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A Novel Method for Tuning a Transistor-Based non-Foster Matching Circuit for Electrically Small Wideband Antennas

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Abstract

This dissertation reviews the application of non-Foster circuits for wideband antenna matching, and introduces a novel, rapid means of "tuning" the circuit to accommodate variations in antenna loadings. The tuning is accomplished via the judicious addition of a common transistor.

A detailed literature search is provided, and non-Foster circuits are discussed in detail, including the myriad of implementations with focus on tuning. A comparison between different tuning methods is presented. The novel tuning method is evaluated via the normalized determinant function to ensure stability.

Evaluations include simulations using commercially available software and experimentation to ensure not only stability but also that noise added by the active circuitry is manageable. Wideband stable operation is confirmed by pairing the tunable non-Foster matching circuit with an electrically small, resistively loaded dipole, and performance gains are demonstrated using the tunability feature. The resistively loaded dipole alone demonstrates reasonable performance at higher frequencies, but performance degrades considerably at lower frequencies, when the dipole is electrically small. The tunable non-Foster circuit is shown to alleviate some of this degradation.

Additionally, applications other than wideband antenna matching can benefit from tunable non-Foster circuits such as tunable filters and phase shifters, and these are discussed as well. Finally, practical limitations of non-Foster circuits are presented.

A Novel Method for Tuning a Transistor-Based non-Foster Matching Circuit for Electrically Small Wideband Antennas

by

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To my parents, Dr. and Mrs. Yassir Salama

for their support, dedication, sacrifice, and for being the best role models

and parents one could ask for,

and

To my young son Abdullah, and any future children, who I aspire to far

surpass their father's abilities and to follow the example their grandparents

have set.

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1. Introduction

The need to miniaturize electronics drives advancement in electronic devices. The improvement of integrated circuit technologies allows more complicated circuitry to be compacted in smaller packages. However, some things cannot to be miniaturized and remain size constrained. An example is the strict dependence of antenna size on resonant frequency. This presents a problem for miniaturizing devices that operate at lower frequencies. The resulting decreased-size wireless devices can sometimes utilize electrically small antennas (ESAs). An ESA is an antenna with a self-resonant frequency that is higher than the operating frequency. For example, a half-wave dipole antenna resonant at 10 MHz would have a length of 15 m. Using a half-wave dipole antenna with a length of 15 cm at this frequency would be considered an ESA. ESAs are attractive because self-resonant antennas can quickly become unreasonably large for many applications. Of course, ESAs typically suffer from low radiation resistance, and high Q [1, 2, 3].

The definition of an ESA is not unique. Burberry [1] adopts the definition as that of an antenna with aperture length of less than $\lambda/30$, where λ refers to wavelength. In 1947, Wheeler [2], who also studied the fundamental limitations of an electrically small antenna, suggested that an antenna is considered electrically small when the radius of the encompassing sphere is much smaller than the “radian length”, or $\frac{1}{2\pi}$ of the operating wavelength [2]. The smaller aperture, in relation to operating frequency, results in an ESA of typically low radiation resistance and relatively high reactance. This makes finding a suitable lossless matching network difficult because resonating the antenna results in additional loss, reducing the maximum radiated power. Further, even an ideal, lossless matching network typically imposes severe bandwidth limitations.

Electrically small antennas with appropriate matching networks have been used for many years. For example, an automobile AM radio with frequencies in the 500 kHz range is accompanied with an ESA. A half-wave resonant length antenna at 500 kHz would have a length of 300 meters. Conventional matching networks apply that result in a narrow instantaneous bandwidth but cover the operational band. The limited instantaneous bandwidth is key. This work addresses alternative methods of achieving wideband ESAs. The remainder of Chapter 1 reviews relevant conventional matching networks and topologies including bandwidth considerations, and Foster's reactance theorem. It concludes with a description of the format of this dissertation.

