tissue sample, fed with $100 V_{pk}$, 50 MHz signal. Cross-sectional view in the YZ plane shown in (a), taken at the midpoint of the plate width. Cross-sectional view in the XZ plane shown in (b), taken at the midpoint of the plate length.

The introduction of the tissue analog between the plates in Fig. 3.30 above substantially alters the electric field distribution, producing a notable increase in field intensity in the regions immediately adjacent to the top and bottom surfaces of the tissue sample. This effect closely resembles the field behavior observed in [7, Fig. 3a] (presented below in Fig. 3.31), where their COMSOL simulations revealed strong electric field enhancement in the air gap regions between the plates and the material under test, with comparatively weak fields within the material itself.

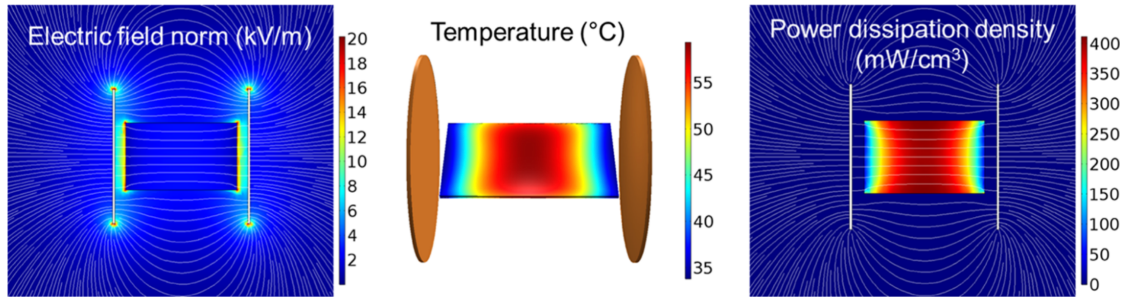


Figure 3.31: COMSOL models showing electric field strength around the sample (2.5 wt% CNT in PLA) and near the electrodes (left), the steadystate temperature field generated in a sample (middle), and the power dissipation density (right). [7, Fig. 3a]

While Sweeney does not provide explicit, quantitative data for the electric field values in Fig. 3.31, the provided color scale suggests that the field strength in the air gap region exceeds 10 kV/m, even reaching values as high as 20 kV/m at the surface of the sample. In contrast, the HFSS simulations performed in this study show that the majority of the electric field values in this same region of Fig. 3.30 exceed 10 kV/m, which is the upper limit of the original color scale used for visualization.

At the same time, the electric field intensity within the tissue itself is significantly weaker by comparison. While significantly weaker than in the surrounding region, an adjusted scale from 0 to 100 V/m in Fig. 3.32 below more clearly shows the field strength within the tissue sample.

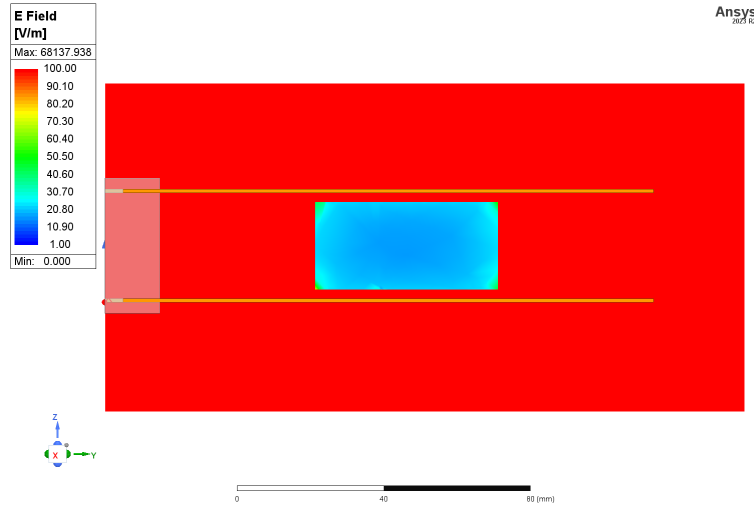


Figure 3.32: Electric field distribution between revised plate design with adjusted scale, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown, taken at the midpoint of the plate width. Results highlight fields induced within tissue sample.

This critical aspect, the power dissipation density within the sample, was examined in the Sweeney study as well, reported in Fig. 3.31. While specific numerical values are again not explicitly listed, the color scale suggests that within the material, power dissipation densities of at least 0.250 mW/cm³ were observed at 50 MHz, with some regions exceeding 350 mW/cm³.

Using an approach similar to the method previously applied to compute the average electric field intensity in Section 3.3, the average volume loss density (analogous to power dissipation density) in the tissue sample at was found to be 10.286 mW/cm³ at 50 MHz. With the known volume of the cylinder, the total loss within the spacer can be calculated by performing a volumetric integration of the volume loss density using the HFSS field calculator (initially detailed by Eq. 3.10). The total volume loss in the tissue sample was computed as 232.66 mW. While the loss in the tissue sample is lower than the value observed in 3.31, direct comparisons are not entirely appropriate due to differences in material properties and test conditions.

A key takeaway from both studies is the qualitative similarity in observed field behavior,

where the introduction of a material between parallel plates results in increases in field intensity at and around the air-material interface while suppressing field intensity within the material itself. This phenomenon, as discussed in [7], aligns with theoretical expectations, as the continuity condition for electric flux density, D (C/m²), at the interface dictates that the normal component of must remain continuous across the boundary. The relationship between the electric field, the electric flux density, and permittivity is given in [14] as

$$E = \frac{D}{\epsilon} \quad (\text{Eq. 3.16})$$

and since the tissue analog exhibits a much higher permittivity (ϵ) than air, the resulting electric field (E) within the tissue is significantly reduced compared to the surrounding air. This redistribution of fields is evident in Fig. 3.30 and Fig. 3.31. Despite differences in material properties and test conditions, the qualitative agreement between these results and those from [7] reinforce the fundamental role of material properties, particularly permittivity in this case, in determining RF field distributions and power transfer.

The presence of the tissue analog not only impacts the localized field behavior but also influences the overall impedance characteristics of the system. Fig. 3.33 below presents the S-parameter response for both configurations over the range of 10 MHz to 100 MHz, with the plates alone in red (previously shown in Fig. 3.27) and with the tissue sample being introduced in blue.

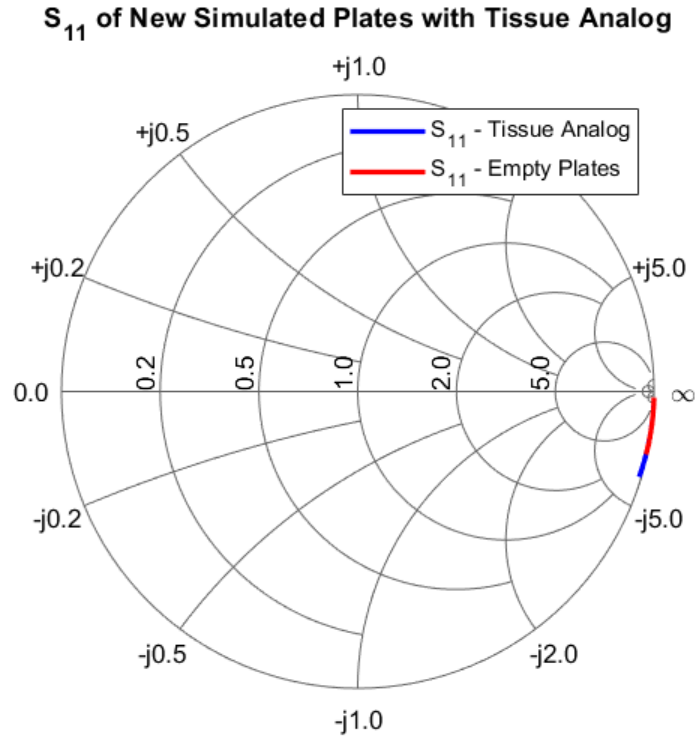


Figure 3.33: S-parameter response of the plates alone (red) and of the plates with a tissue sample (blue). Frequency swept from 20 MHz to 100 MHz. Results highlight tuning needs of network due to change in impedance when tissue sample is added between plates.

The results presented above in Fig. 3.29 indicate that the introduction of the tissue analog alters the impedance response, leading to greater variation in the complex impedance across the same frequency range when compared to the plates alone. This further confirms that the dielectric properties of the tissue influence the behavior of the system, emphasizing the need for a robust matching network capable of adapting to these variations that is yet to be commercially available.

With this in mind, the following section explores the development of a tunable matching network, integral to optimizing power transfer to the plates and accommodating the varying material properties of a load over a range of frequencies.

3.4 Matching Network Simulation and Design

The design of an effective and tunable matching network is essential for optimizing power transfer in an RF-based system for cancer ablation. The network ensures that the maximum possible RF power is directed into the load, being the parallel plates and the material placed between them, while minimizing reflections that reduce the efficiency of power delivery. This chapter details the simulation and design process of the matching network, starting with observations from the original Comdel system and the subsequent development of a refined network with a simplified configuration. The proposed network leverages a new component layout and aims to cover a broad operational frequency range.

3.4.1 Comdel System Design

The matching network that had been previously used in [1], specifically the Comdel MatchPro CPMX-6000, pictured below in Fig. 3.34, served as a baseline for further design and refinement.

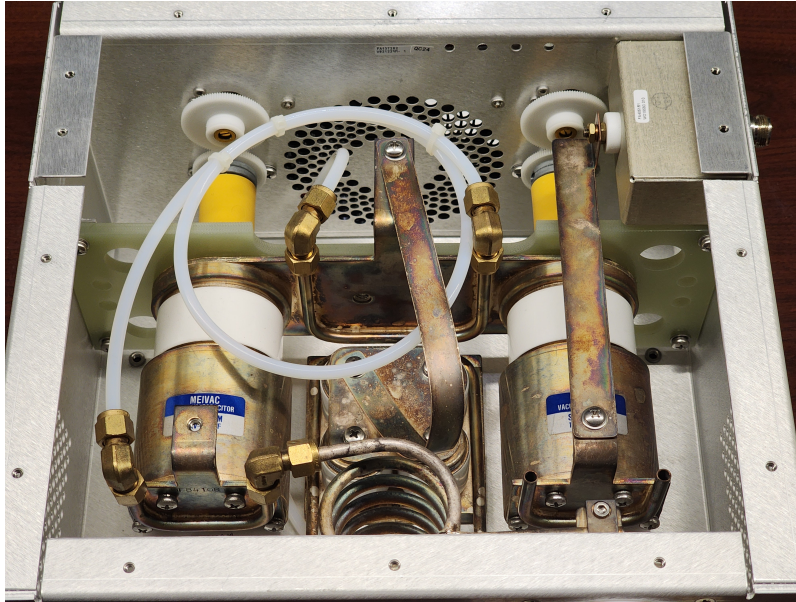


Figure 3.34: Comdel Matching Network used for experimentation in [1], limited to use at a fixed-frequency of 13.56 MHz.

Observations of the existing network design indicate a sequential arrangement of components as follows from source to load: source, a series tunable capacitor, a shunt LC series combination with the capacitor component configured as a capacitor bank, a second series tunable capacitor, a shunt inductor, and lastly the load. To better analyze the Comdel network's behavior, an equivalent circuit, shown below in Fig. 3.35 was constructed in ADS.

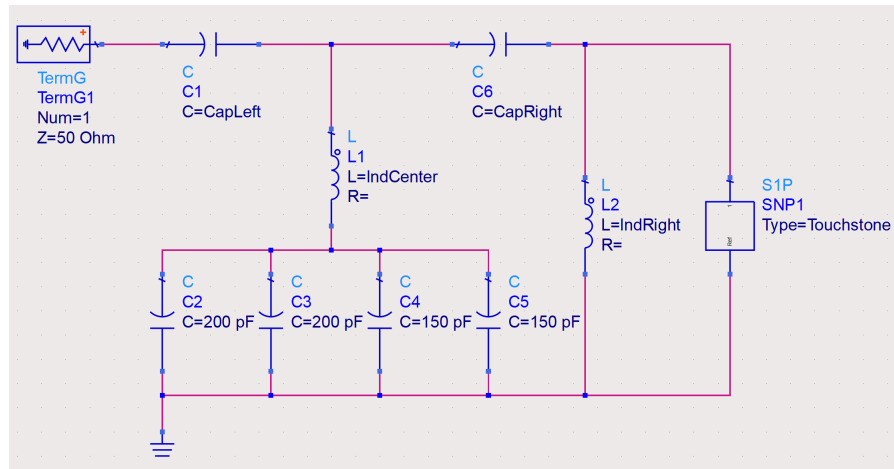


Figure 3.35: ADS schematic of the Comdel CPMX-6000 matching network, modeled to analyze circuit behavior and understand component influence on impedance matching performance.

The S-parameters of the original parallel plates in Fig. 3.12 were imported as an ADS component block, effectively representing the load. This setup allows for use of the tuning feature within ADS, allowing for insights into the behavior of individual components within the network. Notably, the tunable capacitors exhibited a degree of freedom in terms of impedance control, impacting impedance matching while only minimally altering the frequency with the maximum response. Conversely, the tuning of the inductors primarily influenced the frequency with the maximum response while only minimally affecting the distance between the response and the center point of the Smith Chart. The design's ability to operate with two degrees of freedom, better visualized in Fig. 3.36, helps to inform the later development of the refined design.

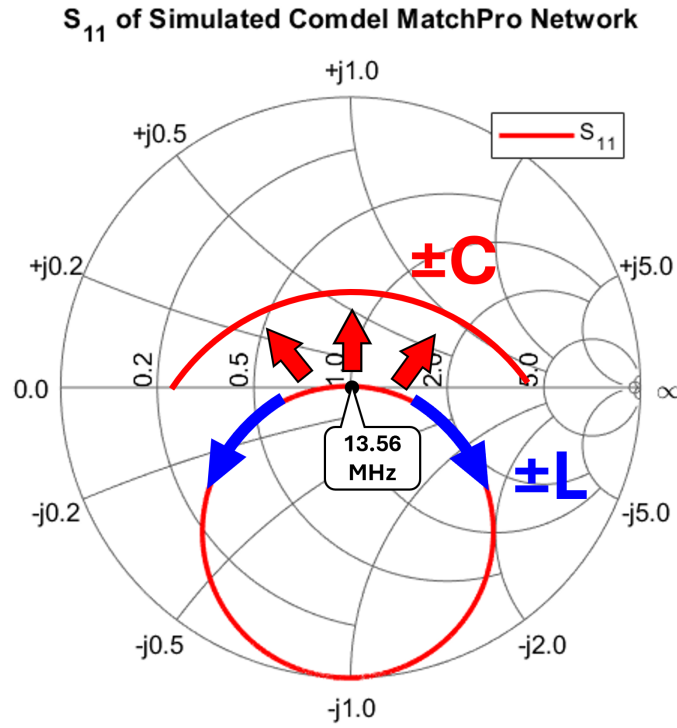


Figure 3.36: Smith chart illustrating tuning behavior of the simulated Comdel MatchPro network. Changes in capacitance (red) and inductance (blue) move the network response along distinct paths, revealing tuning ability to operate primarily in two degrees of freedom.

3.4.2 Novel Network Design

Following the analysis of the Comdel system, opportunities for improvement were identified, particularly in reducing complexity and enhancing tuning flexibility. Keeping in mind the Comdel system was originally designed for operation at a fixed frequency of 13.56 MHz, the particular reliance on a fixed shunt inductor-capacitor bank limiting tuning flexibility suggested the system's design could be improved.

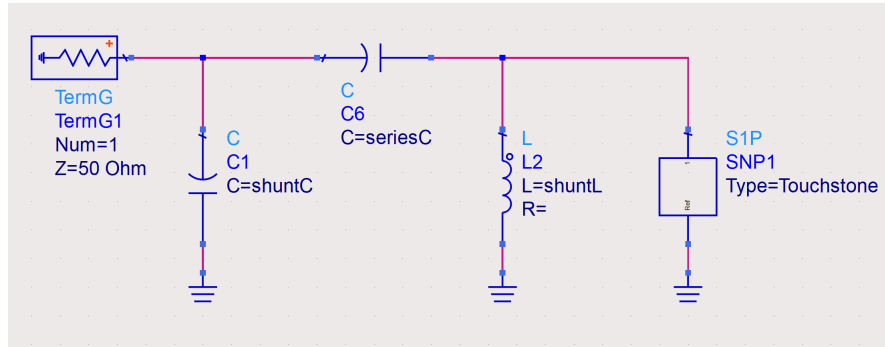


Figure 3.37: Schematic of the novel matching network in ADS, consisting of a shunt capacitor, series capacitor, and shunt inductor to achieve tunable impedance matching across a broad frequency range. The measured S-parameters from the fabricated plates were imported as the load, allowing the network to be tuned to load conditions.

The refined matching network in Fig. 3.37 was designed to achieve impedance matching across a broad frequency range while minimizing circuit complexity and maximizing tunability. The shunt inductor and series capacitor were retained from the previous design; however, the number of reactive components was reduced, simplifying the layout and reducing insertion losses. The measured S-parameter data previously presented in Fig. 3.28 from the newly designed physical plates was included as the load in ADS, enabling a more accurate simulation of the ability of the network to create a match. This configuration provided a basis for evaluating the effects of adding in components before the load, moving towards the source, pictured below in Fig. 3.38, again at the same 50 MHz.

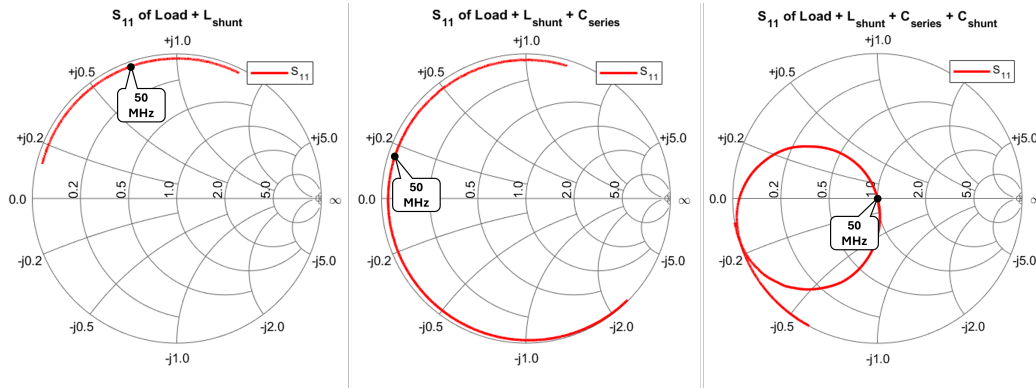


Figure 3.38: Adding in components from load to source, first shunt inductor (100 nH), then series capacitor (125 pF), lastly shunt capacitor (400 pF). Matched at 50 MHz. Frequency swept from 20 MHz to 100 MHz in each figure.

After the load was added, a shunt inductor was introduced (Fig. 3.38, left); the resulting S-parameters were then examined. This inductance adds positive reactance, based on Eq. 2.18, thus bringing the impedance seen by the source closer to a purely real value. Next, a series capacitor was introduced (Fig. 3.38, middle), and its impact on the S-parameters was similarly assessed.

The series capacitance does change the reactance based on Eq. 2.19 but also serves to act as a DC block, step up the voltage, and provide a degree of separation between the other two elements as discussed before in Section 3.4.1 and later on in this section. The increased voltage over the source amplitude is a key ability of this network, formed from the “circulation” of the energy between shunt and series elements which alternate between storing and releasing power over each cycle at the target frequency [13].

Lastly, the shunt capacitor was added (Fig. 3.38, right), with another review of the S-parameter response to confirm a match had been achieved at 50 MHz. The shunt capacitance brings the impedance to a match, moving the impedance along a constant-admittance circle on the Smith Chart as pictured in Fig. 2.2. This step-by-step approach enabled a clear understanding of each component’s influence on the overall impedance matching.

The component values ultimately used in this simulated network to create matches are presented below in Table 3.5.

Frequency	20 MHz	50 MHz	100 MHz
C_{shunt}	1102 pF	412 pF	207 pF
C_{series}	140 pF	140 pF	140 pF
L_{shunt}	486 nH	92 nH	28 nH

Table 3.5: Component values used in ideal network ADS simulation for a match to the physical plates.

Based on the findings in Table 3.5 from work in ADS, the series capacitor can remain fairly consistent across the range of operating frequencies, effectively separating the two shunt elements, reminiscent of the previously employed Comdel network described in Section 3.4.1. This separation enables a significantly easier tuning process by only requiring changes in the shunt component values.

However, just as critical as achieving a proper match is the network's role in increasing the voltage across the plates, directly impacting the field strength and power deposition into the sample (further explored in Section 3.4.4). This voltage increase was discovered to be heavily dictated by the selection of component values, particularly the inductance of the shunt element. By maximizing the inductance value as much as possible, the network is able to create a higher voltage step-up effect, reinforcing field strength (detailed by Eq. 3.17 in Section 3.4.4). However, the practical difficulty in achieving this is discussed next in Section 3.4.3. As such, the component values in Table 3.5 used for ADS simulations were deliberately constrained to generally align with the expected tuning limitations of the selected components detailed in Section 3.4.3.

The result of this initial simulation, representing the scenario in which the network is tuned for a match at 50 MHz, the same frequency used as an example in the HFSS

simulations from Section 3.3, is shown below in Fig. 3.39 in red, illustrating how the designed network successfully transforms the impedance of the plates, confirming that the network operates as intended under ideal conditions.

Building upon this, a second simulation was conducted to assess the theoretical impact of introducing the tissue analog between the plates. In this case, the simulated response of the plates with the tissue analog from Fig. 3.33 was used as the load in ADS, while the matching network itself remained unchanged. This approach provides insight into how the presence of a material may influence the impedance transformation without any compensatory adjustments to the network. The corresponding Smith Chart plot, shown in Fig. 3.39 in blue, demonstrates how the addition of the tissue alters the impedance “seen” by the plates.

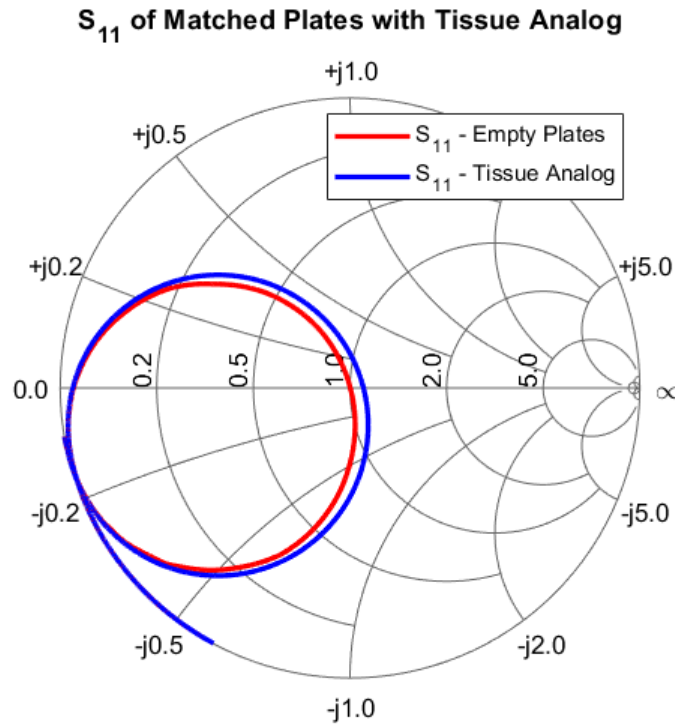


Figure 3.39: S-parameter response of the matched plates at 50 MHz without a tissue sample (red) and with a tissue sample (blue). Identical matching network layout and component values were used for in both cases. Frequency swept from 20 MHz to 100 MHz.

While the fundamental matching behavior remains, the shift in the impedance trace highlights the frequency-dependent nature of the tissue's dielectric properties and its effect on the measured response. This aligns with the observations made in Section 3.3.3, demonstrating the benefit in having the ability to adapt to variations in load conditions.

3.4.3 Component Selection and Implementation

The selection of components for the impedance matching network was guided by the simulated values outlined previously in Table. 3.5. These values were determined as the minimum required to ensure the feasibility of impedance matching across the intended frequency range. However, recognizing that real-world variations and practical constraints may arise, the tuning range of the selected components was deliberately chosen to be larger than the simulated values. This additional range helps to ensure an adequate ability in achieving an impedance across varying conditions.

The Passive Plus PPZN60100 capacitor was chosen for its broad tuning range from 26 pF to 600 pF [39]. This large variance in capacitance provided flexibility in achieving impedance matching across the desired frequency band. Additionally, this capacitor, intended for use in RF equipment, has a voltage rating of 150 VDC [39].

Commercially available tunable inductors generally exhibit a narrow tuning range, making it difficult to cover the required inductance values without additional design accommodations. To overcome this, a switchable inductor bank was implemented using four Radiall R570312000 RF switches. These switches operate from DC to 3 GHz with an impedance of 50 Ω , ensuring compatibility with the rest of the system while their low insertion loss of 0.15 dB and high isolation of 80 dB minimized undesired signal interactions [40]. Each switch can handle 240 W of power, making them suitable for high-power RF applications [40]. The inductor bank itself consists of three discrete inductors, selected

to provide overlapping tuning ranges and allow for stepped variation in the overall inductance. These included the JLC02E030TRSM, JLC04E065TRSM, and JLC06E110TRSM. The JLC02E030TRSM has a nominal inductance of 33 nH and a tuning range of 32 to 39 nH, the JLC04E065TRSM is centered around 65 nH with a tuning range of 60 to 78 nH, and the JLC06E110TRSM provides a base inductance of 110 nH, tunable from 94 to 123 nH [41]. These inductors were selected to maximize coverage as much as possible while maintaining reasonable tuning steps. Additionally, a fourth inductance was added using a 7 mm diameter single loop of 18 AWG wire, providing a small, additional inductance estimated to be approximately 18 nH, used in parallel to aid in achieving a match at higher frequencies where a small inductance is necessary.

The narrow range of tunable inductors prevented using sole use of the inductor bank to adequately achieve impedance matches. As a result, the series capacitor was implemented using the PPZN60100 tunable capacitor rather than a the fixed-value capacitor suggested by Table 3.5. This decision, while reducing the ability to achieve matching through tuning just two components, ultimately provided a workable implementation given the constraints of component availability. In future work, the ability to develop or source inductors with a greater continuous tuning range could allow for the series capacitor to remain at a fixed value, simplifying tuning and reducing the number of actively adjusted components.

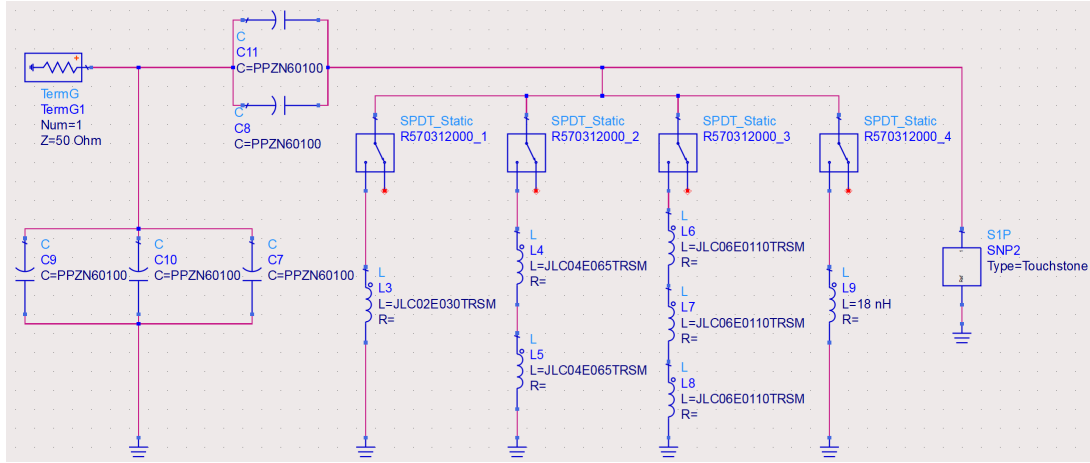


Figure 3.40: Schematic of novel network with components in place.

The circuit schematic presented above in Fig. 3.40 illustrates how the selected components were implemented into the design from Fig. 3.37, replacing the ideal elements with the real counterparts.

In this implementation, the shunt capacitor is formed by three PPZN60100 capacitors in parallel able to tune from 78 pF to 1800 pF, while the series capacitor consists of two PPZN60100 capacitors in parallel able to tune from 52 pF to 1200 pF.

The tunable inductor bank is realized through four Radiall R570312000 RF switches, each connected to a specific inductor or inductor combination, enabling discrete adjustments to the total inductance. The shunt inductor network is structured as follows: switch 1 connects to a small single-loop inductor formed by the 7 mm diameter loop of 18 AWG wire, switch 2 connects to a single JLC02E030TRSM inductor, switch 3 connects to a series combination of two JLC04E065TRSM inductors, and switch 4 connects to a series combination of three JLC06E0110TRSM inductors. This configuration provides a range of selectable inductance values, with a minimum and maximum values being calculated as 11.57 nH and 387.13 nH, respectively. While, due to the nature of this configuration of inductors, there are gaps in the achievable values within this range, the implementation of

the series capacitor with tunable components helps to mitigate the impact of these discontinuities.

3.4.4 Simulations with Novel Network Design

The analysis of the impedance matching network requires a comprehensive validation of its effectiveness in facilitating power transfer. As such, this section focuses on the validation of its performance through circuit co-simulation in HFSS. The objective is to systematically investigate the impact of the matching network on voltage distribution, electric field intensity, and power transfer.

Unlike the prior simulations in Sections 3.3.2 and 3.3.3 that modeled the system as a standalone HFSS structure excited directly by a terminal wave port, the co-simulation method integrates a circuit schematic with a network source and a connected HFSS model. This approach allows for analysis of how the network affects the behavior of the plates beyond simply showing a match to their complex impedance. The co-simulated setup enables further voltage and field calculations while still maintaining a source impedance and power across a frequency sweep consistent with previous HFSS simulations, ensuring that all comparisons between matched and unmatched cases remain meaningful.

To establish a baseline for comparison, the first simulation focuses on an unmatched case, without any impedance transformation, where the circuit design consists solely of a $50\ \Omega$ source of 100 W and 100 V_{pk} connected to the revised plate design used in Fig. 3.23. This circuit layout is shown below in Fig. 3.41.

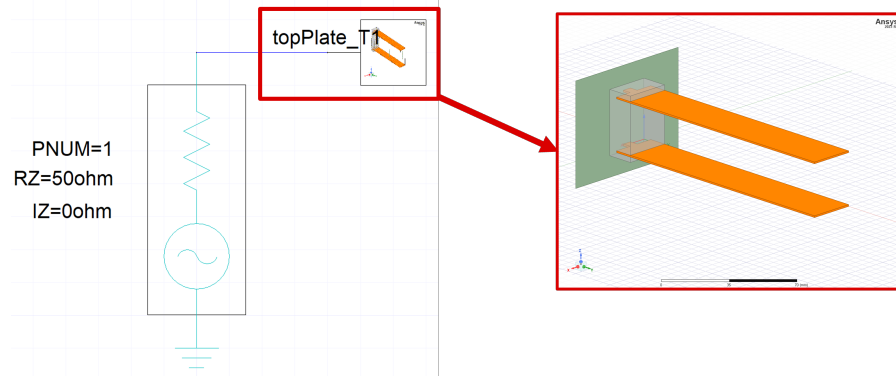


Figure 3.41: Circuit layout for unmatched co-simulation with HFSS. A source alone is first used while no matching network is applied, ensuring comparable analysis to unmatched simulation with HFSS alone.

After running the simulation, the voltage and fields across the plates can be reviewed. The voltage and field intensities from this setup help to verify that the results from this new unmatched simulation are in agreement with those obtained from previous HFSS-only setups. The average field intensity between the plates is calculated as before, using Eq. 3.7. The voltage between the plates in simulation can be computed by integrating the magnitude of the electric field in the full-volume between the plates as

$$V_{plates} = \left(\frac{\int_{Volume} |E| dV}{\int_{Volume} dV} \right) \times d_{plates} \quad (\text{Eq. 3.17})$$

where $|E|$ is the electric field within the full-volume and d_{plates} is the distance of separation between the plates. This approach aims to provide a voltage calculation that accurately reflects the potential difference that would be observed in an experimental measurement.

In the unmatched co-simulation, the average field intensity was found to be 7,186 V/m and the voltage between the plates was 214.1 V_{pk} at 50 MHz. Similarly, in the previous simulation of the revised plate design shown in Fig. 3.23, the average field intensity between the plates was found to be 7,170 V/m and the voltage was 213.1 V_{pk} at 50 MHz.

While there are minor variations between these two sets of results, the relative consistency confirms that the co-simulation method accurately represents the behavior observed

in prior simulations solely using HFSS and thus serves as a valid baseline for further analysis. Consequently, any significant changes observed in subsequent simulations, particularly when the matching network is introduced, can be confidently attributed to the impact of the matching network rather than from the simulation technique itself.

Following confirmation that the unmatched cases align, the new matching network pictured in Fig. 3.37 was added to the circuit design as pictured below in Fig. 3.42. Notably, the Q values for the selected components from the previous section were included to further ensure accurate simulation.

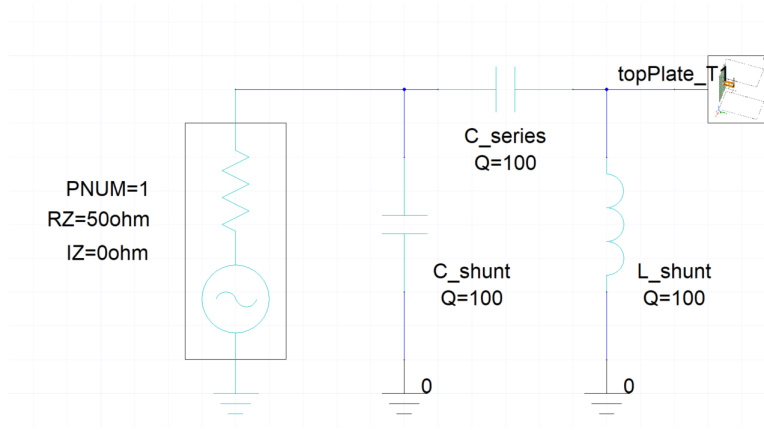


Figure 3.42: Circuit layout for co-simulation of the matching network, inserted between the source and the HFSS plate model to achieve an impedance match and analyze the simulated field response.

Matches were then found at the previously used benchmark frequencies of 20 MHz, 50 MHz, and 100 MHz. The component values used, the resulting voltages, and resulting average field intensities from these matches are detailed below in Table 3.6. It is important to emphasize that these simulations were performed without considering the practical limitations of real-world components or physical tolerances in fabrication. The objective is to evaluate how the network topology works to impact voltage and field enhancement rather than focus on specific implementation constraints, thus allowing for an assessment of how the matching network ultimately improves power transfer.

Frequency	20 MHz	50 MHz	100 MHz
C_{shunt}	370 pF	220 pF	135 pF
C_{series}	25 pF	25 pF	25 pF
L_{shunt}	2370 nH	394 nH	104 nH
$V_{pk,plates}$	1,412.2 V	1,055.9 V	700.6 V
$ E _{avg}$	49,501 V/m	35,196 V/m	23,355 V/m
$V_{pk,plates,unmatched}$	213.8 V	215.6 V	222.3 V
$ E _{avg,unmatched}$	7,126 V/m	7,186 V/m	7,409 V/m
$\Delta V_{pk,plates}$	$\times 6.9$	$\times 4.9$	$\times 3.2$
$\Delta E _{avg}$	$\times 6.9$	$\times 4.9$	$\times 3.2$

Table 3.6: Component values used to achieve an impedance match to empty plates at selected frequencies along with corresponding voltage and average field intensity. The component values used along with the voltages and fields from matched simulation are shown in the top section. Results from unmatched simulation of empty plates are also presented in the middle section. The increases in voltage and field intensity resulting from use of the matching network are presented in the bottom section. Calculations made before rounding.

The results presented above in Table 3.6 clearly demonstrate the impact of the matching network on voltage and field enhancement. Across all frequencies, the voltage across the plates increased significantly when the matching network was applied, with corresponding increases in the average field intensity.

At 50 MHz, for instance, the voltage increased from 215.6 V to 1,055.9 V, a factor of $\times 4.9$. Similarly, the field intensity increased from 7,186 V/m to 35,196 V/m, a comparable $\times 4.9$ increase. This proportionality between voltage and field strength confirms the expected behavior from Eq. 3.7, where an increased potential difference across the plates results in a proportionally stronger electric field. Furthermore, the results indicate a trend whereby the effectiveness of the matching network decreases slightly as frequency in-

creases. While the matching network still provides a substantial enhancement at 100 MHz, the improvement factor is reduced in comparison to the match at 20 MHz. These observations also help to confirm that increases in the voltage and field are attributed to the addition of the network before introducing the tissue sample, discussed below.

The next step involved introducing the tissue sample between the plates to determine power dissipation. However, unlike previous simulations with the tissue sample between the plates, the tissue sample was positioned where there is a small amount of contact with the bottom plate, a more realistic scenario taking into account that an object would likely be resting on the bottom plate. Except for the change in location of the tissue sample, a simulation identical to that shown in Fig. 3.30 was performed and the resulting fields are pictured below in Fig. 3.43.

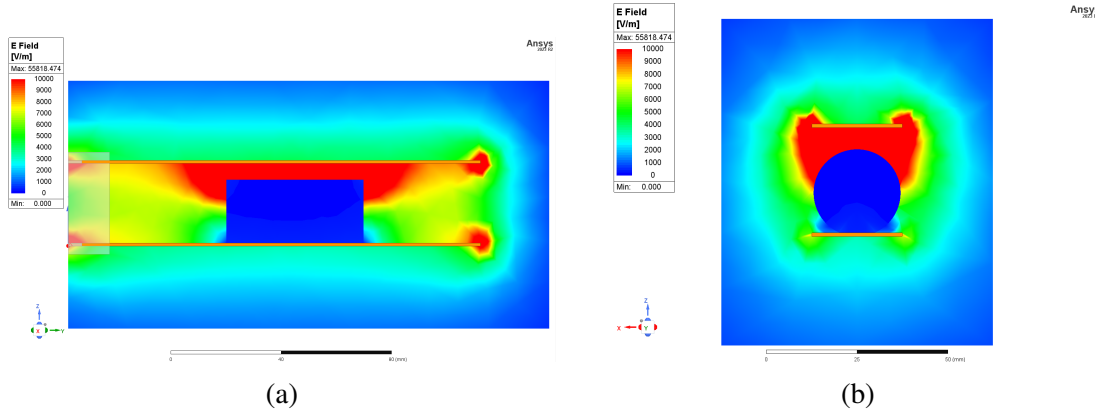


Figure 3.43: Electric field distribution between revised plate design with added tissue sample making contact with the bottom plate, fed with 100 V_{pk}, 50 MHz signal. Cross-sectional view in the YZ plane shown in (a), taken at the midpoint of the plate width. Cross-sectional view in the XZ plane shown in (b), taken at the midpoint of the plate length.

Similarly to the results seen in previous simulation with the tissue sample in Fig. 3.30, the results in the figure above show an increased field strength in the volume between the plates, especially in the region between sample and plates. Adding the same matching network layout from Fig. 3.42 alongside the HFSS design used in Fig. 3.43, another co-

simulation was carried out solely with a $50\ \Omega$ source of 100 W and 100 V_{pk} but with the addition of the tissue sample.

At 50 MHz, the unmatched co-simulation produced 236.3 V across the plates, an average field intensity of 7,878 V/m, and a total loss in the tissue of 0.5 W. Similarly, the unmatched simulation from Fig. 3.43 using HFSS alone produced 236.3 V across the plates, an average field intensity of 7,878 V/m, and a total loss in the tissue of 0.5 W. The strong agreement between these two simulation setups again supports that the co-simulation method accurately represents the behavior observed in prior simulations solely using HFSS and thus serves as a valid baseline for further analysis. Subsequent observations can be confidently attributed to the impact of the matching network rather than from the simulation technique itself.

Following confirmation that unmatched simulations with the tissue aligned, the matching network from Fig. 3.42 was added to the circuit design and matches were again found at frequencies of 20 MHz, 50 MHz, and 100 MHz. The results from these simulations are presented below in Table 3.7.

Frequency	20 MHz	50 MHz	100 MHz
C_{shunt}	156 pF	110 pF	73 pF
C_{series}	10 pF	10 pF	10 pF
L_{shunt}	4422 nH	724 nH	183 nH
$V_{pk,plates}$	2529.7 V	1,510.6 V	1,034.3 V
$ E _{avg,plates}$	84,322 V/m	50,352 V/m	34,475 V/m
$ E _{avg,tissue}$	1,845.7 V/m	2,090.0 V/m	1,943.8 V/m
P_{loss}	14.4 W	21.5 W	23.1 W
$V_{pk,plates,unmatched}$	233.9 V	236.3 V	246.4 V
$ E _{avg,plates,unmatched}$	7,797 V/m	7,878 V/m	8,215 V/m
$ E _{avg,tissue,unmatched}$	166.9 V/m	319.7 V/m	452.3 V/m
$P_{loss,unmatched}$	0.1 W	0.5 W	1.2 W
$\Delta V_{pk,plates}$	$\times 10.8$	$\times 6.4$	$\times 4.2$
$\Delta E _{avg,plates}$	$\times 10.8$	$\times 6.4$	$\times 4.2$
$\Delta E _{avg,tissue}$	$\times 11.1$	$\times 6.5$	$\times 4.3$
$(\Delta V_{pk,plates})^2$	$\times 116.7$	$\times 41.0$	$\times 17.6$
$(\Delta E _{avg,tissue})^2$	$\times 122.3$	$\times 42.7$	$\times 18.5$
ΔP_{loss}	$\times 123.1$	$\times 42.8$	$\times 18.5$

Table 3.7: Component values used to achieve an impedance match to plates with tissue sample at selected frequencies along with corresponding voltage and average field intensity. The component values used along with the voltage, field, and power values from matched simulation are shown in the top section. Results from unmatched simulation are also presented in the middle section. The increases in voltage, field intensity, and power loss resulting from use of the matching network are presented in the bottom section. Calculations made before rounding.

As expected, the matching network results in an increase in the voltage across the plates

and field intensity between the plates. However, the most critical observation in this case is that the power loss in the tissue scales closely to the square of the factor of voltage increase, a reflection of the fundamental relationship between power dissipation and electric potential. This relationship is essential in proving that the matching network is responsible for enhancing power transfer and establishes a strong correlation between an increased voltage across the plates (and thus an increased field intensity) and the energy absorbed by the tissue.

To further strengthen the analysis, another set of simulations were carried out, incorporating the frequency-dependent characteristics of the fabricated plates and spacer shown in Fig. 3.25 by replacing the co-simulated HFSS plates with the measured S-parameters of the physical plates as shown below in Fig. 3.44.

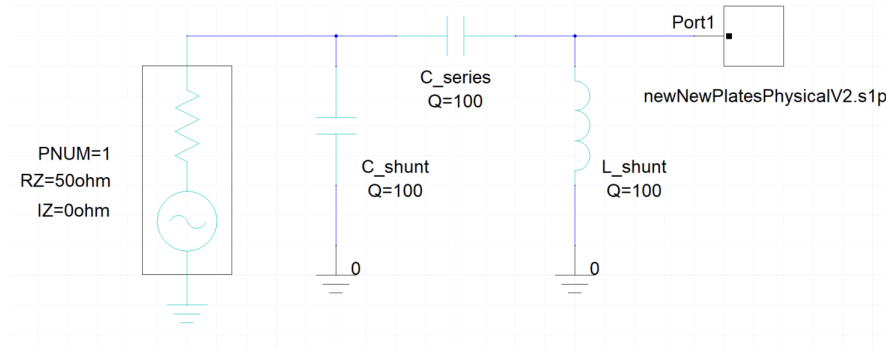


Figure 3.44: Circuit setup for validating matching performance against the fabricated plate system. The physical S-parameter response of the plates was imported as the load, allowing evaluation of how well the idealized matching network could match to the real-world fabricated system and predict experimental behavior.

Again, the matching network was tuned to achieve matches at 20 MHz, 50 MHz, and 100 MHz where the resulting voltage across the plates is analyzed. The voltage increase over the unmatched case in these simulations, as shown below in Table 3.8, follows the same general trend observed from co-simulations with the simulated plates from HFSS, strongly reinforcing that the fundamental behavior of the matching network holds true in

practical implementations.

Frequency	20 MHz	50 MHz	100 MHz
C_{shunt}	175 pF	74 pF	56 pF
C_{series}	10 pF	10 pF	10 pF
L_{shunt}	4140 nH	690 nH	175 nH
$V_{pk,plates}$	2,410.7 V	1,023.3 V	708.0 V
$V_{pk,plates,unmatched}$	199.9 V	198.3 V	195.2 V
$\Delta V_{pk,plates}$	$\times 12.1$	$\times 5.2$	$\times 3.6$
$\Delta P_{loss,expected} \approx (\Delta V_{pk,plates})^2$	$\times 145.4$	$\times 26.6$	$\times 13.2$

Table 3.8: Component values used along with the voltage values from matched simulation are shown in the top section. Results from unmatched simulation are also presented in the middle section. The increases in voltage expected and power loss resulting from use of the matching network are presented in the bottom section. Calculations made before rounding.

While field and loss calculations are unavailable in this approach due to the absence of a corresponding fully modeled HFSS volume, the fact that similar voltage scaling behavior is observed when using the S-parameter-based model of the physical plates demonstrates that the expected power transfer to the tissue, as calculated in Table 3.8, follows suit. This trend is further validated through calculated results presented in the next chapter.