1.1 Conventional Matching Networks

Conventional matching networks rely on the Q factor of the load to create a matched circuit. Generally, these matching networks consist of an LC circuit which is placed between the load and the generator. The goal is to conjugate match the source impedance to the load impedance to ensure maximum power transfer from the generator to the load. The Q factor, which can generally be defined as the ratio of reactance to resistance, can be a very strong indicator of the bandwidth of the resulting combination of the matching network and antenna. For an individual passive component, the Q factor can be defined as:

$$Q(\omega) = \frac{X}{R} \tag{1.1}$$

where the impedance is Z = R + jX. An ESA will typically have a small radiation resistance and a large reactance, which results in a large Q factor. The Q factor can be an indication of the frequency and size disparity of an ESA. Therefore, the fractional bandwidth of an antenna is inversely proportional to the Q factor. Typical generators will have a near zero reactance with a modest resistance, and the larger the difference between the load and generator, the more complications will result when utilizing a conventional matching network. Matching networks that need to resolve a relatively minor difference in Q of the load and generator can provide a practical solution to obtaining maximum power transfer with little cost. It is important to note the frequency dependency of the equations dictating the Q factor. The components of the matching network are also dependent on frequency. The impedance of an LC network, comprised of inductors and capacitors, vary with frequency:

$$Z_L(\omega) = j\omega L \tag{1.2}$$

$$Z_C(\omega) = \frac{-1}{j\omega C} \tag{1.3}$$

where L is the inductance value in Henrys and C is the capacitance value in Farads. As the frequency varies from the designed frequency of the matching network, the match is expected to degrade. Unfortunately, the impedance slope across frequency of the passive matching network is counterproductive to creating a wideband match for typical scenarios. A much more detailed discussion on this follows in Section 1.3. Furthermore, an ESA will generally have a higher Q value at the desired frequency of operation which leads to complications of conventional matching networks, including bandwidth limitations.

1.2 Matching Network Topologies

There are several different common methods to implementing matching networks. Most involve adding sections of inductance and capacitance which result in an observed impedance that "matches" the desired value, implying that the load impedance is the complex conjugate of the antenna impedance. Lumped element versions use small discrete inductors and capacitors to form a matching network. Figure 1.1 shows the different combinations of matching networks referred to as "L Networks" [5].

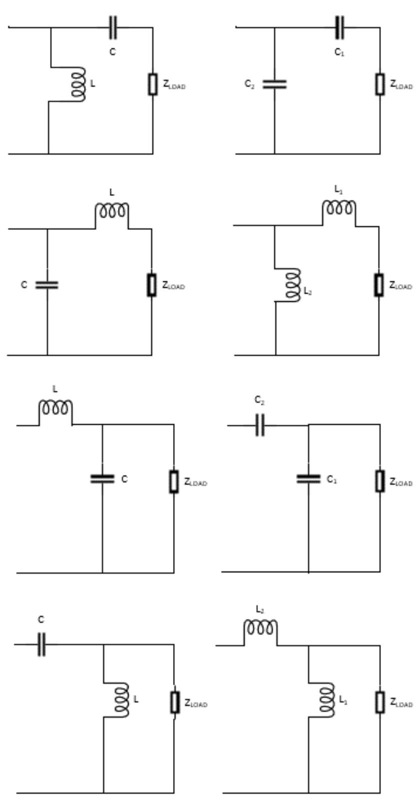

Figure 1.1: L Network configurations reproduced originally from [5]

The different L network combinations are used where appropriate. Some configurations are intended for circuits where the load impedance exceeds the source impedance, and other configurations are for the opposite scenario [5]. Using the appropriate circuit analysis equations, the values for the L and C components of the matching network can be determined to ensure maximum power transfer.

Sometimes, adding an additional component can help lower the Q value in addition to matching the source and load. This can be done with a Tee network or a Pi network, where three separate components are used for matching [5]. An example of this is shown in Figure 1.2. Once again, each configuration is used for a particular source and load impedance scenario.