The strong agreement between the simulated and physical plate responses further reinforces the real-world applicability of these results. In turn, the observations made from the simulations in this section validate the role of the impedance matching network in enhancing power transfer to the material between the plates, in this case, the tissue sample.

With a well-characterized network topology and demonstrated improvements in transferring losses to the load, the next step begins the transition from simulation to experimental results. In the following section, the focus shifts towards fabrication of the matching network and integrating a variety of real components into a system for experimental use.

3.5 Summary

This chapter detailed the design, modeling, and simulation efforts that form the backbone of the RF system developed in this thesis. Beginning with the analysis of CNT behavior under RF excitation in Section 3.2, simulations in HFSS revealed the dipole-like behavior of CNTs, with oscillatory electric fields and charge distributions. These results provided confirmation into the observed electromagnetic interactions of CNTs, supporting their potential use in targeted heating applications and informing the system design.

In Section 3.3, the development of the parallel plates was explored, starting with modifications to the original Comdel plate design. Simulations demonstrated that the refined plate geometry resulted in improved field containment and reduced field presence radiating outside the plates, addressing the limitations of previous designs. These refinements ensured that the plates could serve as effective terminations for energy delivery while maintaining a high degree of specificity in energy delivery.

Section 3.4 focused on the development of a tunable matching network, which was designed to address the limitations of fixed-frequency commercial solutions such as the Comdel MatchPro network used previously, enabling impedance matching across a broad frequency range. Simulations in ADS and HFSS validated the matching network's ability to adapt to varying complex impedances, a critical feature for accommodating evolving material properties. Component selection was guided by the need to balance tunability, performance, and practicality, while ensuring compatibility with the broader system. Additionally, simulations incorporating a human tissue analog demonstrated the network's ability to effectively operate and match with materials under test, further validating its design.

Finally, Section 3.5 detailed the physical implementation of the system, including the fabrication and testing of the matching network, integration of the RF amplifier and power

supplies, and incorporation of monitoring components such as the power check circuit and directional coupler. These efforts transitioned the system from simulation to practical assembly, culminating in a system capable of experimental evaluation.

Together, the modeling, design, and fabrication efforts described in this chapter represent a comprehensive approach to developing the RF system. This work bridges theoretical simulations with practical implementation, setting the stage for the experimental validation and performance analysis presented in the next chapter.

Chapter 4

System Assembly and Testing

The assembly of the RF system involved integrating all major components, including the matching network, plates, power supply, and measurement tools, into a functional setup. Ultimately, the system will be laid out as shown below in Fig. 4.1, which details the interconnections and spatial arrangement of components. This diagram can be referred back to for clarity on the system as described in following sections. Note that the VNA has been used routinely used to take S-parameter measurements at low power (≤ 10 dBm), it is left out of Fig. 4.1 below and will not be used at any point alongside the amplifier. When the amplifier is employed later in Section 5.2, it is only ever fed from the signal generator.

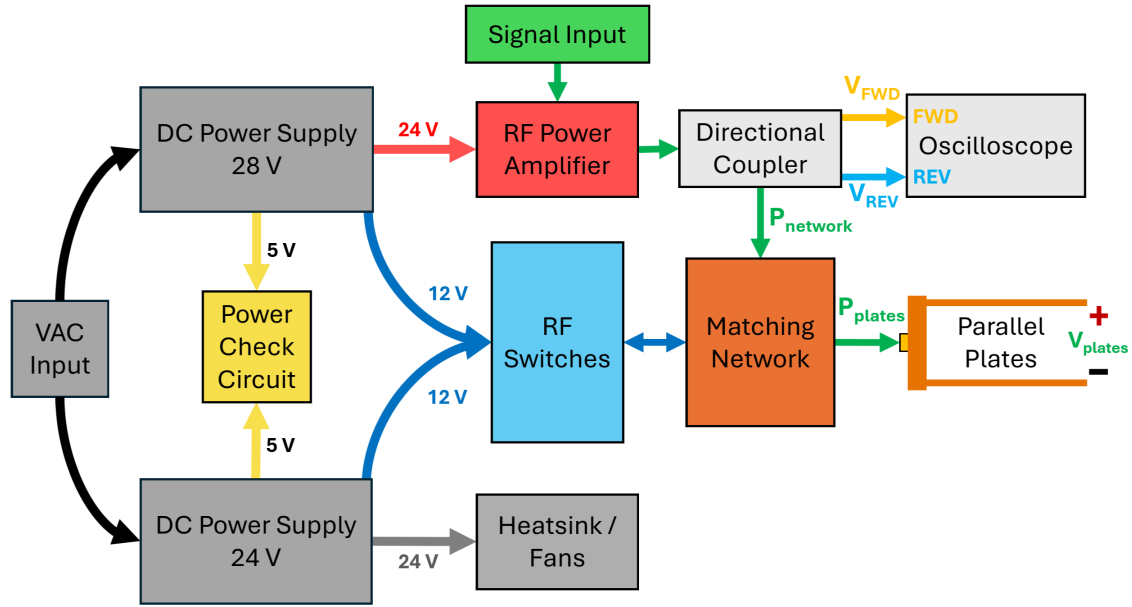


Figure 4.1: System-level block diagram showing the complete experimental setup. RF signal is fed through the matching network and delivered to the parallel plates, while directional coupler measurements provide real-time monitoring of forward and reflected power. Supporting DC power supplies are connected to the amplifier, RF switches, fans to ensure proper operation.

4.1 Matching Network Fabrication

Most RF components, including signal generators, amplifiers, directional couplers, and VNAs, are standardized to operate at $50\ \Omega$ [13]. As such, it's practical to design the network's microstrip transmission lines on the PCB for $50\ \Omega$ to ensure compatibility with ancillary system components discussed later in Chapter 4.

The trace width required for a line impedance of $50\ \Omega$ was found to be roughly 3 mm (118 mil) by using

$$Z_0 = \frac{87}{\sqrt{\epsilon_r + 1.41}} \ln \left(\frac{5.98h}{0.8w + t} \right) \quad (\text{Eq. 4.1})$$

where the desired Z_0 as well as board properties such as ϵ_r , substrate thickness, h , and copper thickness, t , can be substituted to find the required trace width, w [42]. Based on

the information provided from [43] for the RO4350B laminate selected for use, $\epsilon_r = 3.48$, $h = 60$ mil, and $t = 17.5$ μm .

To ensure the microstrip traces can safely handle the expected current, the maximum allowable current to keep the temperature under a limit can be found by

$$I = (k)(\Delta T^{0.44})(A^{0.725}) = (k)(\Delta T^{0.44})(A^{0.725}) \quad (\text{Eq. 4.2})$$

where $k = 0.048$ is a constant given in [44], ΔT ($^{\circ}\text{C}$) is the change in temperature, and A (mil^2) is the cross sectional area of the trace. Based on the information given in [43], the temperature of the 4350B laminate should remain below 280°C ; however, to ensure proper operation, the maximum operating temperature of 85°C listed in [39] for the PPZN60100 capacitors connected to the primary trace was used as the desired maximum for the entire PCB. With a conservative ambient temperature of 25°C , the maximum allowable change in temperature would be $\Delta T = 60^{\circ}\text{C}$. Given this value for ΔT and the previously discussed values for t and w , the maximum allowable current for the trace can be found by

$$I = (0.048)(60^{0.44})((118)(0.69))^{0.725} = 7.05A \quad (\text{Eq. 4.3})$$

Given the maximum power from the source discussed within the scope of this work of 100 W with a port impedance of 50Ω , Ohm's law would give an RMS voltage, V_{rms} , of 70.71 V and a current, I_{rms} , of 1.41 A. The calculation from Eq. 4.3 shows a that the trace can handle up 7.05 A of current, a considerable amount more than the 1.41 A expected during operation.

Furthermore, Eq. 4.2 can be used to calculate the expected increase in temperature with a current of 1.41 A by

$$1.41 = (k)(\Delta T^{0.44})(((118)(0.69))^{0.725})^{0.725} \longrightarrow \Delta T = 1.55^{\circ}\text{C} \quad (\text{Eq. 4.4})$$

This result indicates that under the maximum expected operating conditions, the calculated temperature rise (1.55°C) is not consequential.

In order to test the matching network design with the components described in Section 3.4.3, a PCB to mount everything together had to first be created. Component footprints were designed for each component laid out in Fig. 3.40 using the available dimensions found in data sheets. Using these component footprints, a board design using Roger's RO4350B was put together in ADS' layout view.

The finalized board design is shown below in Fig. 4.2 along with each of the four boards for the inductors in Fig. 4.3. Having four standalone boards for the shunt inductance in the system due to a desire to minimize any potential interference or other effects that could arise from having the whole network on a contiguous PCB.

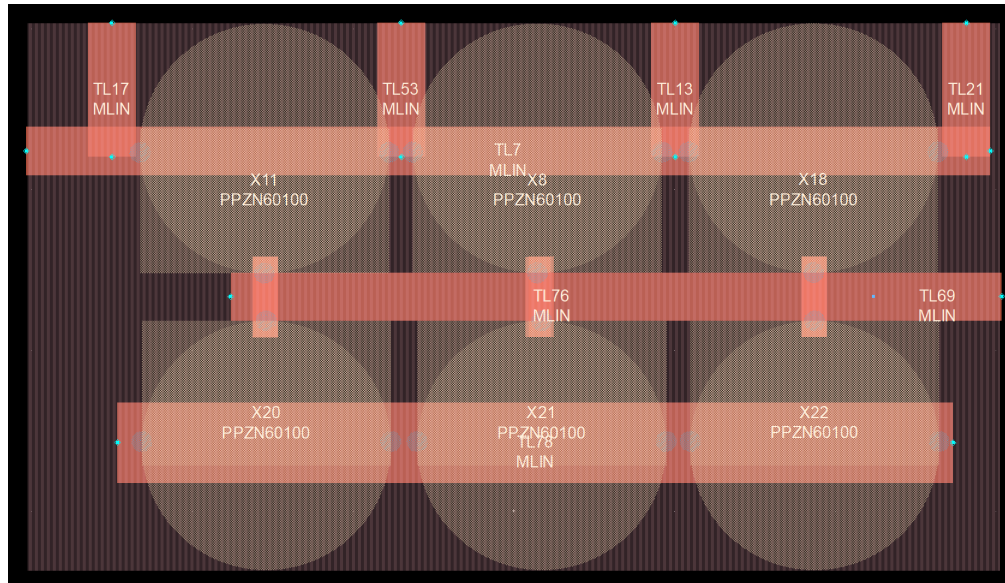


Figure 4.2: ADS layout of the main matching network PCB. Placement of tunable capacitors and interconnecting microstrip transmission lines are shown. This layout was later fabricated on Rogers 4350B substrate using an LPKF protolaser system.

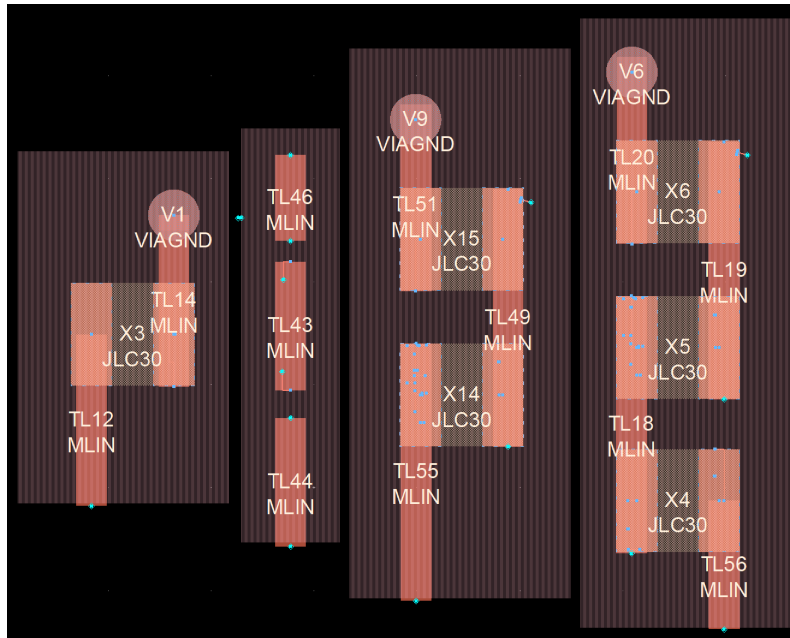


Figure 4.3: ADS layout of the four inductor matching network PCBs. Placement of tunable inductors and interconnecting microstrip transmission lines are shown. This layout was later fabricated on Rogers 4350B substrate using an LPKF protolaser system.

The finalized board layout was exported as a set of gerber files and imported into an LPKF ProtoLaser to be fabricated. The fabricated board is pictured below in Fig. 4.4 and the board populated with components and edge mount SMA connectors is also pictured below, in Fig. 4.5.

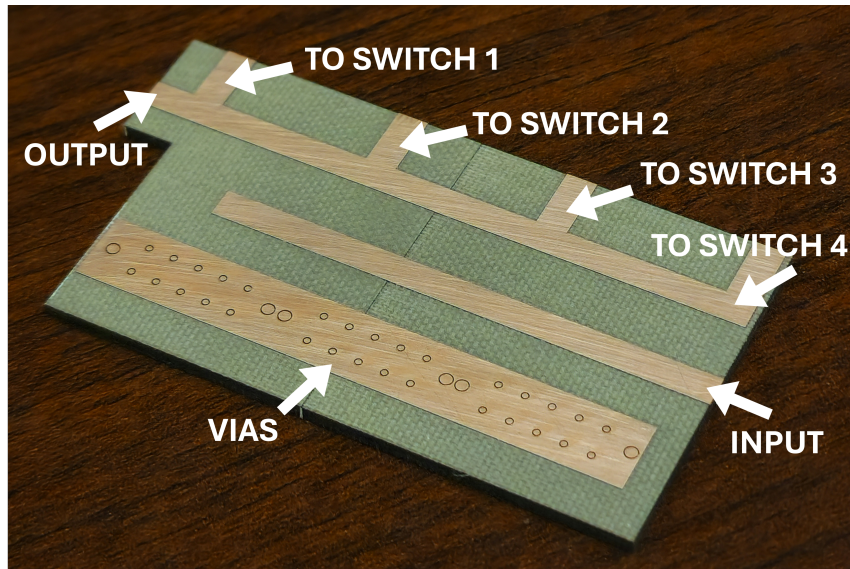


Figure 4.4: Fabricated matching network PCB on Rogers 4350B substrate. Labeled traces show the planned input, output, and connections to each of the four RF switches controlling the tunable inductors. Grounding vias were incorporated to improve grounding and minimize parasitic effects across the board.

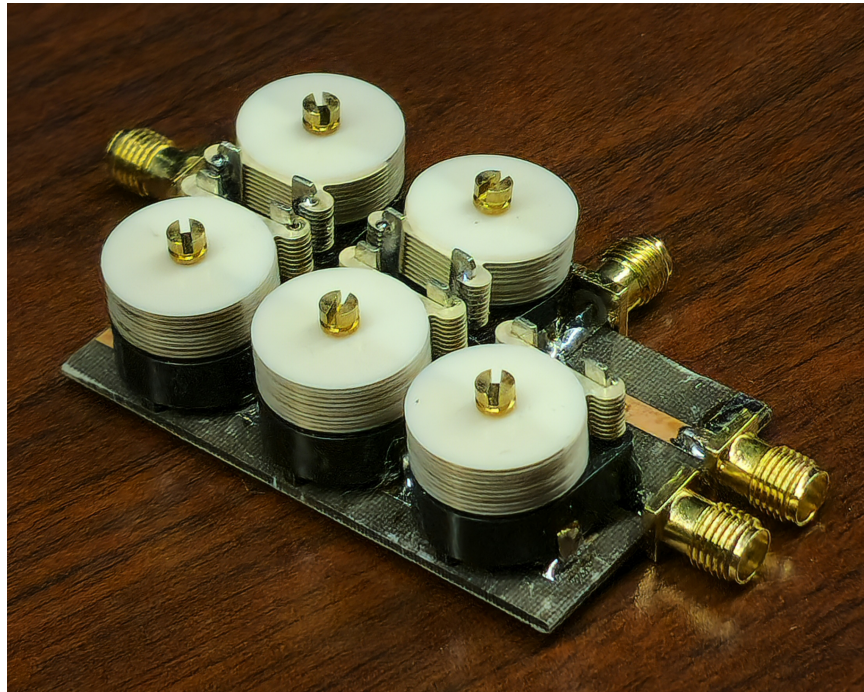


Figure 4.5: Fully assembled matching network PCB populated with tunable capacitors and SMA connectors.

Careful attention was given to the assembly process to minimize parasitic effects and ensure consistency with the simulated design. For instance, multiple vias pictured above in Fig. 4.4 were incorporated between the pad connected to the shunt capacitors and the ground plane to reduce inductance and mitigate unwanted effects introduced by the PCB itself. With the network assembled, the next step involved testing to validate its ability to achieve an adequate impedance matching across the target frequency range, as detailed in the following section.

4.2 Matching Network Testing

The matching network was connected to the parallel plates, which served as the load, and a VNA, acting as the source. To achieve a match, the various inductor boards were manually connected or disconnected from the main board as needed, while the capacitors were adjusted accordingly. This tuning process was verified through confirmation on the VNA's Smith Chart at frequencies of 20 MHz, 50 MHz, and 100 MHz, preliminarily demonstrating the ability of the network to create a match. While the VNA still swept over the same frequency range from 10 MHz to 100 MHz used in previous simulation and testing, 20 MHz was used as the lowest frequency to create a match. This is due to the amplifier's minimum operation frequency of 20 MHz, described in Section 4.3 below. Smith Chart plots of each of these matches are shown below in Fig. 4.6.

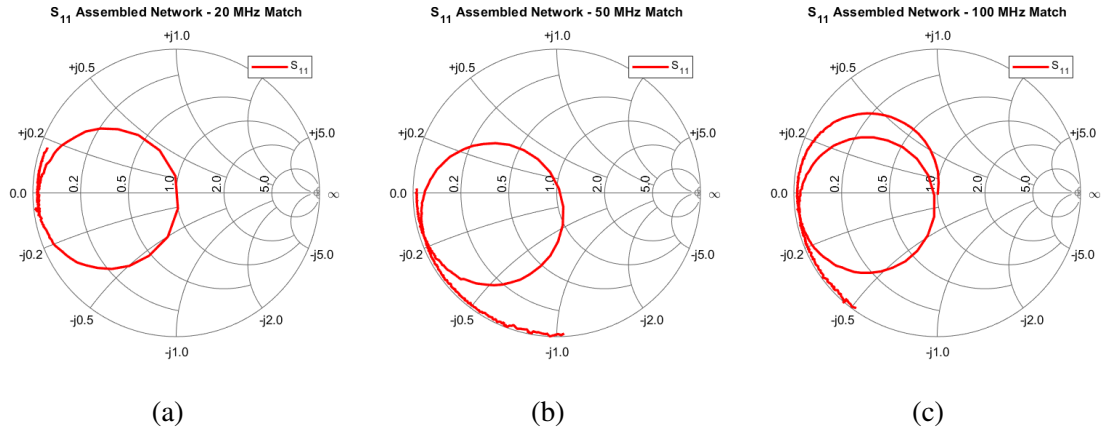


Figure 4.6: Assembled network with matches at (a) 20 MHz, (b) 50 MHz, (c) 100 MHz. Frequency swept from 20 MHz to 100 MHz in each figure.

Manually connecting and disconnecting the inductors during this phase ensured that the results were not influenced by any unintended effects introduced by the RF switches. Once this baseline was confirmed, the same process was repeated with the RF switches used to automate the connection and disconnection of the inductors for the same set of frequencies. Smith Chart plots of each of these matches are shown below in Fig. 4.7. Although the switches do have an effect on the results, it appears to be very minimal and indicates that they do not significantly affect the network's ability to match.

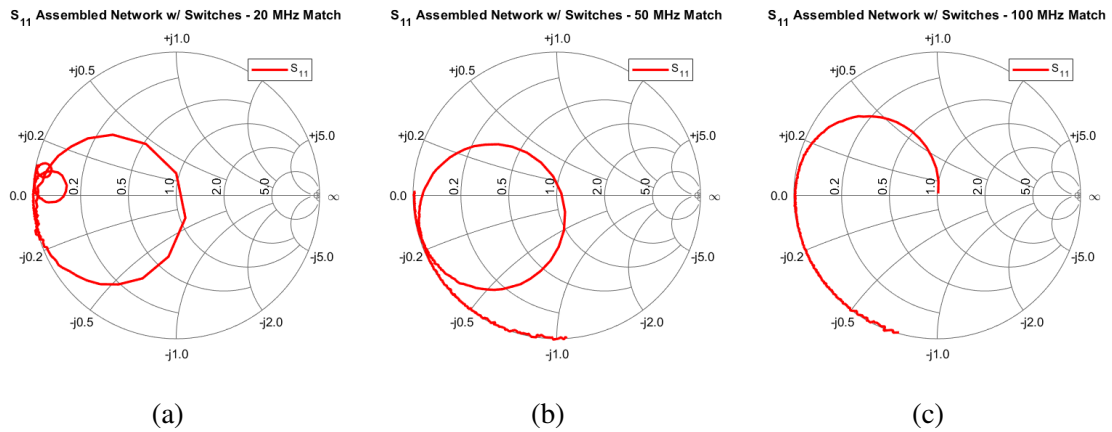


Figure 4.7: Assembled network with matches using RF switches at (a) 20 MHz, (b) 50 MHz, (c) 100 MHz. Frequency swept from 20 MHz to 100 MHz in each figure.

The testing of the matching network demonstrated its ability to adaptively match a range of impedances across the target frequency band, validating the design's effectiveness and consistency with simulation. With the matching network's functionality preliminarily confirmed, attention was turned to the integration of ancillary components such as the RF amplifier and power supplies, as discussed in the following section.

4.3 RF Amplifier and Power Supplies

As mentioned in Section 1.2 and Section 3.2, the target maximum power is 100 W; however, the signal generator used to feed the system is not able to deliver this amount of power. To address this, a Pasternak PE15A5065 high-power amplifier was integrated. This amplifier provides up to 100 W of output power across a frequency range of 20 MHz to 520 MHz, with a gain of 51 dB [45]. To maintain safe operating temperatures, the amplifier was mounted on a heatsink equipped with two fans. Notably, a VNA used in the initial tests carried out in Chapter 5 has a maximum output of only 10 dBm, or 0.01 W.

In order to power the amplifier and the heatsink fans, two Astrodyne ASM550 DC power supplies were used. One of the power supplies has a 28 V output for the amplifier while the other power supply has a 24 V output to power the fans [46]. Both power supplies have auxiliary 12 V, 500 mA outputs, used to power the RF switches [46]. With the RF switches requiring 250 mA of current to operate according to [40] and each of the 12 V outputs have a current rating of 500 mA, each power supply was connected to two of the four total switches.

4.4 Power Check Circuit

To monitor the power supplies, a 'power check' circuit was designed using an LM293 comparator IC, BS170 MOSFETs, resistors, and LEDs. From each power supply there

is a 'PowerOK' line and a 'PS ON' control line [46]. The PowerOK line outputs a logic high of >3.3 V when the output is enabled and functioning correctly, whereas a logic low of <3.3 V is output when disabled or if an error is detected [46]. Measurement with a handheld multimeter found that the logic high operated at just under 5 V, while the logic low operated very near to 0 V. The PS ON line controls the state of the power supply's outputs. When this line is connected to the 5 V ground, both the main 28 V and 24 V outputs as well as the auxiliary 12 V outputs become enabled according to the information found in [46].

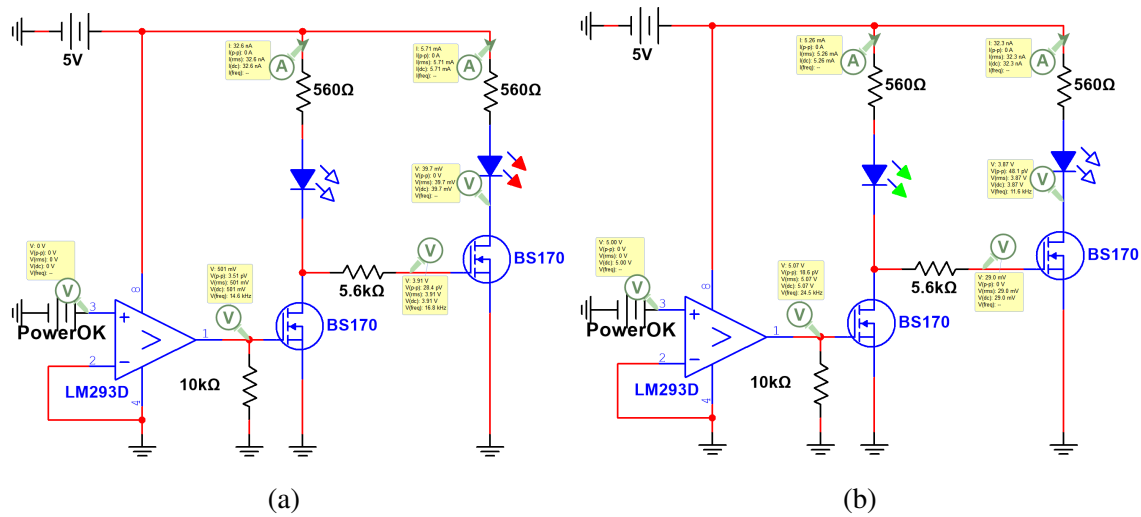


Figure 4.8: Power check circuit where (a) PowerOK = 0 V, (b) PowerOK = 5 V.

The monitoring circuit shown above in Fig. 4.8, simulated in Multisim, helps to provide a clear visual indication of the status of the power supply. Initially, when the power supply receives any AC power, a red LED is illuminated, indicating that while the power supply is receiving power, the PS ON line has not yet been connected to ground and the outputs are disabled. When a switch is flipped to connect the PS ON line to ground and no issues are detected, the red LED is turned off and a green LED illuminated to confirm that the outputs were enabled and functioning correctly. If the switch is flipped to connect the PS ON line to ground and there is an error, the red LED will remain on while the green LED

will remain off. Fig. 4.9 shows the circuit assembled on a breadboard.

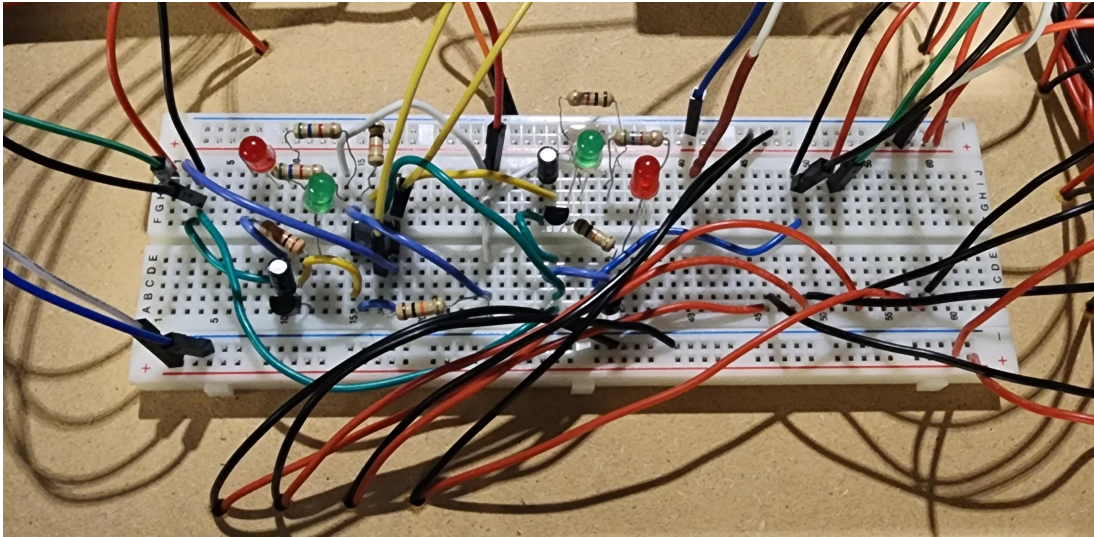


Figure 4.9: Power check circuit constructed on a breadboard. The circuit monitors DC supply voltages and provides visual status indication via LEDs to help observe system operation.

The integration of the RF amplifier and power supplies ensured the system could deliver sufficient power to drive the matching network and maintain stable operation across the target frequency range. Additionally, the power check circuit served as a method for monitoring the power supplies, ensuring their healthy operation and stability during system use. These ancillary system components were mounted on a board of MDF and are shown in Fig. 4.10 below.

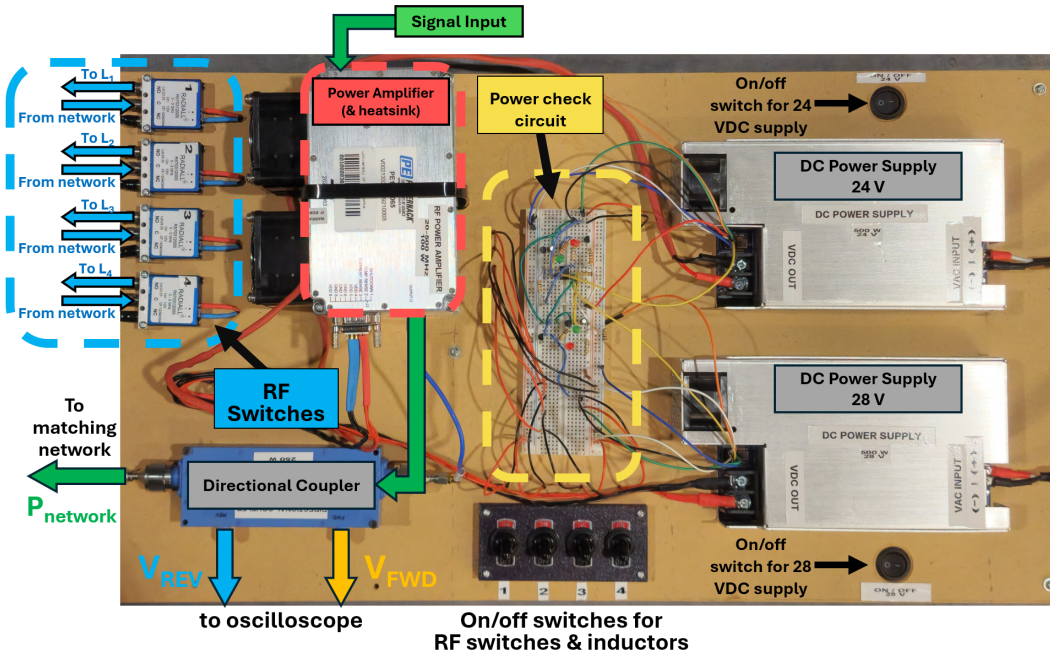


Figure 4.10: Layout of full system mounted on board of MDF. Key components include DC power supplies, RF power amplifier with heatsink, RF switches, directional coupler, custom power check circuit, and the manual switch panel for controlling RF paths and inductor selection. The integrated setup enables controlled RF signal delivery through the tunable matching network to the parallel plates.

With an established method of power delivery into the system, the next step involved the addition of a directional coupler and oscilloscope to aid in gathering detailed measurements, discussed in the following section.

4.5 Directional Coupler and Oscilloscope

In order to measure the network and system performance during testing, a directional coupler was positioned between the amplifier and the matching network. This directional coupler, a Pulsar C40-111-481/5N, with an insertion loss of 0.4 dB and a coupling factor of 40 dB, operates over a frequency range from 0.1 MHz to 250 MHz with a source power of up to 250 W, making it more than adequate for the requirements of this system [47].

Additionally, the coupler's minimum directivity of 20 dB ensured isolation of forward and reflected signal lines [47]. The forward and reverse lines of the directional coupler, with the aforementioned 40 dB coupling factor, were connected to an oscilloscope to take voltage measurements.

High-impedance probes with an 100x attenuation factor were also connected to the oscilloscope and used to measure the voltage across the parallel plates. These measurements, providing a comparison to the forward voltage from the directional coupler's forward line, help to characterize the performance of the matching network as well as help in verifying proper impedance matching at a given target frequency. Lastly, the auxiliary trigger output of the VNA was connected to the oscilloscope, helping with data acquisition shown in the following sections.

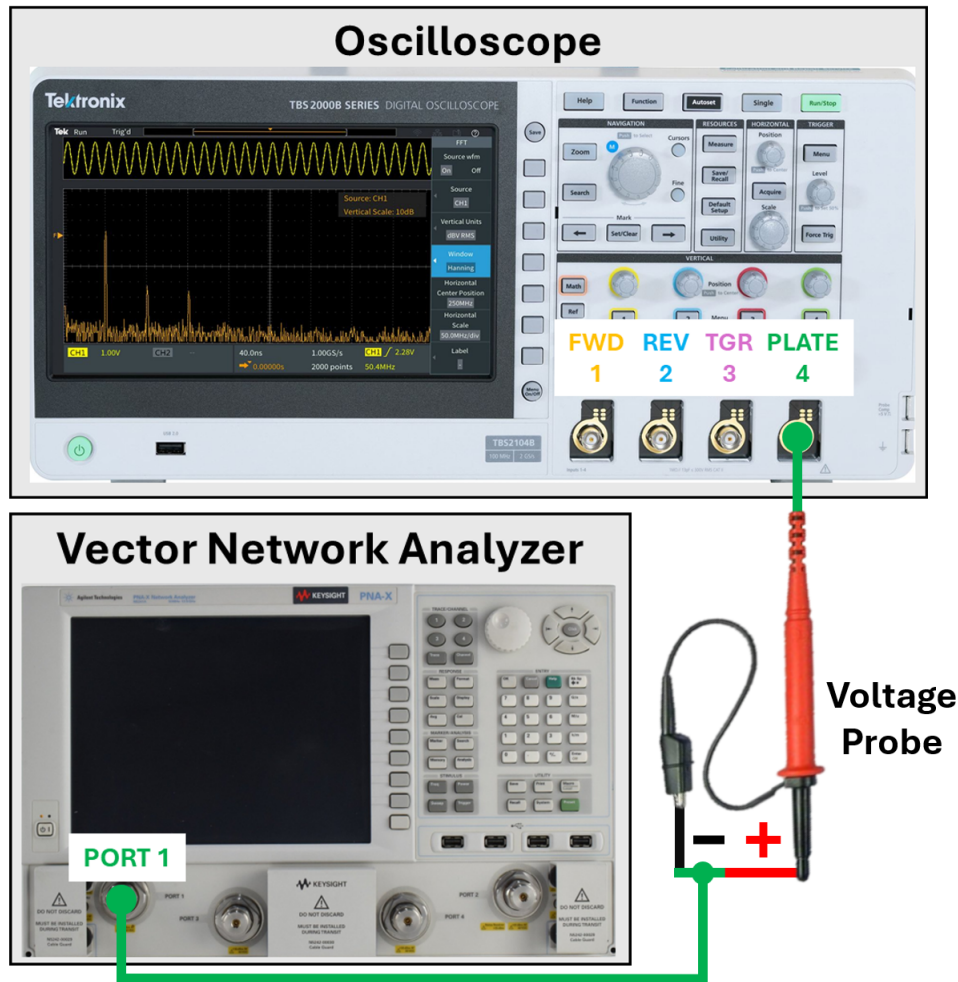


Figure 4.11: Layout showing how a VNA was used to measure the S-parameters of the voltage probes and oscilloscope connection, allowing for an understanding of their loading effects when connected to the parallel plates.

However, the oscilloscope and associated probes introduce unavoidable modification to the impedance, affecting voltage measurements across the plates. While these effects must be considered, they do not invalidate the overall trends observed in the simulations and experimental results. To quantify the influence of the probes, the S-parameters of the probes and oscilloscope were measured as depicted above in Fig. 4.11 and integrated into the HFSS co-simulation circuit as pictured below in Fig. 4.12, allowing for a representation of their effect on the system.

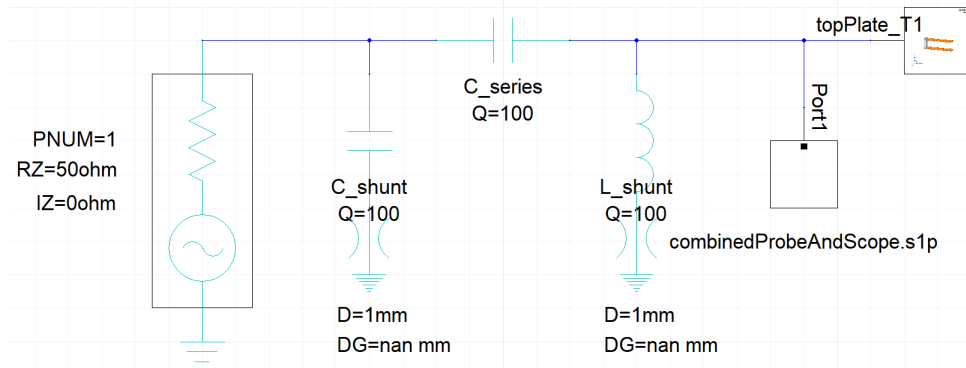


Figure 4.12: Circuit layout for matched simulation with measured S-parameter data of oscilloscope and voltage probes connected in parallel, used to understand effect of probes on simulated response.

As discussed previously in Section 3.4.4, in the matched condition without the probes, the voltage across the plates increased significantly compared to the unmatched case, detailed in Table 3.6, demonstrating the effectiveness of the impedance matching network. When the probes were included in the simulation, the voltage across the plates still increased but to a lesser extent. The probes were introduced and matches were again found at 20 MHz, 50 MHz, and 100 MHz. The results of these simulations are shown below in Table 4.1.

Frequency	20 MHz	50 MHz	100 MHz
C_{shunt}	125 pF	45 pF	28 pF
C_{series}	200 pF	200 pF	200 pF
L_{shunt}	386 nH	150 nH	63.5 nH
$V_{pk,plates}$	178.9 V	125.9 V	115.2 V
$V_{pk,plates,unmatched}$	148.3 V	123.4 V	116.4 V
$\Delta V_{pk,plates}$	$\times 1.2$	$\times 1.0$	$\times 0.9$

Table 4.1: Component values used to achieve an impedance match to empty plates at selected frequencies along with corresponding voltage and average field intensity when oscilloscope and probes were connected in parallel. Results from unmatched simulation of empty plates with oscilloscope and probes connected in parallel also presented and used to show increases in voltage and field intensity resulting from use of the matching network.

The factor of voltage increase across the plates was reduced to 1.2, 1.0, and 0.9, at 20 MHz, 50 MHz, and 100 MHz, respectively. This reduction can be attributed to the additional impedance introduced by the oscilloscope and probes, altering the impedance seen by the network and subsequently affecting the voltage measured across the plates. Despite this reduction in the magnitude of the observed voltage increase, the matching network still showed an increase in the voltage across the plates in these simulations and thus remains a meaningful method for validating network behavior.

In the context of this work, specifically in verifying matches in an experimental setup as seen in Section 5.1, observing a peak in the measured voltage coinciding with a normalized impedance of $Z_n = 1 + j0$ at the center of the Smith Chart serves as a key indicator that a proper match has been achieved. While the oscilloscope probes introduce an additional impedance into the system, affecting the component values required for matching, the underlying process of tuning the network remains the same.

Notably, in an actual system, an effective approach to implementing the network would

involve first estimating approximate component values through simulation-based design. Once assembled, the network can be fine-tuned at low power using only the Smith Chart, S-parameter, and impedance response, ensuring a match without the need for direct voltage measurements across the plates. Following this process, the final system can operate without probes, preventing any interference induced by the use of probes/oscilloscope and maintaining more optimal power transfer.

4.6 Summary

This chapter detailed the design, modeling, and simulation efforts that form the backbone of the RF system developed in this thesis. Beginning with the analysis of CNT behavior under RF excitation in Section 3.2, simulations in HFSS revealed the dipole-like behavior of CNTs, with oscillatory electric fields and charge distributions. These results provided confirmation into the observed electromagnetic interactions of CNTs, supporting their potential use in targeted heating applications and informing the system design.

In Section 3.3, the development of the parallel plates was explored, starting with modifications to the original Comdel plate design. Simulations demonstrated that the refined plate geometry resulted in improved field containment and reduced field presence radiating outside the plates, addressing the limitations of previous designs. These refinements ensured that the plates could serve as effective terminations for energy delivery while maintaining a high degree of specificity in energy delivery.

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material properties. Component selection was guided by the need to balance tunability, performance, and practicality, while ensuring compatibility with the broader system. Additionally, simulations incorporating a human tissue analog demonstrated the network's ability to effectively operate and match with materials under test, further validating its design.

Finally, Section 3.5 detailed the physical implementation of the system, including the fabrication and testing of the matching network, integration of the RF amplifier and power supplies, and incorporation of monitoring components such as the power check circuit and directional coupler. These efforts transitioned the system from simulation to practical assembly, culminating in a system capable of experimental evaluation.

Together, the modeling, design, and fabrication efforts described in this chapter represent a comprehensive approach to developing the RF system. This work bridges theoretical simulations with practical implementation, setting the stage for the experimental validation and performance analysis presented in the next chapter.

Chapter 5

Results and Discussion

This chapter presents the experimental evaluation of the RF system described in earlier chapters. Building on the simulations and design considerations previously established, this section demonstrates the real-world performance of the tunable matching network and parallel plate system when applied to a tissue analog. Impedance matching is first assessed through a series of matching tests, verified through measurements of S_{11} in conjunction with voltage waveforms captured on the oscilloscope.

S-parameter data collected from the physical network and plates is then analyzed to determine voltage amplification and power transfer, following the process first introduced Section 2.3.2. These results are used to validate the system's performance and highlight how practical component constraints affect behavior. Together, these findings reinforce the system's effectiveness while providing insight into both its capabilities and limitations.

5.1 Initial Testing

5.1.1 Baseline Measurements

To establish a reference for later comparisons, measurements were made for an initial low-power baseline case. The plates were directly connected to the output of the directional

coupler, while the input of the coupler was fed by a VNA, serving as the source. A layout of this setup is shown in Fig. 5.1. It is worth noting here that in all subsequent tests, the VNA and amplifier are never used in the system at the same together.

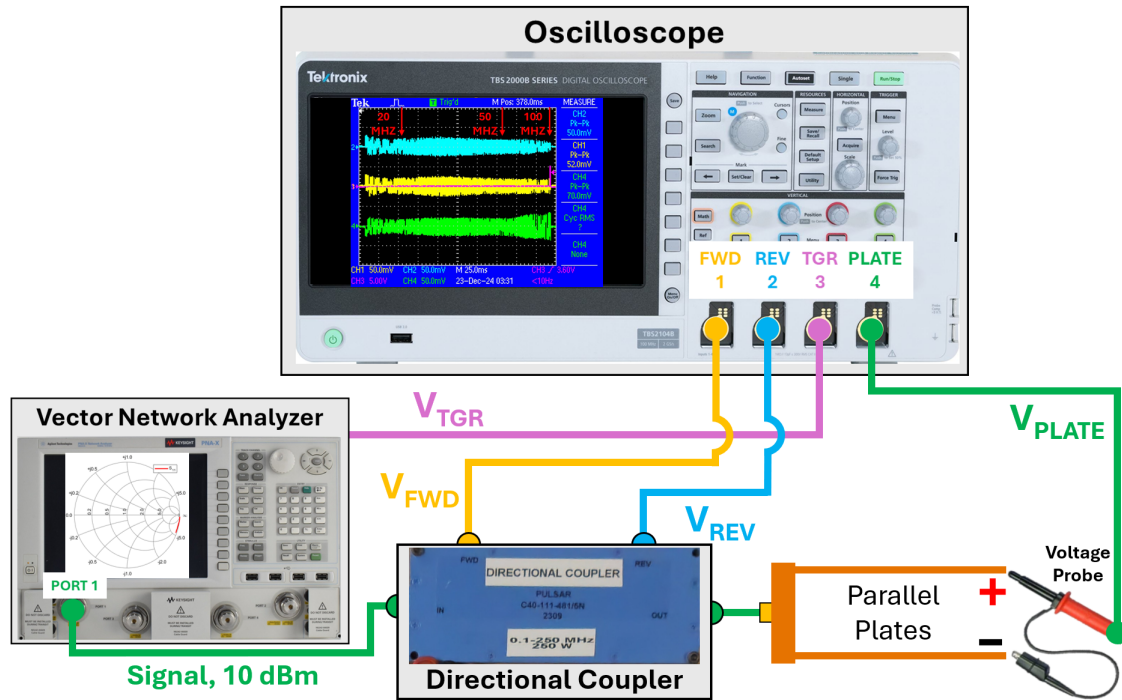


Figure 5.1: Layout of setup used for obtaining an unmatched baseline measurement, where the signal was fed through a directional coupler before exciting the parallel plates, allowing measurement of forward, reflected, and plate voltages across frequency.

The VNA was configured to output 10 dBm and sweep from 10 MHz to 100 MHz on port 1, fed into the input of the directional coupler. With the output of the directional coupler bypassing the matching network and being connected directly to the plates, the oscilloscope captured the data shown in Fig. 5.2. With the plates connected alone, the measured voltage across them serves as a reference for an unmatched state.

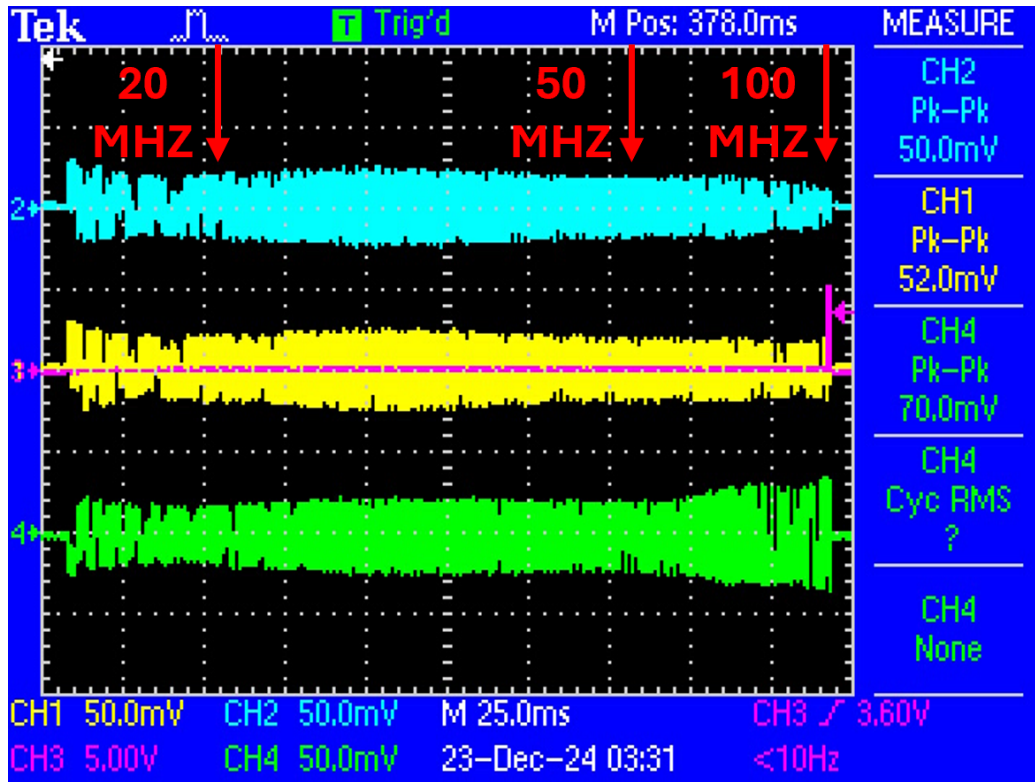


Figure 5.2: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep at 10 dBm. Directional coupler output connected directly to plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

The maximum plate voltage (green) across the sweep, at 100 MHz, is $70.0 \text{ mV}_{\text{pkpk}}$ ($24.8 \text{ mV}_{\text{rms}}$). The voltage across the plates remains below this value for the majority of the sweep. At 20 MHz, a value of approximately $40 \text{ mV}_{\text{pkpk}}$ ($14.1 \text{ mV}_{\text{rms}}$) was observed while at 50 MHz, a value of approximately $50 \text{ mV}_{\text{pkpk}}$ ($17.7 \text{ mV}_{\text{rms}}$) was observed. To reiterate, this configuration does not represent an optimized or tuned system but rather provides a baseline to determine whether proper impedance matching results in an increase in measured voltage across the plates.

In a second setup, the matching network was added to Fig. 5.1 between the plates and the output of the directional coupler, laid out in Fig. 5.5 below.

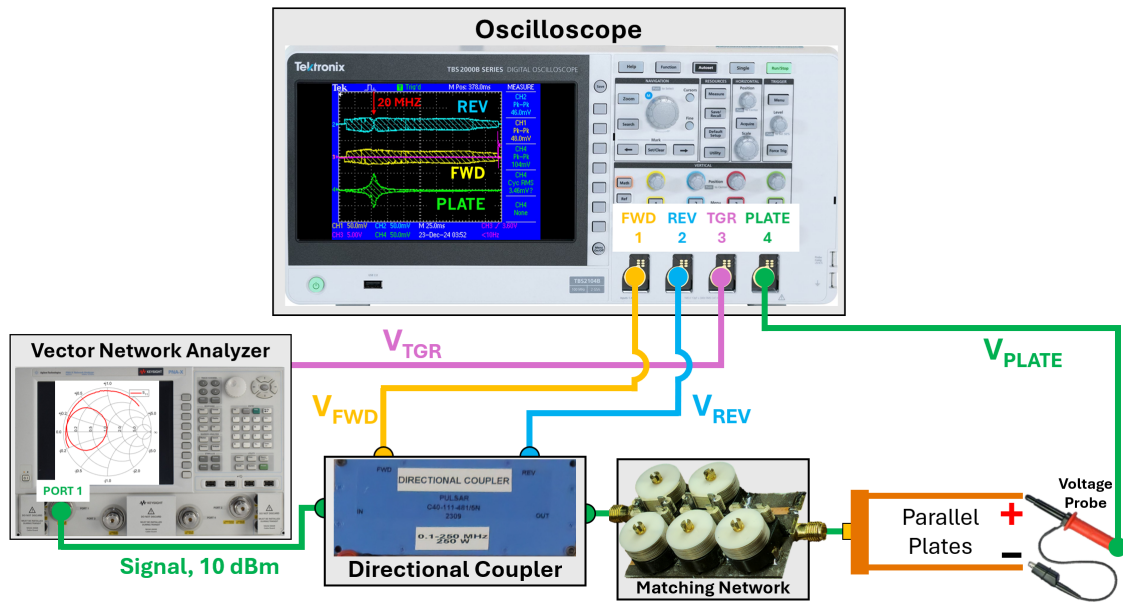


Figure 5.3: Layout of setup used for obtaining an matched measurements. Signal was fed through a directional coupler and the matching network before exciting the parallel plates, allowing measurement of forward, reflected, and plate voltages across frequency.

With all capacitors in the network set to their minimum values and all RF switches turned off to disconnect the inductors, the oscilloscope captured the data shown in Fig. 5.4.

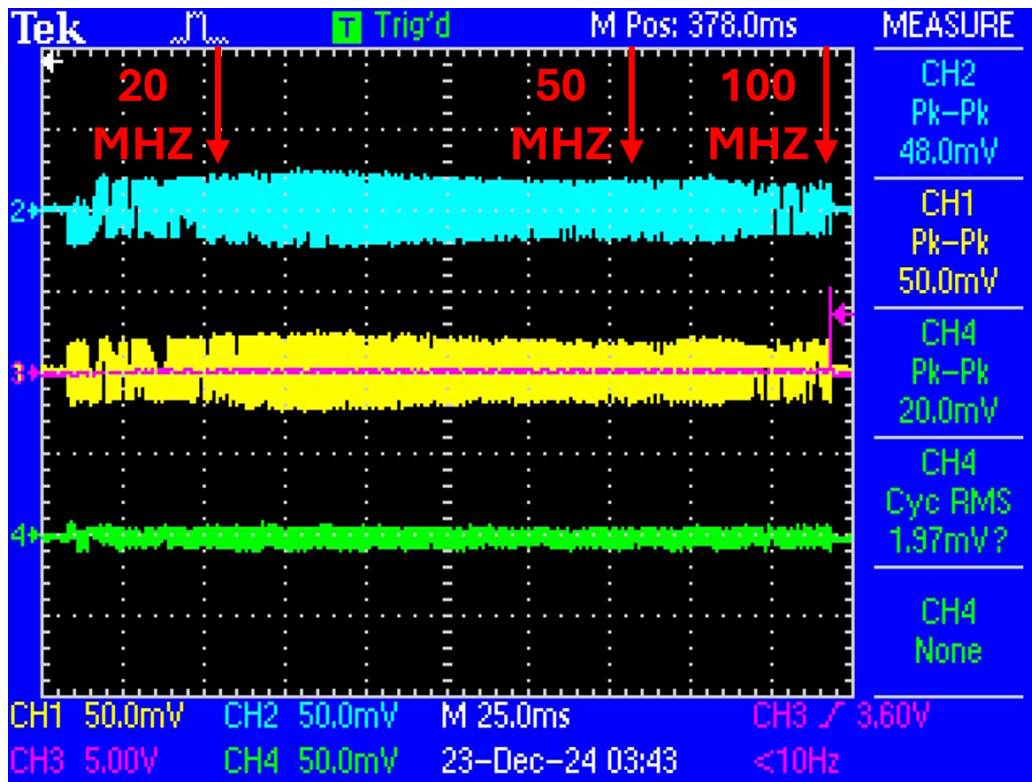


Figure 5.4: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. All capacitors set to minimum and all inductors disconnected. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

As expected without a proper match, there are significant reflections (blue) and minimal voltage being presented across the plates compared to the plates-alone case. Importantly, this setup demonstrated that changes in the network directly impacted the voltage across the plates. In both Fig. 5.2 and Fig. 5.4, the forward voltage (yellow) measured from the oscilloscope appears to gradually decrease with frequency.

The measurements from this section serve to provide a baseline, allowing for later comparison with observations made when tuning the matching network in the following section.

5.1.2 Matching Tests

Focus was then shifted towards demonstrating impedance matches over a range of example frequencies. Again the VNA was configured to output 10 dBm and sweep from 10 MHz to 100 MHz on port 1, fed into the directional coupler, the output of which was then connected to the input of the matching network. A visual of this setup is again provided below.

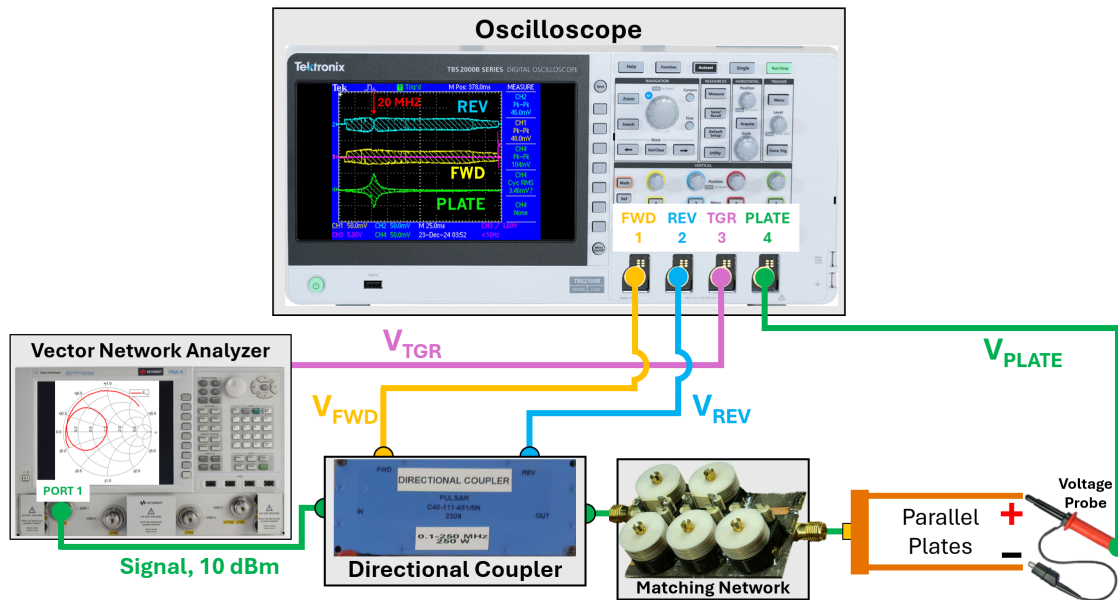


Figure 5.5: Layout of setup used for obtaining an matched measurements. Signal was fed through a directional coupler and the matching network before exciting the parallel plates, allowing measurement of forward, reflected, and plate voltages across frequency.

A match was found at 20 MHz by tuning the network's components while monitoring S_{11} plotted on the VNA in Fig. 5.6 alongside the oscilloscope measurements in Fig. 5.7.

S₁₁ Assembled System, Empty Plates, 20 MHz Match

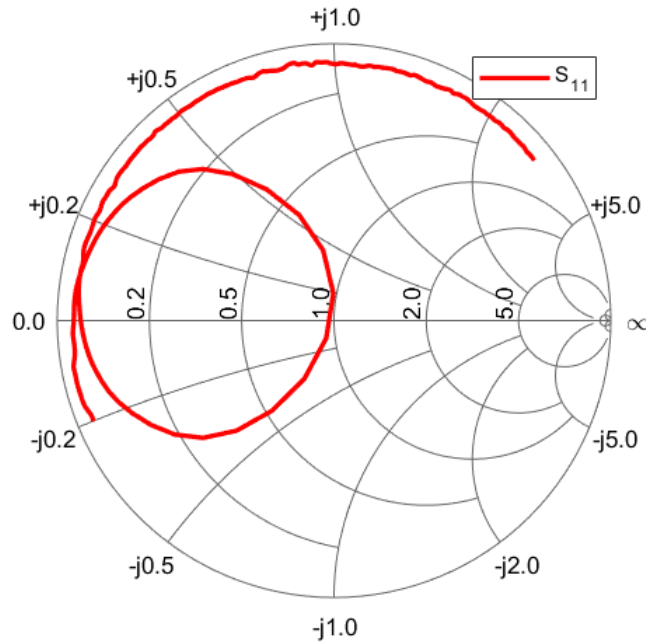


Figure 5.6: S-parameter response for a match at 20 MHz with no material between the plates. Results at middle of Smith Chart support the network’s ability to impedance match well. Frequency swept from 10 MHz to 100 MHz.

In Fig. 5.6 above, the normalized impedance is very close to $Z = 1 + j0$ at 20 MHz, indicating a well-matched system based on the discussion provided in Section 2.3. Given this, it would be expected that a corresponding reduction in reflection and increase in voltage across the plates would be observed at 20 MHz. This is confirmed in Fig. 5.7 below, presenting the oscilloscope measurements for this configuration.

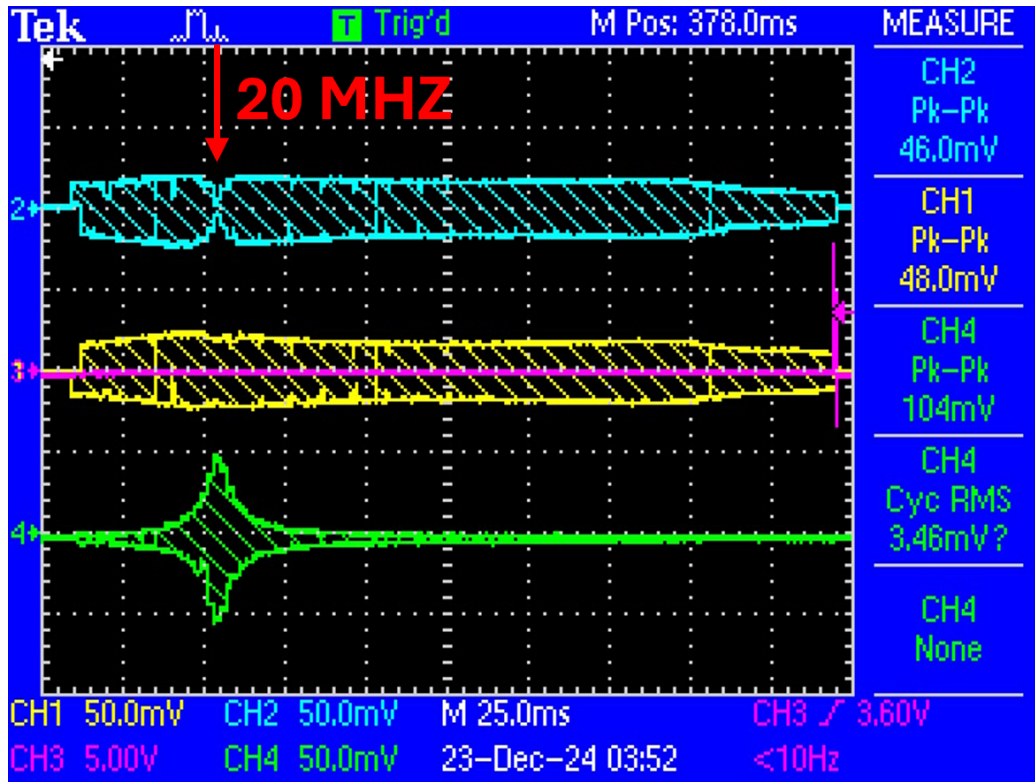


Figure 5.7: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for a match at 20 MHz, no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

It can be seen in Fig. 5.7 that around 20 MHz there is a sharp decrease in the reflection, as measured from the directional coupler, approaching zero. Simultaneously, measured voltage across the plates rises sharply at 20 MHz, reaching a peak value of $104 \text{ mV}_{\text{pkpk}}$ ($36.8 \text{ mV}_{\text{rms}}$). This indicates an increase in the voltage measured across the plates by a factor of approximately $\times 2.61$ compared to Fig. 5.2. As discussed in Section 2.3, the two observations of decreased reflection and increased voltage across the load are key indications that an impedance match has been achieved at this frequency, allowing for more efficient power transfer from the source to the plates as compared to the case where the plates were connected alone.

After creating a match for 20 MHz, a cylindrical sausage cut to approximate the dimen-

sions of the tissue model simulated in Section 3.3.3 was placed between the plates. While this material is not an exact match for the estimated dielectric properties which were used in simulation, this choice removed the need for living tissue, such as an animal subject, which would have introduced unnecessary complexity into what is intended to be a set of initial tests. The addition of this “tissue analog” caused a slight decrease in the voltage in Fig. 5.9 across the plates to 94 mV_{pkpk} (33.2 mV_{rms}) and a slight corresponding shift in S_{11} on the Smith Chart in Fig. 5.8.

S_{11} Assembled System with Tissue Analog, 20 MHz Match

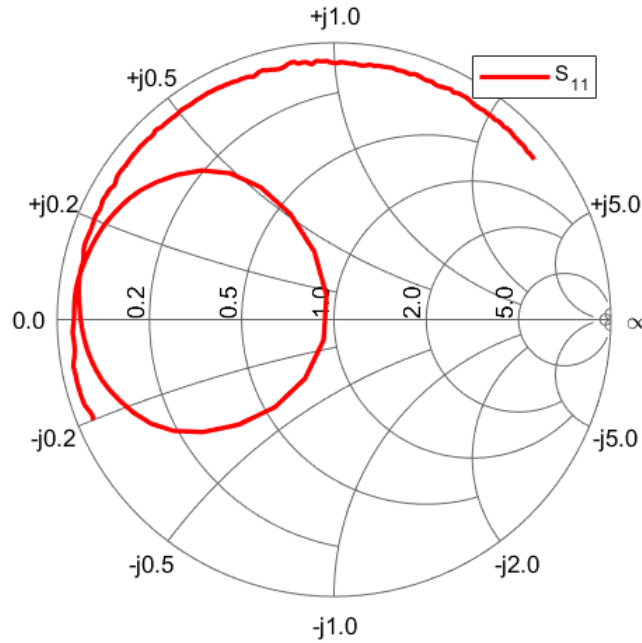


Figure 5.8: S-parameter response for a match at 20 MHz with tissue analog between the plates. Compared to the empty case, the impedance is shifted, highlighting the additional loading introduced by the material. Frequency swept from 10 MHz to 100 MHz.

As compared to the case where the plates were empty in Fig. 5.6, the normalized impedance at 20 MHz has shifted in Fig. 5.8, consistent with the expected changes in impedance due to the dielectric properties of the tissue analog previously observed in simulation in Fig. 3.33 and Fig. 3.39. While this shift results in a deviation from the match

seen above in Fig. 5.6, the impedance remains fairly close to $Z = 1 + j0$ at the center of the Smith Chart, indicating that the system still retains a partial match. Thus, it would be expected to still observe a corresponding decreased reflection and increased voltage across the plates at 20 MHz, but to a lesser extent than seen with the proper match in Fig. 5.7. This is examined in Figure 5.9 below, which provides the oscilloscope measurements for this configuration.

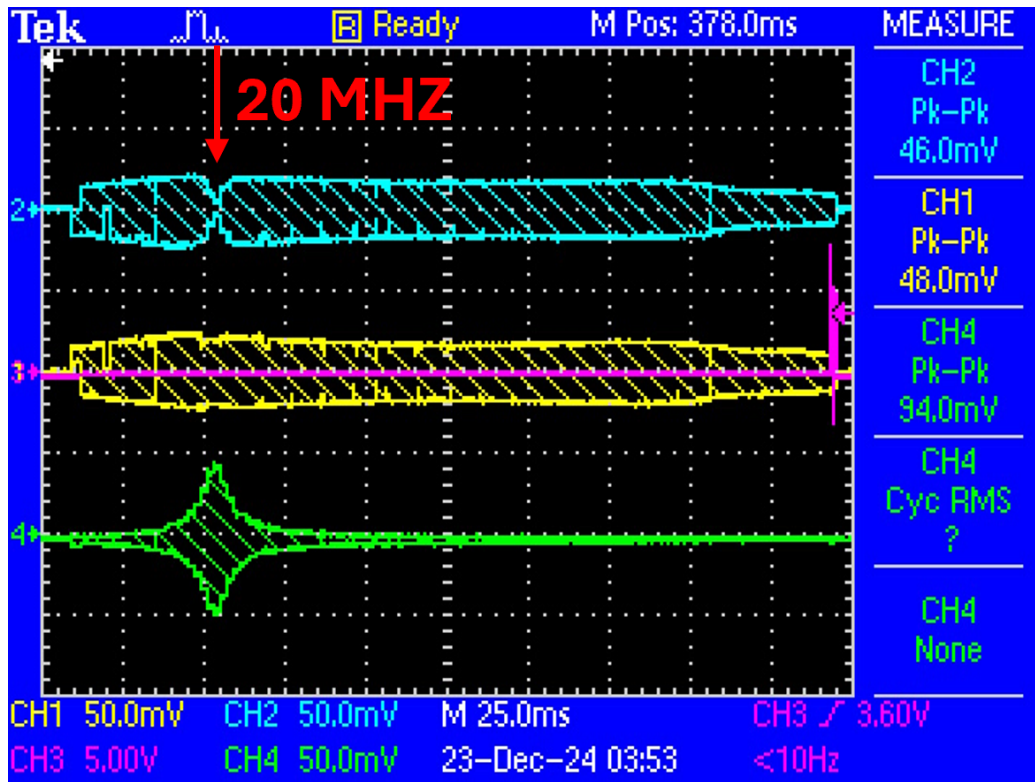


Figure 5.9: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for a match at 20 MHz, tissue analog between plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

The changes seen by placing the tissue analog between the plates in Fig. 5.9 are expected, as the tissue analog alters the impedance seen by the matching network and thus results in a reduction in the quality of the match, aligning well with observations made from simulated data in Section 3.4.2. The partial match that remains with the addition of

the tissue analog still results in an increased voltage across the plates at 20 MHz.

The same process was repeated at 50 MHz, where a match was created through a similar tuning procedure. The results for the match at 50 MHz without the tissue analog are shown below in Fig. 5.10 and Fig. 5.11.

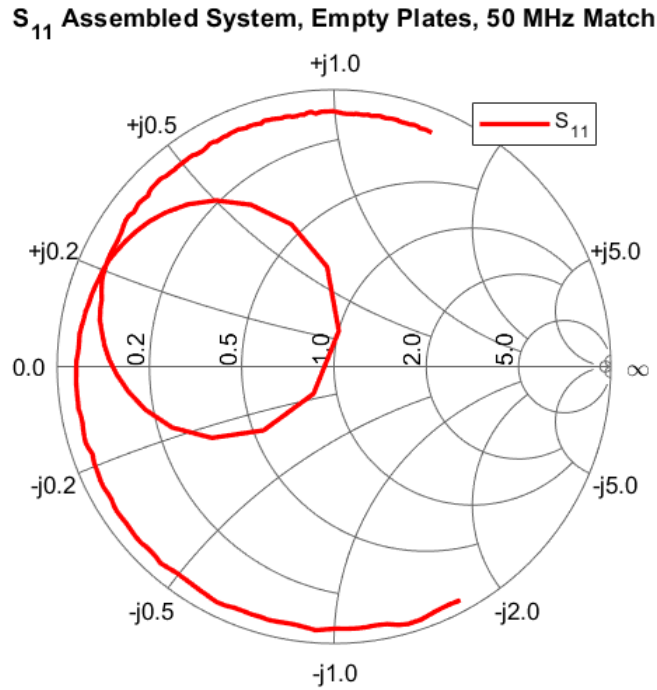


Figure 5.10: S-parameter response for a match at 50 MHz with no material between the plates. Results at middle of Smith Chart support the network's ability to impedance match well. Frequency swept from 10 MHz to 100 MHz.

The S-parameter plot in Fig. 5.10 above again shows a normalized impedance very close to $Z = 1 + j0$ at 50 MHz, again indicating that the system has been well-matched at 50 MHz. Further supporting this, a corresponding decrease in reflection and increase in voltage at this frequency is observable in Fig. 5.13 below.

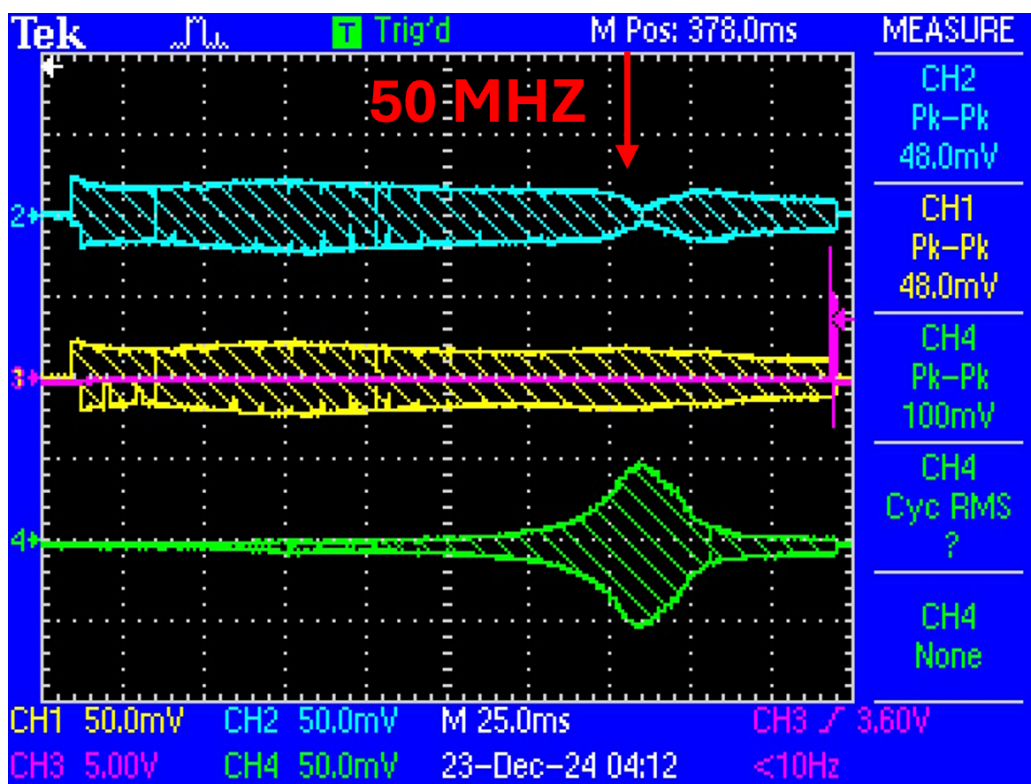


Figure 5.11: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for a match at 50 MHz, no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

At 50 MHz, the behavior was consistent with the results at 20 MHz, with a rise in the maximum voltage across the plates, in this case, to $100 \text{ mV}_{\text{pkpk}}$ ($35.4 \text{ mV}_{\text{rms}}$). Compared to the voltage across the unmatched plates shown in Fig. 5.2, the voltage is increased by a factor of approximately $\times 2.0$ with a match at this frequency.

The introduction of the tissue analog into the 50 MHz matched system also mirrors the results observed at 20 MHz, with an impedance shift, shown below in Fig. 5.12.

S_{11} Assembled System with Tissue Analog, 50 MHz Match

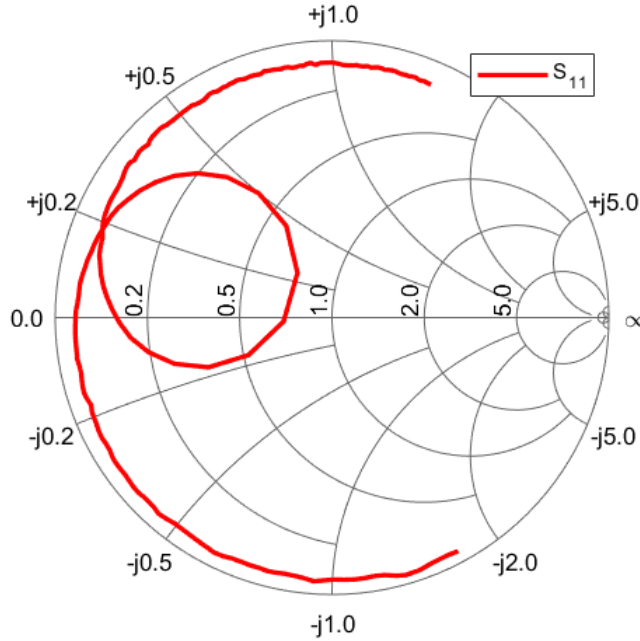


Figure 5.12: S-parameter response for a match at 50 MHz with tissue analog between the plates. Compared to the empty case, the impedance is shifted, highlighting the additional loading introduced by the material. Frequency swept from 10 MHz to 100 MHz.

The observations provided between the Smith Charts in Fig. 5.6 and Fig. 5.8 are very consistent with the observations that could be made to compare the Smith Charts in Fig. 5.10 and Fig. 5.12- as compared to the case where the plates were empty in Fig. 5.10.

The observations provided between the Smith Charts in Fig. 5.6 and Fig. 5.8 (with and without the tissue analog, respectively, at 20 MHz) are consistent with the observations between Fig. 5.10 and Fig. 5.12 (with and without the tissue analog, respectively, at 50 MHz). The normalized impedance has shifted as expected, due to the dielectric properties of the tissue analog influencing the quality of the match. Alongside this change in impedance, a reduction in reflection and increase in voltage across the plates can still be observed at 50 MHz, examined in Fig. 5.13 below.

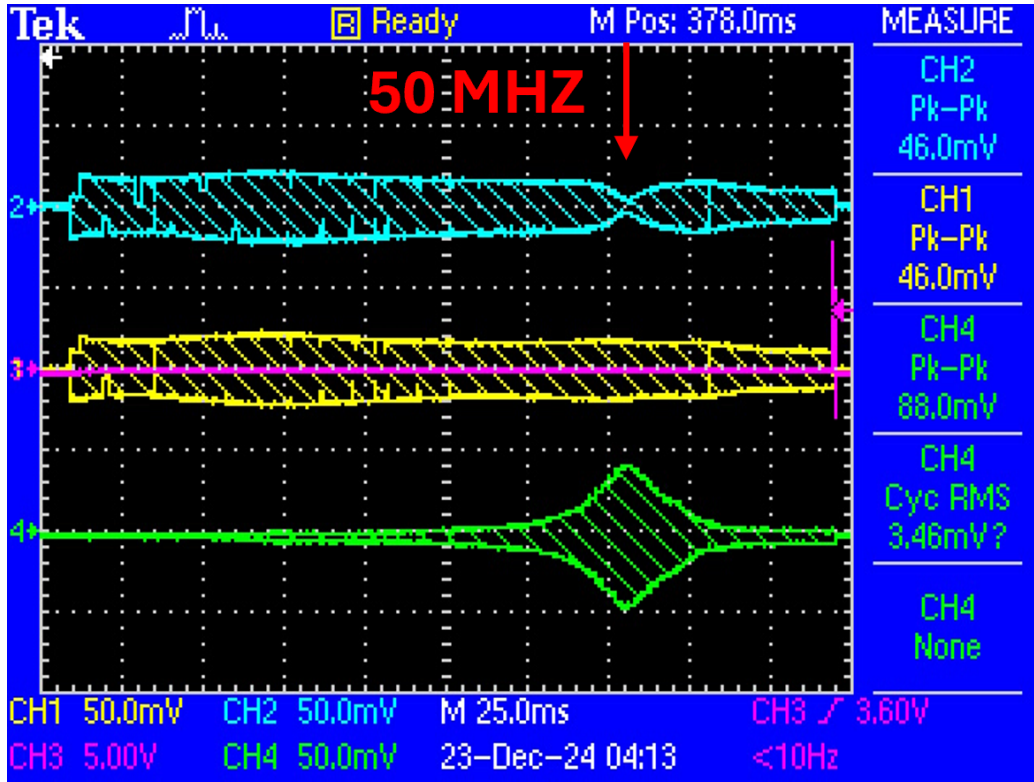


Figure 5.13: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for a match at 50 MHz, tissue analog between plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

As expected, the addition of the tissue analog into the matched system at 50 MHz in Fig. 5.13 above shows a partial match, but with the voltage across the plates instead reaching a decreased maximum of $88 \text{ mV}_{\text{pkpk}}$ ($31.1 \text{ mV}_{\text{rms}}$). This follows the trends observed at 20 MHz.

A similar tuning procedure was again performed for a match at 100 MHz. The Smith Chart plot of S_{11} in Fig. 5.14 shows a normalized impedance very close to the center of the Smith Chart at $Z = 1 + j0 \Omega$, indicating a strong impedance match as discussed in Section 2.3.2.

S₁₁ Assembled System, Empty Plates, 100 MHz Match

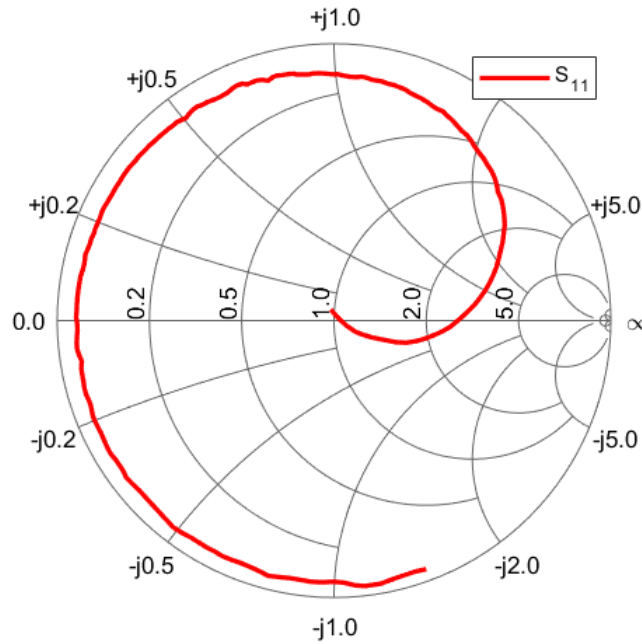


Figure 5.14: S-parameter response for a match at 20 MHz with no material between the plates. Results at middle of Smith Chart support the network's ability to impedance match well. Frequency swept from 10 MHz to 100 MHz.

However, as seen below in Fig. 5.15, the maximum voltage measured across the plates for the match at 100 MHz was 52 mV_{pkpk} (18.4 mV_{rms}), a value 74% of that presented in Fig. 5.2 when the plates alone were connected for measurement. In contrast, the voltage across the plates when matched at 20 MHz was increased by $\times 2.61$ over the unmatched value. While at 50 MHz, the voltage increased by $\times 2.0$ over the unmatched value. This discrepancy in measured results initially raised concerns about the validity of the match. Yet, it can be seen in Fig. 5.15 that the reflected voltage still goes to near zero as it did with the matches at 20 MHz and 50 MHz.

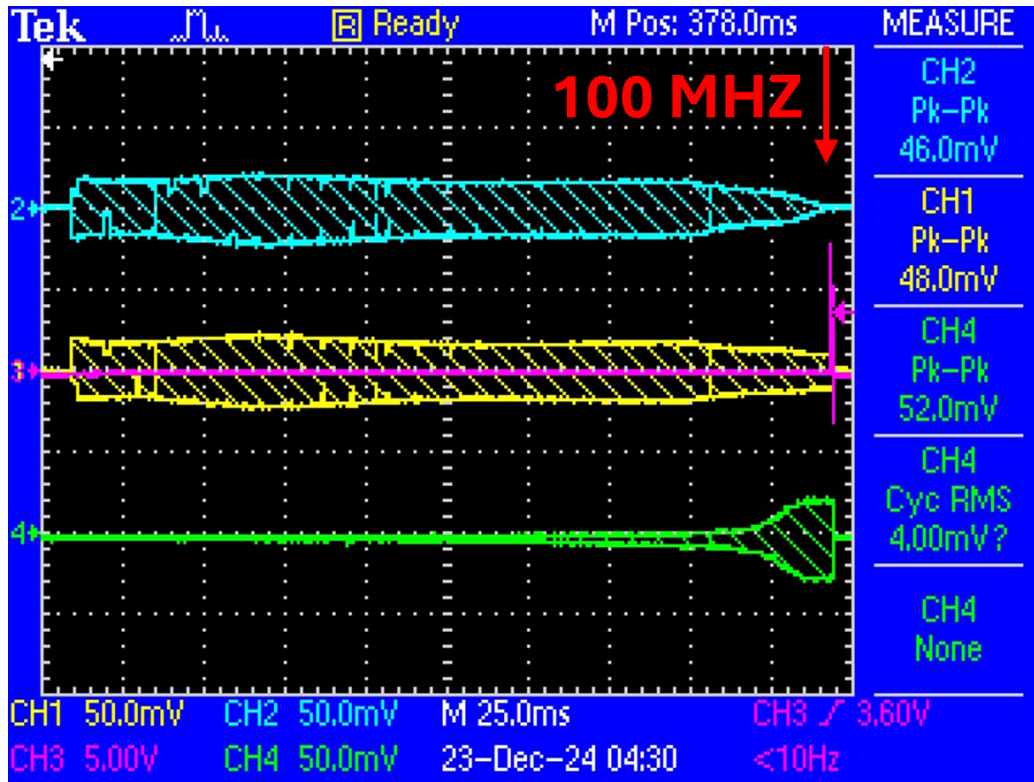


Figure 5.15: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for a match at 100 MHz, no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

Notably, the observed forward voltage at 100 MHz, $\approx 25 \text{ mV}_{\text{pkpk}}$ ($8.84 \text{ mV}_{\text{rms}}$) in Fig. 5.15, is only 52% of the forward voltage measured in Fig. 5.7 and Fig. 5.11 of $\approx 48 \text{ mV}_{\text{pkpk}}$ ($16.97 \text{ mV}_{\text{rms}}$). This reduction in the forward voltage at 100 MHz suggests that less power is reaching the plates in the first place, which in turn corresponds to the reduced voltage across the plates. In fact, the voltage across the plates at 100 MHz is also 52% of the voltage observed at the other frequencies, reinforcing the relationship between forward voltage and the voltage developed across the plates.

When comparing the ratio of the voltage across the plates to the forward voltage, the relationship remains fairly consistent across all frequencies- the voltage across the plates is greater than the forward voltage by a factor of 2.17, 2.09, and 2.08 at 20 MHz, 50 MHz, and

100 MHz, respectively. This trend across frequencies is similar to that experienced when matches were made in simulation with the probes connected with the voltage, increasing by factors of 1.2, 1.0, 0.9 at 20 MHz, 50 MHz, and 100 MHz, respectively. The similarity in these trends further supports the consistency of the impedance matching behavior across frequencies when the probes are connected, despite differences between simulated and experimental results.

During the process of tuning for the previously described matches, an example of such behavior was captured below in Fig. 5.16. The effects seen in Fig. 5.16 are likely the result of factors inherent to the RF system and its components at higher frequencies- power losses in the matching network increasing due to the parasitic effects of the components and probes.

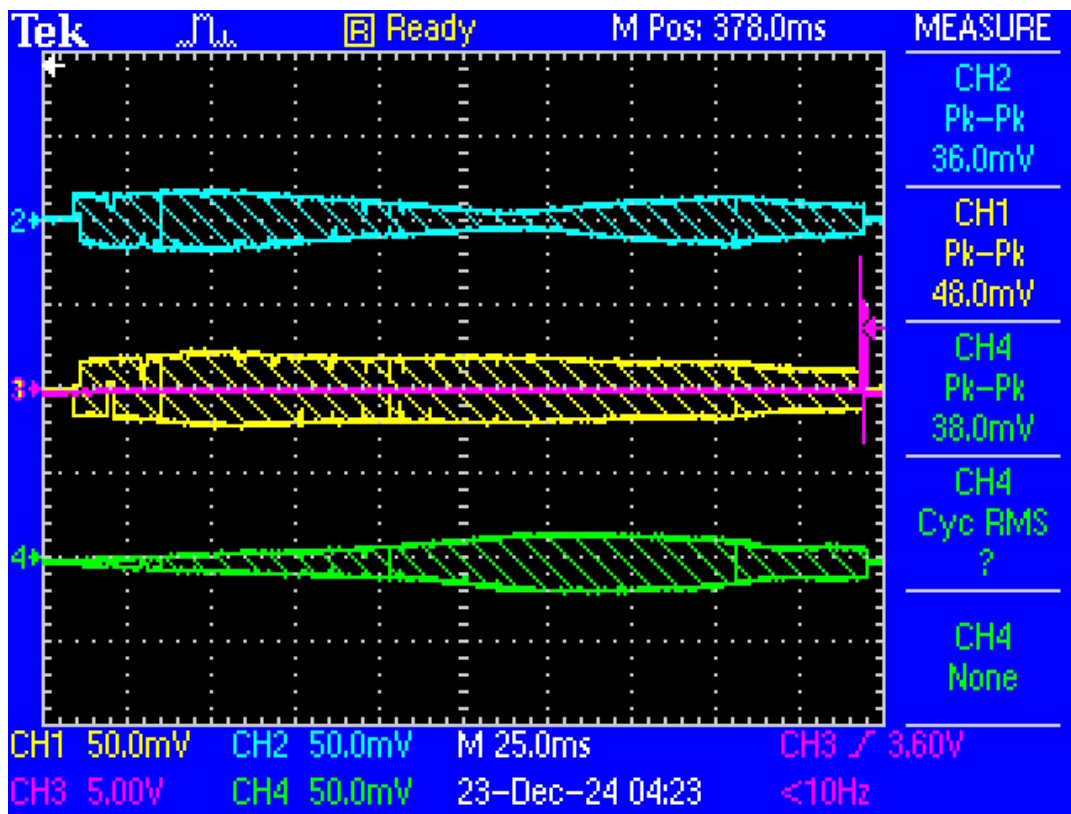


Figure 5.16: Oscilloscope screenshot for a 10 MHz to 100 MHz sweep. Network tuned for lossy effects, no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green), and the VNA trigger signal (pink).

For instance, the reflection voltage over the frequency sweep gradually decreases and eventually reaches a minimum, occurring over a broad frequency range and does not drop to near-zero as observed when matched properly. Similarly, the voltage across the plates shows a gradual rise and fall over the same frequency range, rather than the sharp increase and decrease over a small set of frequencies or so seen in cases of proper matching.

Standing out the most is that the minimum reflection and the maximum voltage across the plates do not align at the same frequency, likely a result of the natural resonances and filtering properties of the matching network, as the voltage observed across the plates could represent a filtered signal that includes losses from the capacitors and other components

in the network. Future work could aim to address these issues by selecting components with higher Q ratings to aid in reducing losses. While this was a known limitation during the design phase, budget constraints restricted the choice of components. Nevertheless, these differences are significant compared to cases where a match was achieved, where the reflection sharply dips over a narrow frequency range and aligns with the peak voltage across the plates. Additionally, Fig. 5.16 provides support to the conclusion that the other test cases represent impedance matches.

The results in this section demonstrate the effectiveness of the network in achieving a match across the tested frequencies to reducing reflections and maximize the voltage across the plates. However, further validation is required to assess the system's performance. The following section involves single-frequency tests, where the network's impact on voltage and power distribution is analyzed more closely to quantify its effectiveness.

5.1.3 Single Frequency Tests

To further investigate the interaction between the plates and the tissue analog, and to gain a more precise understanding of the behavior, particularly the reflection, results were gathered with a set of constant-frequency input signals. Otherwise, the setup used to obtain these experimental results remains the same as depicted in Fig. 5.5.

The network was tuned for a match at 50 MHz *with the tissue analog* placed between the plates before configuring the VNA to output at a constant 50 MHz rather than sweeping over a range of frequencies.

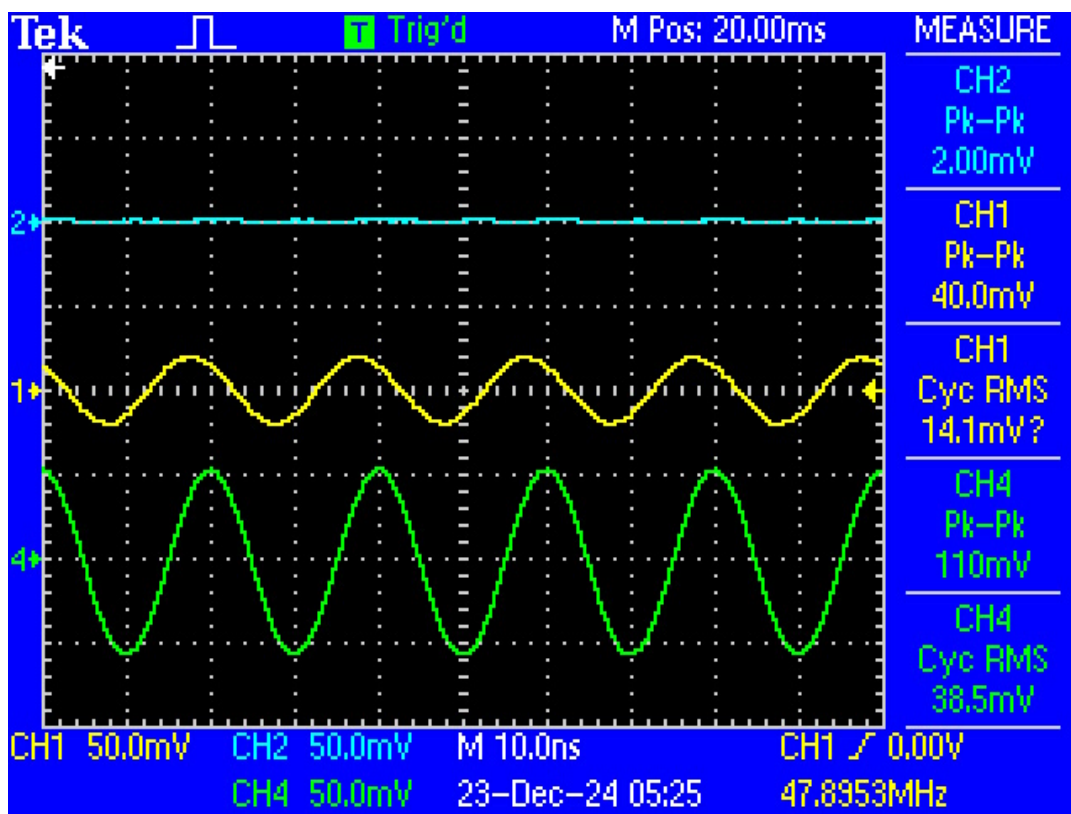


Figure 5.17: Oscilloscope screenshot for 50 MHz signal. Network tuned for a match at 50 MHz, tissue analog between plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green).

With the tissue analog between the plates and the network tuned for a match at 50 MHz, the reflection was observed to be $2 \text{ mV}_{\text{pkpk}}$ ($0.7 \text{ mV}_{\text{rms}}$) while the voltage across the plates was observed to be $110 \text{ mV}_{\text{pkpk}}$ ($38.9 \text{ mV}_{\text{rms}}$). Removing the tissue analog caused an expected shift in the impedance, as seen below in Fig. 5.18, with the reflection rising to $18 \text{ mV}_{\text{pkpk}}$ ($6.36 \text{ mV}_{\text{rms}}$) and the voltage across the plates dropping to $100 \text{ mV}_{\text{pkpk}}$ ($35.4 \text{ mV}_{\text{rms}}$).

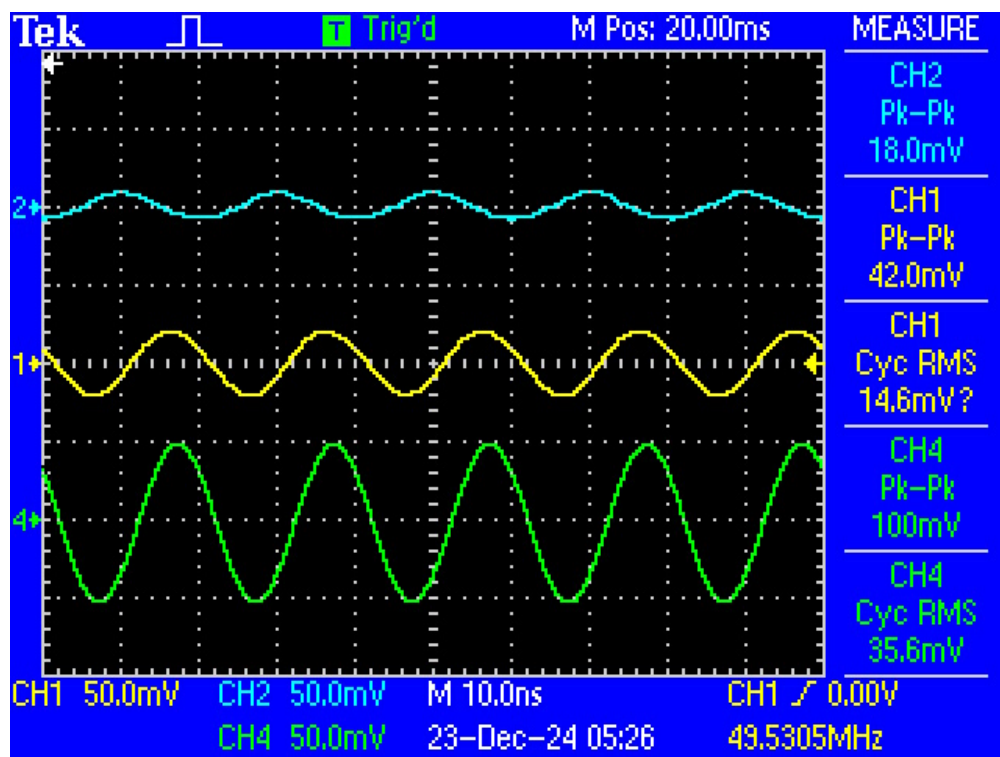


Figure 5.18: Oscilloscope screenshot for 50 MHz signal. Network tuned for a match at 50 MHz, no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green).

This increase in reflected voltage signifies a degradation in the quality of the match, reducing power transfer to the plates. The reflected voltage of $18 \text{ mV}_{\text{pkpk}}$ ($6.36 \text{ mV}_{\text{rms}}$) still remains lower than the forward voltage of $42 \text{ mV}_{\text{pkpk}}$ ($14.8 \text{ mV}_{\text{rms}}$), further supporting claims made in Section 5.1.2 that while the quality of the match has decreased, there still remains some power at the plates. This is backed up by looking to the frequencies where a match was not made in the tests from Section 5.1.2- the unmatched conditions resulting in reflection voltages comparable to the forward voltage, again supporting that the system in Fig. 5.17 above still retains a partial match.

In addition, equivalent tests carried out at 20 MHz and 100 MHz, first finding a match with the tissue analog between the plates before removing it and observing the change in voltages, produced the results shown below in Fig. 5.19 and Fig. 5.20. The observed

changes in voltage before and after removing the tissue analog indicate that it influences the system's impedance, validating that the plates are effectively interacting with the material. This interaction is essential in demonstrating that the system functions as intended, with the matching network facilitating power transfer to the material placed between the plates.

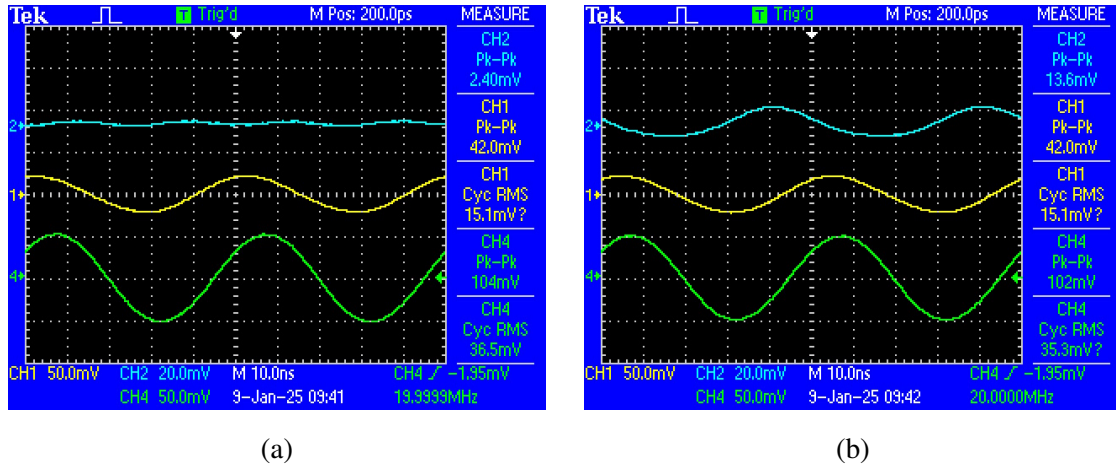


Figure 5.19: Oscilloscope screenshots for 20 MHz signal. Network tuned for a match at 20 MHz, (a) tissue analog between the plates, (b) no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green).

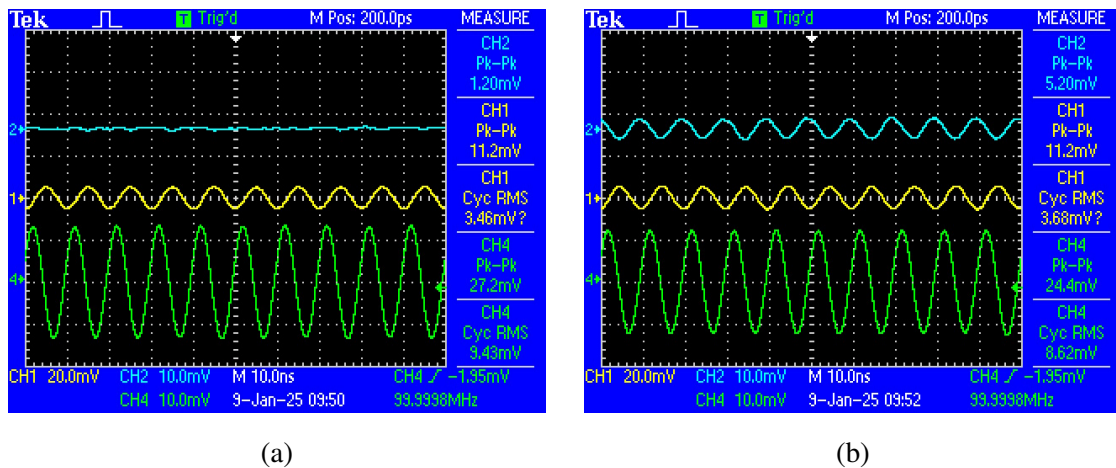


Figure 5.20: Oscilloscope screenshots for 100 MHz signal. Network tuned for a match at 100 MHz, (a) tissue analog between the plates, (b) no material between the plates. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green).

The results of the single-frequency tests further reinforce the previously observed trend in which the voltage across the plates shows an increase over the forward voltage, with the factor of increase gradually decreasing with increases in frequency. As seen in Fig. 5.17, Fig. 5.9, and Fig. 5.20a, the voltage across the plates increases by a factor of 2.75, 2.48, and 2.43 at 20 MHz, 50 MHz, and 100 MHz, respectively.

However, some caveats must be considered when comparing the results from this section to those from Section 5.1.2. One important factor is the sampling rate and measurement resolution of the oscilloscope, which is likely to introduce small variations in the recorded voltages. Additionally, as previously touched on in Section 3.4.2, the exact voltage increase depends on the component values within the matching network. Since these components are analog and tuned manually, slight deviations in their values could lead to small differences in the observed voltage increase, making perfect agreement between the two testing approaches unlikely. Nevertheless, the agreement between these two independent methods of evaluation strengthens the validity of the results, helping to confirm that the matching network effectively increases the voltage across the plates in different measurement conditions.

The single-frequency tests in this section provided a focused evaluation of the system's ability at fixed operating points, complementing the matching tests conducted earlier in Section 5.1.2. These tests validated the matching network's performance at selected frequencies and the observations made form a background for the detailed analysis presented in the following section, where the data is examined to assess the overall system performance and identify opportunities for refinement.

5.1.4 Analysis

To further understand the amount of power delivered to the load through the matching network, the series of equations originally introduced in Section 2.3 can be used to estimate the voltage across the plates and power delivered to the tissue analog, which together act as the load.

The matching network was first tuned for the target frequency before being disconnected from the system and connected to the VNA for collection of two-port S-parameter data, depicted below in Fig. 5.21, where the input of the matching network is connected to port 1 and the output of the matching network is connected to port 2.

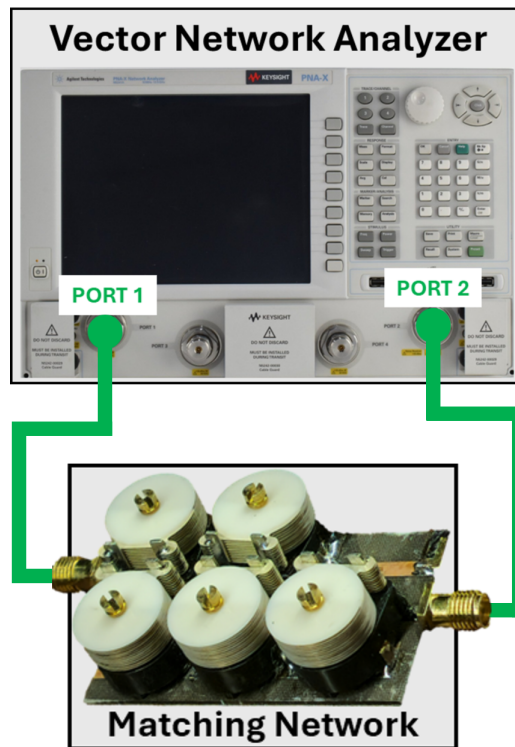


Figure 5.21: Layout of setup used for obtaining two-port S-parameter measurements from the matched network.

Similarly, two ports were connected to the HFSS circuit shown in Fig. 3.42. The

S-parameter measurements from simulation were first used to validate the accuracy and reasonableness of the calculations introduced in Section 2.3. The calculated results from the data derived in simulation should align well with the values presented in Table 3.7 which were extracted directly from HFSS. This supports the validity of applying this same approach to the physically measured data.

To expand on the equations introduced in Section 2.3 and provide a more concrete demonstration, the equations were applied to the simulated data gathered from a match at 50 MHz, a frequency used throughout this work as an example. At 50 MHz, the S-parameters of the network, $S_{network}$, along the plates and tissue sample acting together as the load, S_{load} , were measured as

$$S_{network} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} -0.53 - j0.77 & 0.12 + j0.08 \\ 0.12 + j0.08 & 0.97 + j0.14 \end{bmatrix} \quad S_{11,load} = 0.99 - j0.15 \quad (\text{Eq. 5.1})$$

As used previously in simulation, the voltage at the 50 Ω port with a 100 W of power can be calculated as 100 V_{pk} from [13] as

$$P_{source,watts} = \frac{10^{\frac{P_{in,dB}}{10}}}{1000} = 100 \text{ W} \quad (\text{Eq. 5.2})$$

$$V_{pk,source} = \sqrt{2} \times \sqrt{P_{source} Z_0} = 100 \text{ V}_{pk} \quad (\text{Eq. 5.3})$$

Eq. 2.20 and Eq. 2.21 can then be used with $S_{11,load}$ from the plates, acting as the load, to find voltage and power when unmatched.

$$\begin{aligned} V_{load,unmatched} &= (1 + |S_{11,load}|) \times V_{source} \\ &= (1 + |0.99 - j0.15|) \times 100 \\ &= 199.75 \text{ V}_{pk} \end{aligned} \quad (\text{Eq. 5.4})$$

$$\begin{aligned}
P_{load,unmatched} &= (1 - |S_{11,load}|^2) \times P_{source} \\
&= (1 - |0.99 - j0.15|^2) \times 100 \\
&= 0.51 \text{ W}
\end{aligned} \tag{Eq. 5.5}$$

These results for when the matching network is not in use provide a baseline for comparison, with the aim to increase the voltage across the plates when the tuned matching network is in use. In order to find the voltage across the plates when the matching network is in use, the following equations from [13] were used.

$$V_1^- = S_{11,network} V_1^+ + S_{12,network} V_2^+ \tag{Eq. 5.6}$$

$$V_{2-} = S_{21,network} V_{1+} + S_{22,network} V_{2+} \tag{Eq. 5.7}$$

These equations have been used to derive Eq. 2.24, with the data at 50 MHz plugged-in yielding

$$\begin{aligned}
V_{pk,plates} &= V_2 \\
&= \left[\frac{S_{21,network} (1 + S_{11,load})}{1 - S_{11,load} S_{22,network}} \right] V_1^+ \\
&= \left[\frac{0.12 + j0.08 (1 + 0.99 - j0.15)}{1 - (0.99 - j0.15)(0.97 + j0.14)} \right] 100 \\
&= 1258.12 V_{pk}
\end{aligned} \tag{Eq. 5.8}$$

The total amount of power entering the system through the network can then be determined from Eq. 2.25 as

$$\begin{aligned}
P_{network} &= (1 - |S_{11,system}|^2) \times P_{source} \\
&= (1 - |0.004 + j0.01|^2) \times 100 \\
&= 99.98 \text{ W}
\end{aligned} \tag{Eq. 5.9}$$

However, the portion of $P_{network}$ leaving the network and going to the load is found from Eq. 2.26 as

$$\begin{aligned}
P_{load} &= P_{network} \times |S_{21,network}|^2 \times \frac{(1 - |S_{11,plates}|^2)}{|1 - S_{22,network}S_{11,plates}|^2} \\
&= 99.9 \times |0.12 + j0.08|^2 \times \frac{(1 - |0.99 - j0.15|^2)}{|1 - (0.97 + j0.14)(0.99 - j0.15)|^2} \quad (\text{Eq. 5.10}) \\
&= 20.25 \text{ W}
\end{aligned}$$

Applying this process to the full data set extracted from the simulated plate and circuit designs, again with a theoretical power of 100 W and voltage of 100 V_{pk} from the source, yielded the results presented below in Table 5.1. For reference, also included are the corresponding values extracted directly from the HFSS simulation performed in Fig. 3.43 and previously presented in Table 3.7.

Frequency	Calculated from Simulated Data			Extracted from HFSS (Table 3.7)		
	20 MHz	50 MHz	100 MHz	20 MHz	50 MHz	100 MHz
$V_{pk,plates,unmatched}$	199.9 V	199.7 V	199.3 V	233.9 V	236.3 V	246.4 V
$P_{unmatched}$	0.1 W	0.5 W	1.3 W	0.1 W	0.5 W	1.2 W
$V_{pk,plates,matched}$	1,968.3 V	1,258.1 V	809.9 V	2,529.7 V	1,510.6 V	1,034.3 V
$P_{network}$	99.9 W	99.9 W	99.9 W	-	-	-
P_{load}	12.2 W	20.3 W	22.0 W	-	-	-
$P_{loss,tissue}$	-	-	-	14.4 W	21.5 W	23.1 W
$\Delta V_{pk,plates}$	×10.3	×6.3	×4.1	×10.8	×6.4	×4.1
$(\Delta V_{pk,plates})^2$	×105.8	×39.7	×16.5	×116.7	×41.0	×17.6
ΔP_{load}	×105.8	×39.9	×16.9	×123.1	×42.8	×18.5

Table 5.1: Comparison of voltage and power parameters from calculated results and those pulled from HFSS (Table 3.7) across three target frequencies of 20 MHz, 50 MHz, and 100 MHz. Input set to 100 W, 100 V_{pk} from 50 Ω source.

Notably, due to the setup in HFSS and calculations performed, no values for $P_{network}$ or P_{load} were sourced as a direct measurement from HFSS. Likewise, there is no calculated value for the loss in the tissue from S-parameter data; however, the power lost in the tissue sample as measured in HFSS, $P_{loss,tissue}$, provides a useful corresponding value in comparison to P_{load} . While there are small differences between the values for P_{load} and $P_{loss,tissue}$, these can likely be attributed to variations between how calculations are made in HFSS compared to the calculations made with S-parameter data. For example, the voltages (and thus fields as well) calculated with S-parameter data are lower than those pulled directly from HFSS, carrying through to a slightly lower calculated power.

P_{load} accounts for the power that successfully exits the network and reaches the plates after accounting for transmission losses and reflections within the matching network. However, it is not simply the total power arriving at the plates, but rather the portion that is actively being consumed by the plates. This distinction is important, as power not absorbed by the tissue is either reflected back toward the network or oscillates within the plate structure without contributing to energy transfer into the medium- a factor accounted for in the third term of Eq. 5.10 above.

In parallel, the values for $P_{loss,tissue}$ pulled directly from simulation represent the amount of power that is absorbed within the tissue sample. The calculated values for P_{load} show reasonable agreement with values for $P_{loss,tissue}$. This agreement further suggests that P_{load} calculated using Eq. 5.10 primarily represents real power transfer rather than reactive energy stored within the plates. If a significant portion of P_{load} was stored as reactive power rather than being dissipated within the tissue sample, a much larger discrepancy between the two values would be expected. Instead, the close alignment indicates that a significant majority of power calculated by Eq. 5.10 is being transferred to the tissue. This further reinforces the validity of using the previously outlined series of calculations to evaluate power transfer in experimental conditions.

Calculated from Experimental Data			
Frequency	20 MHz	50 MHz	100 MHz
$V_{pk,plates,unmatched}$	199.9 V	199.8 V	199.3 V
$P_{unmatched}$	0.07 W	0.36 W	1.41 W
$V_{pk,plates,matched}$	511.8 V	259.4 V	224.5 V
$P_{network}$	84.9 W	96.9 W	99.9 W
P_{load}	0.4 W	0.6 W	1.8 W
$\Delta V_{pk,plates}$	$\times 2.6$	$\times 1.3$	$\times 1.1$
$(\Delta V_{pk,plates})^2$	$\times 6.6$	$\times 1.7$	$\times 1.3$
ΔP_{load}	$\times 5.6$	$\times 1.6$	$\times 1.3$

Table 5.2: Comparison of voltage and power parameters from calculated results using experimentally-gathered S-parameter data across three target frequencies of 20 MHz, 50 MHz, and 100 MHz. Theoretical input set to 100 W, 100 V_{pk} from 50 Ω source.

When applying the calculations to the experimentally measured S-parameter data above in Table 5.2, the key trends observed in simulation persist. The relationship between the voltage across the plates and the power delivered remains consistent with expectations—the voltage increase achieved when using the matching network results in a corresponding increase in power delivered. Furthermore, the square of the increase in voltage continues to serve as a reasonable predictor for the power increase, aligning with trends established in simulation.

However, while the general behavior of the system is preserved, the absolute values for P_{load} are lower than those observed in simulation, indicating that less power is ultimately reaching the plates. When compared to previous results, such as those from Table 3.7, which presented simulated plate voltages with increases of factors of 12.1, 5.2, and 3.6 at 20 MHz, 50 MHz, and 100 MHz, respectively, the experimentally observed voltage increases of 2.6, 1.3, and 1.1 indicate a reduction in the effectiveness of the physical network in increasing power delivery.

This reduction in the factor of the voltage increase, and thus the power delivered, can be primarily attributed to constraints in the physical implementation of the matching network. In the simulated circuit, component values were allowed to vary freely to achieve an optimal match, whereas in the experimental setup, practical limitations previously touched on in Section 3.4.3 on the tunable inductor and capacitor values restricted the achievable matching conditions. As discussed in Section 3.4.3, the inductor tuning range in the physical network was limited, preventing the use of higher inductance values that would have been necessary to provide stronger impedance transformation at certain frequencies. Conversely, while the tunable capacitors had sufficient range, their inability to reach sufficiently low values introduced further limitations in optimizing the network response.

Despite these constraints, the experimental matching network was able to successfully impedance match over a range of frequencies, as evidenced by the calculated results in Table 5.2 from measured S-parameter data. This reinforces that the network and plate design remain effective in improving power transfer even under component limitations, and furthermore suggests significant performance improvements could be achieved through expanded component tunability. Overall, these results represent a clear success in demonstrating the matching network's ability to increase the observed voltage across the plates while secondarily increasing overall power transfer and efficiency in comparison to an unmatched case. Testing at higher power levels, explored in the following section, will provide further insights into the system's behavior at more demanding, practical power levels.

5.2 High Power Testing

After obtaining promising results from the tests in Section 5.1, the next step involved testing at higher power using the complete system shown in Fig. 4.1. First, the network was tuned for 20 MHz at low power using the VNA and oscilloscope as observation methods to

confirm a match was achieved as described previously. For all following descriptions of the system, the tissue analog was in use and the fans connected to the amplifier's heatsink were turned on. A Klein Tools IR1 Infrared Thermometer was used to measure various parts of the system (measured in place, sans the tissue analog which was measured from the point making contact with the bottom plate).

The provided signal generator was set to output a 20 MHz signal at 10 dBm and was connected directly to the oscilloscope where a voltage of $4.2 V_{pkpk}$ ($1.48 V_{rms}$) was measured as pictured below in Fig. 5.22. When the VNA was previously set to output the same 20 MHz signal at 10 dBm in Fig. 5.19a, the observed voltage of 42 mV_{pkpk} (14.8 mV_{rms}), when adjusted for the directional coupler's 40 dB coupling factor, is $4.2 V_{pkpk}$ ($1.48 V_{rms}$). The consistency of these measured results validate the acceptable performance of the signal generator moving forward.

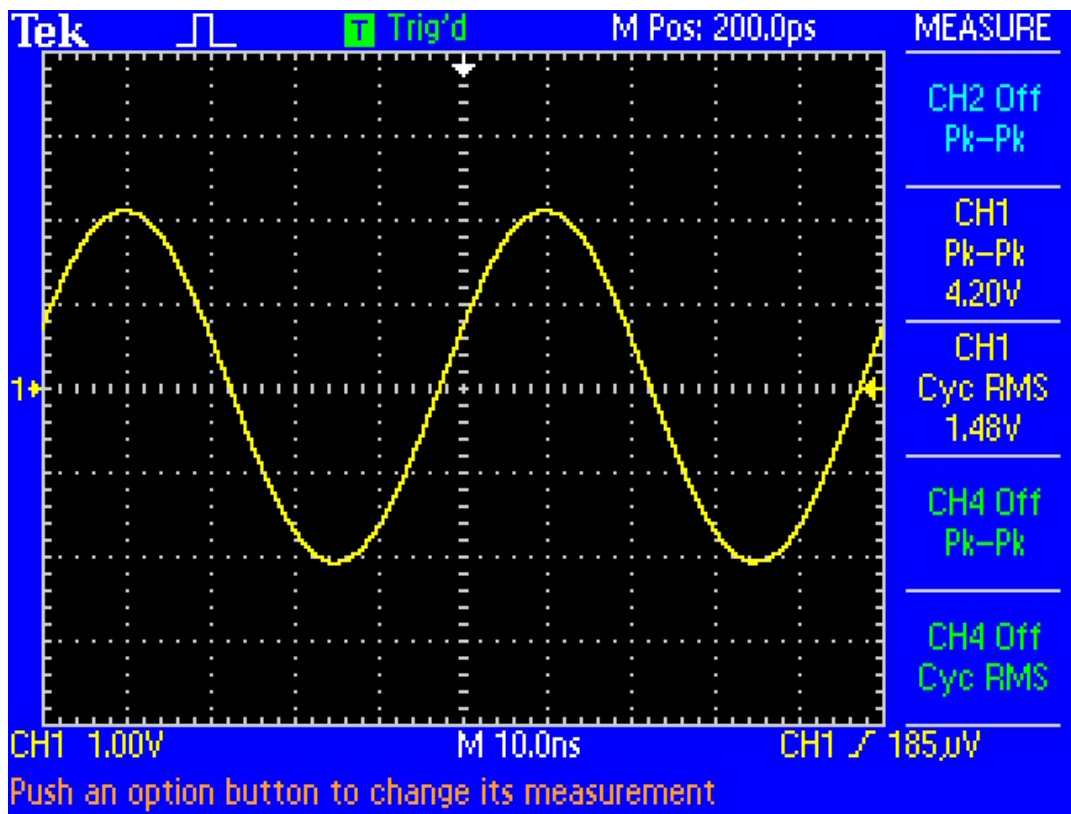


Figure 5.22: Oscilloscope screenshot for 20 MHz signal at 10 dBm, input directly from signal generator.

With the system assembled, the signal generator was set to output a -41 dBm signal at 20 MHz. According to the amplifier's datasheet claim of 51 dB, an input of -41 dBm should result in an output of approximately 10 dBm with the same 4.2 V_{pkpk} from Fig. 5.22 [45]. However, as shown below in Fig. 5.23, the measured forward voltage from the directional coupler was 112 mV_{pkpk} (39.6 mV_{rms}), or 11.2 V_{pkpk} (3.96 V_{rms}) after accounting for the coupling factor.

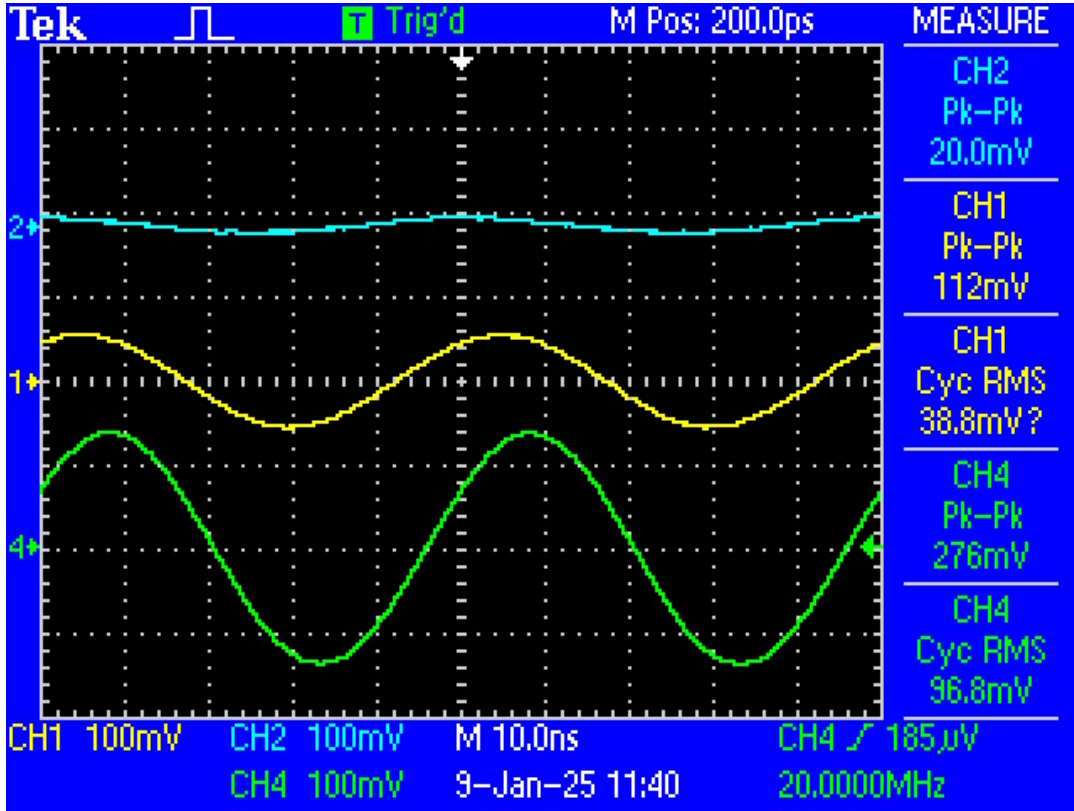


Figure 5.23: Oscilloscope screenshots for 20 MHz signal. Network tuned for a match at 20 MHz, tissue analog between plates, -41 dBm input into amplifier from signal generator. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler, the probe voltage across the plates (green).

From [13], it's given that

$$dB = 20 \log_{10} \left(\frac{V_{out}}{V_{in}} \right) \quad (\text{Eq. 5.11})$$

where the gain is found from the ratio of the output voltage, V_{out} , over the input voltage, V_{in} . With an expected gain of 51 dB and expected output voltage of 4.2 V_{pkpk}, the input voltage from the source at -41 dBm can be found as

$$\begin{aligned}
dB &= 20 \log_{10} \left(\frac{V_{out}}{V_{in}} \right) \\
51 &= 20 \log_{10} \left(\frac{4.2}{V_{in}} \right) \\
V_{in} &= 0.01184 \text{ V}
\end{aligned} \tag{Eq. 5.12}$$

This value for V_{in} , along with the measured voltage of 11.2 V_{pkpk} , can then both be plugged back into Eq. 5.11 to determine the actual amplifier gain as

$$\begin{aligned}
dB &= 20 \log_{10} \left(\frac{V_{out}}{V_{in}} \right) \\
&= 20 \log_{10} \left(\frac{11.2}{0.01184} \right) \\
&= 59.5 \text{ dB}
\end{aligned} \tag{Eq. 5.13}$$

suggesting that the amplifier's gain is closer to 59.5 dB rather than the 51 dB stated in the datasheet. With the -41 dBm input power experiencing a gain of 59.5 dB, the power output from the amplifier would actually be around 18.5 dBm (0.071 W). After documenting these results, the system was allowed to run for 10 minutes. As expected at this power level, no significant heating was measured in any part of the system after the time period had elapsed.

The signal generator output was increased to -11 dBm, which would amplify to 40 dBm (10 W) based on a gain of 51 dBm. On the other hand, if the observed gain seen above in Fig. 5.23 of 59.5 dB holds true, the output would amplify to 48.5 dB (70.8 W).

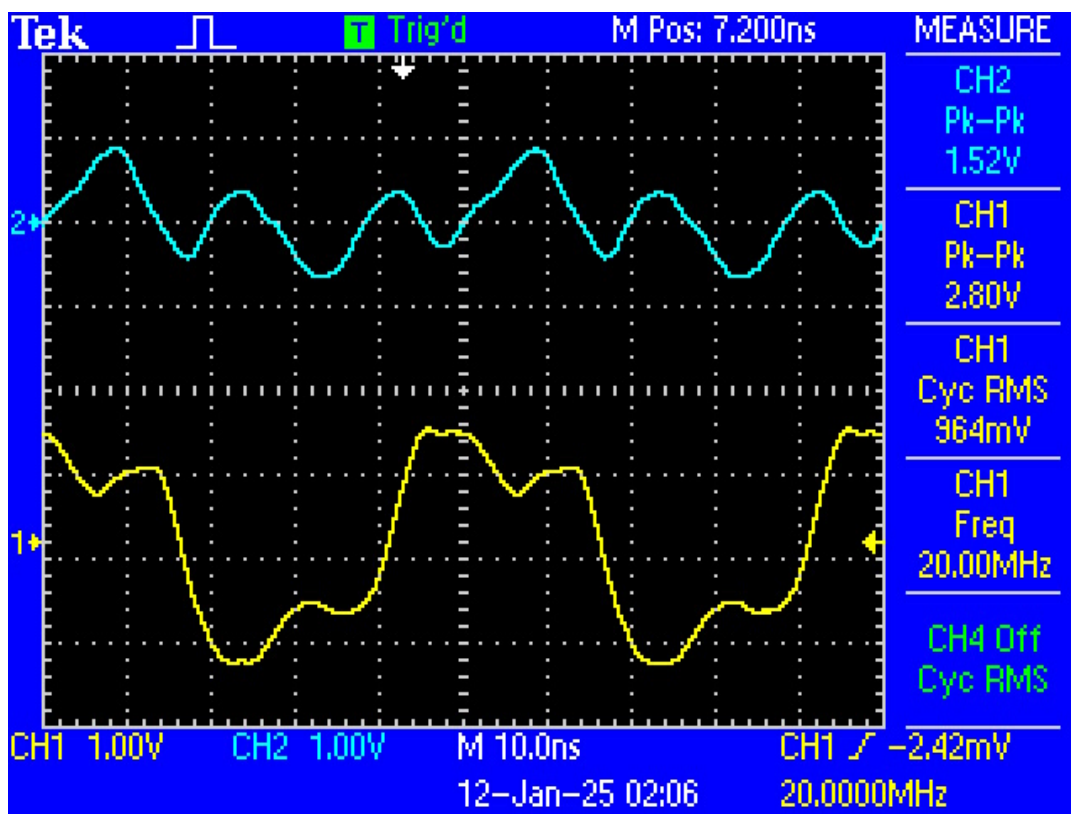


Figure 5.24: Oscilloscope screenshots for 20 MHz signal. Network tuned for a match at 20 MHz, tissue analog between plates, -11 dBm input into amplifier from signal generator. Screenshot shows the forward (yellow) and reverse (blue) lines from the directional coupler.

The resulting signal from Fig. 5.24 was measured as 2.8 V_{pkpk} (990.1 mV_{rms}), or 280 V_{pkpk} (99.01 V_{rms}) after accounting for the coupling factor. By again using Eq. 5.11, the gain can be determined as 57.5 dB, fairly consistent with the previous observation. This would bring the signal up to around 46.5 dBm (44.6 W).

After documenting these results, the system was allowed to run for 10 minutes. After this time had elapsed, some heating was observed in the system. The 28 VDC power supply feeding the amplifier was measured at 88°F/31°C, the power amplifier at 81°F/27°C, the first series capacitor at 91°F/33°C, and the fourth and only inductor in use at 97°F/36°C. As such, all components are well within temperature ratings. However, the observed signal quality in Fig. 5.24 is significantly degraded. Without any changes made to the system,

the reverse voltage, corrected for the coupling factor, had risen to $152 V_{\text{pkpk}}$ ($53.75 V_{\text{rms}}$), indicating an issue with the system. After consulting with Dr. Fulton on the source of the issue, the amplifier was determined as a likely culprit. As a result, it was decided the best course of action would be to suspend further use of the power amplifier. The decision to halt additional high-power experiments was justified by safety considerations and the desire to preserve equipment.

Prior to increasing the input signal to the amplifier in Section 5.2, the matching network had already been tuned. As shown in Fig. 5.23, the ratio of the reflected voltage over the incident voltage is ≈ 0.18 , which when used as the reflection coefficient in Eq. 2.15, finds that $\text{VSWR} = 1.43$. This value is well below the continuous limit at 50 W of $\text{VSWR} = 3$ found in [45]. Additionally, the input signal fed into the amplifier from the signal generator was never raised to or even near the 10 dBm input limit found in [45]. Given all this information, it was not initially deemed necessary to include a circulator between the amplifier and directional coupler. Nevertheless, the use of a circulator in future experiments would offer an added layer of protection in the system that was left out in this work.

Although the high power testing was limited to 20 MHz, the rest of this chapter describes the expected behavior at other possible frequencies. These claims are supported by the agreement between the calculated data presented in Table 5.2 and the observed trends of the matching network across various scenarios presented throughout Section 5.1.2.

5.3 Summary

The results outlined in this chapter provide a comprehensive analysis of the RF system's performance, spanning baseline measurements, initial low-power tests, and high-power testing. Section 5.1.1 established a control by characterizing the behavior of the plates as well as the connected network and plates together without proper impedance matching,

offering a benchmark against which subsequent improvements could be measured.

The matching tests in Section 5.1.2 demonstrated the efficacy of the designed network in achieving impedance matches across the target frequency range, with oscilloscope and S-parameter data confirming the ability to identify matches through minimizing reflections and maximizing the voltage across the plates. These tests also highlighted that while parasitics and other effects can materialize under certain conditions, their impact on the network's performance is largely minimized when a proper impedance match is achieved, as opposed to the significant effects observed for a poor match, demonstrated in Fig. 5.16.

Rather than observing the response of the network with a frequency sweep, Section 5.1.3 measured the system's behavior at single input frequencies and allowed for a better understanding of the interaction between the tissue analog and system response. Furthermore, the observed trends seen in different experimental scenarios reinforce the trends observed in the calculated results presented in Section 5.1.4.

Section 5.1.4 first outlines the approach used for estimating power transfer using S-parameter measurements taken from both the network and parallel plates. The methodology, based on established RF power equations, was first applied to simulated S-parameter data, where it showed strong agreement with directly extracted HFSS results, validating its accuracy. Applying the same approach to experimentally gathered S-parameter data enabled an evaluation of the power reaching the load in physical experimentation, reinforcing the observed voltage-power relationship and highlighting the effect of practical component limitations on power transfer efficiency. While physical constraints on inductor and capacitor tunability resulted in less energy transfer than seen in simulation, the calculations confirmed that the network was able to form an impedance match and provide improvements in voltage and power over the unmatched case. These findings demonstrate the effectiveness of the network and plate design in experimental implementation while also highlighting how further improvements could enhance energy transfer.

Chapter 6

Conclusion

6.1 Conclusion

This thesis has presented the design and analysis of a tunable impedance matching network to facilitate VHF power delivery to a target medium, with parallel efforts dedicated to modeling and simulating the behavior of CNTs under RF excitation. This work built on previous efforts, including work by Dr. Roger Harrison and collaborators, addressing the limitations in system tunability and field concentration seen in earlier designs such as the Comdel system. These challenges associated with, in part, delivering RF energy and the lack of commercially available solutions necessitated the development of an effective, tunable impedance matching network.

The initial motivation for this research stemmed from the promise shown by CNTs in absorbing RF energy and potentially converting it into localized heat, thus enabling targeted ablation of cancerous tissues. Through simulations using HFSS, the CNTs were observed developing oscillatory electric field and charge distributions at opposing end, a dipole-like behavior, when excited by an electric field. This electromagnetic-oriented analysis helps to confirm how CNTs might absorb RF energy and convert it into heat, supporting past observations and potentially aiding future researchers seeking to exploit the unique properties of CNTs.

The design and fabrication of the parallel plates was a critical step towards achieving a localized and targeted electric field. Initial iterations of the plate design used prior to this work demonstrated suboptimal field concentrations and broad dispersion. To address these shortcomings, the plates were redesigned with significant improvements. The inclusion of a spacer was shown to play a significant role in shaping the electric field distribution between the parallel plates. This behavior was, in part, linked to the dielectric constant of the spacer. These changes, validated through simulations, resulted in greater field intensity between the plates while also reducing the field intensity outside the plate and by effect, a reduction in unintended exposure. The new plate design was fabricated and tested, and although the plates exhibited some variability compared to simulations, their overall performance aligned with expectations in order to meet design objectives.

A key component of this work was the design and optimization of a tunable impedance matching network. Research started with an analysis of the Comdel MatchPro CPMX-6000 system, intended to work at a single frequency. Simulations in ADS provided a clear understanding of how individual components influenced the ability to match to a variety of impedance. These simulations and research helped to identify areas for improvement, and a matching network was developed using a simplified configuration that showed an ability to match across a broad frequency range. The matching network incorporated a combination of shunt capacitors, series capacitors, and shunt inductors. The network was fabricated and tested, demonstrating an ability to create a match across the target frequency range. This network provides a critical step toward a fully realized dynamic RF system, particularly as commercially available matching networks are typically designed for fixed-frequency applications. This tunability allows the system to dynamically adapt to evolving material properties, such as the dielectric constants of CNT suspensions expected in future work, enabling energy delivery across a broader range of applications.

The matching network, parallel plates, and ancillary components required for operation

were then integrated into a cohesive system. S-parameter data collected through a VNA in parallel with voltage measurements collected through an oscilloscope helped in providing an analysis of the response of the network, load, and system as a whole. Additionally, calculations were made from both simulated and experimental datasets to provide a quantitative analysis of voltage, current, and power through the network and into the load.

In evaluating the experimental results, the voltages measured directly on the plates were, in part, used as metrics of performance. This approach differs from many prior works, which primarily focus on the resulting heating effects. By focusing on the characterization of the voltages and field strength, this research offers an understanding of the system and induced field behavior at low power levels. The use of S-parameter data for approximate power calculations provides a validated approach for assessing power transfer efficiency in an experimental setting. The comparison between simulation-based and experimentally derived power calculations showing an agreement between the estimated power at the plates and the power dissipated in the tissue sample reinforce this method. The ability to systematically quantify power flow through the system using measured S-parameters enables a deeper understanding of how the impedance-matching network enhances energy delivery, bridging the gap between theoretical matching conditions and real-world performance. While high-power testing is ultimately necessary to fully assess the system's heating capabilities and its effectiveness in transferring energy to a material under test, the system's performance characteristics and behavior have been evaluated in detail, demonstrating how the network and plate design facilitate energy transfer even in the absence of observed heating.

The included discussions on nanotube heating, dielectric heating / behavior, RF effects on tissue, and regulatory limits on RF exposure, while outside the primary scope of this work, help to provide critical context for understanding the broader implications of this research. These sections also serve to better inform future engineers and researchers working

on subsequent iterations of this system, especially those unfamiliar with the contexts of its applications- by presenting relevant background information, this thesis offers insights into the challenges and considerations that must be addressed.

Before this body of work, commercially available RF-based systems for similar applications lacked the ability to optimize across multiple frequencies dynamically, relying on static configurations that limited flexibility and efficiency.

After this work, the presented matching network and refined plate design provide a foundation for future RF systems capable of adapting to diverse material conditions. These scientific contributions mark a significant step toward the development of fully realizable, tunable and efficient RF system.

6.2 Future Work

Despite the results and success of this thesis, several opportunities for refinement and further development present themselves to enhance performance, versatility, and applicability.

One area for future refinement involves improving the components used in the system. While the current design prioritized practicality, cost, and ease of implementation at an early stage, these constraints necessitated trade-offs in the component selection. For example, the use of more tunable and/or less lossy components, such as vacuum capacitors akin to those employed in the Comdel system, could improve efficiency and performance. However, such design choices would also necessitate additional trade-offs. For instance, the Comdel system required the use of an additional water cooling loop in order to function properly. Another avenue for improvement is the potential development of custom designs to expand the inductance tuning range, providing greater flexibility in impedance matching. The addition of a transformer between the matching network and the plates warrants

further exploration as well, acting to scale the impedance seen by the matching network.

Future iterations of the plate design could also be explored, tailoring the geometry to optimize field characteristics for specific material properties and test setups. Continued improvement could refine the electric field distribution and flux between the plates, potentially improving the system's efficiency and selectivity for particular materials, applications, and environments.

The addition of more complex and automated monitoring systems represents another opportunity for improvement. Currently, measurements rely on manual processes, which, while effective, are labor-intensive and less suited for high power testing. Real-time monitoring systems, even capable of dynamically adjusting operating parameters based on feedback, would improve both safety and efficiency.

After addressing some, if not all, of these challenges in future work, testing at high power will remain critical in fully characterizing the system's performance and validating its capabilities. High-power testing would allow researchers to assess the system's ability to induce heating effectively and safely in target materials, particularly CNT suspensions or biological tissues.

Lastly, future work could extend the applicability of subsequent iterations of the work presented in this thesis. Beyond its potential use in biomedical applications, such as targeted RF heating for cancer therapy, the tunable system could be adapted for industrial processes discussed in the literature-improving the performance of existing systems designed for composite curing, additive manufacturing, and selective material processing.

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