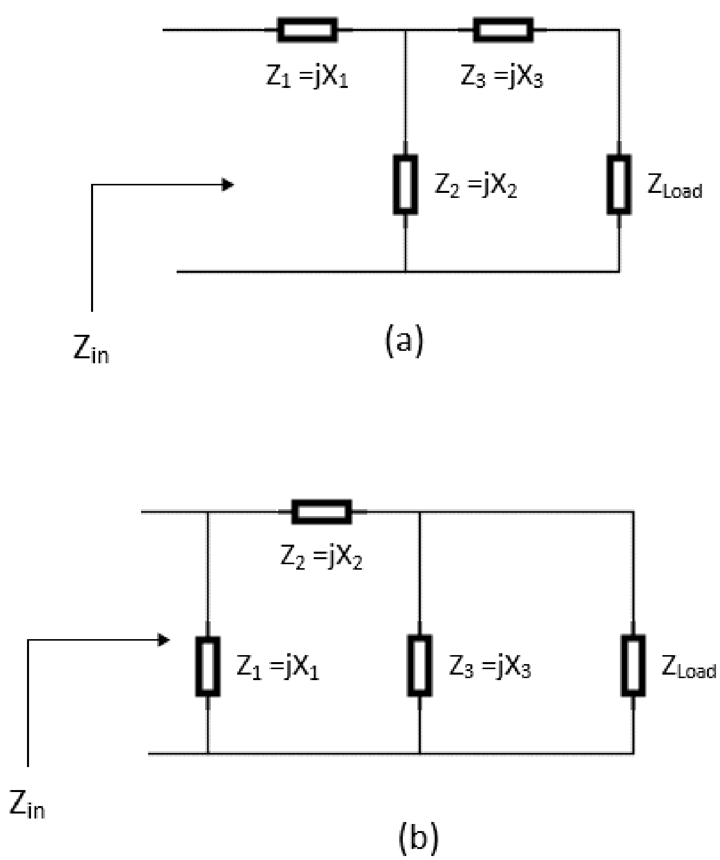

Figure 1.2: a) Tee network and b) Pi network reproduced originally from [5]

For cases where an antenna is printed on a circuit board, a transmission line matching network might be preferred. Since the antenna's impedance will be relatively constant in free space, transmission line matching networks provide a precise way to fix the input impedance of the antenna circuit. This allows the antenna circuit input impedance to be matched to a common source impedance with high precision. Lumped element matching circuits are forced to rely on common reactive element values which are available for commercial purchase, and which also will be affected by implementation challenges such as parasitic effects of soldering, etc. A transmission line matching network is not limited by these constraints. The width of a transmission line on a printed circuit board, which generally controls the characteristic impedance, and the length of the transmission line, which controls the distance traveled along the Smith chart can be chosen to match any arbitrary impedance. One of the major disadvantages of a transmission line matching network is its size.

Several considerations enter the choice of matching network, one is loss factor. The loss factor is defined as the insertion loss from the input to the output of the matching network. For applications where antenna gain is not a critical consideration, a larger loss factor might be tolerable. For example, receive-only applications could tolerate a higher loss factor if the system is externally noise limited. However, for transmit applications, a high loss factor can result in unacceptable heat dissipation.

Another closely related consideration is the power handling constraints of a matching network. A transmission line matching network usually can handle more power than a lumped element network with discrete inductors and capacitors. Because these high-frequency lumped components are typically small in size, they can quickly overheat when subjected to high power levels, whereas transmission line matching networks are typically constructed of large traces of copper which can better dissipate heat. On the other hand, as indicated above, the lumped element solution will require a physically smaller footprint than a transmission line solution. For antennas at high frequencies where an antenna itself is small, a transmission line solution might prove to be too large, even larger than the antenna at times which could dominate that trade-off study.

1.3 Bandwidth Limits of Conventional Matching Networks

With conventional matching networks, whether they are implemented with lumped elements, transmission lines, or another method, the impedance is matched at only one frequency. The Q factor of the load and source circuits should be taken into consideration as this will likely have a large effect on the bandwidth of the impedance. As the Q factor gets larger, the resulting matching network that is required to match the two impedances will yield a more narrowband match. By their nature, electrically small antennas usually exhibit a very high Q factor [1, 3, 4]. Because their size compared to that of a naturally resonant antenna is so small, the radiation resistance of the antenna becomes a less dominant part of the impedance when compared to the Ohmic resistance. Thus, besides being narrow band, ESAs are typically lossy. As discussed earlier, the resulting matching networks when conventional techniques are used will be narrowband. For applications where the bandwidth of operation is narrow, a conventional matching network might be a suitable solution. Radio receivers and other similar devices often employ these matching networks to mitigate the effects of electrically small antennas at the radio frequencies where wavelengths are impractically large for resonant antennas. However, applications which require wideband operation will need a different form of matching network or some other solution. Figure 1.3 shows the scenario where a capacitive load of 10pF is matched at 100MHz. Under the conventional matching techniques, one could use a 253nH inductor in series to bring the reactance to zero.

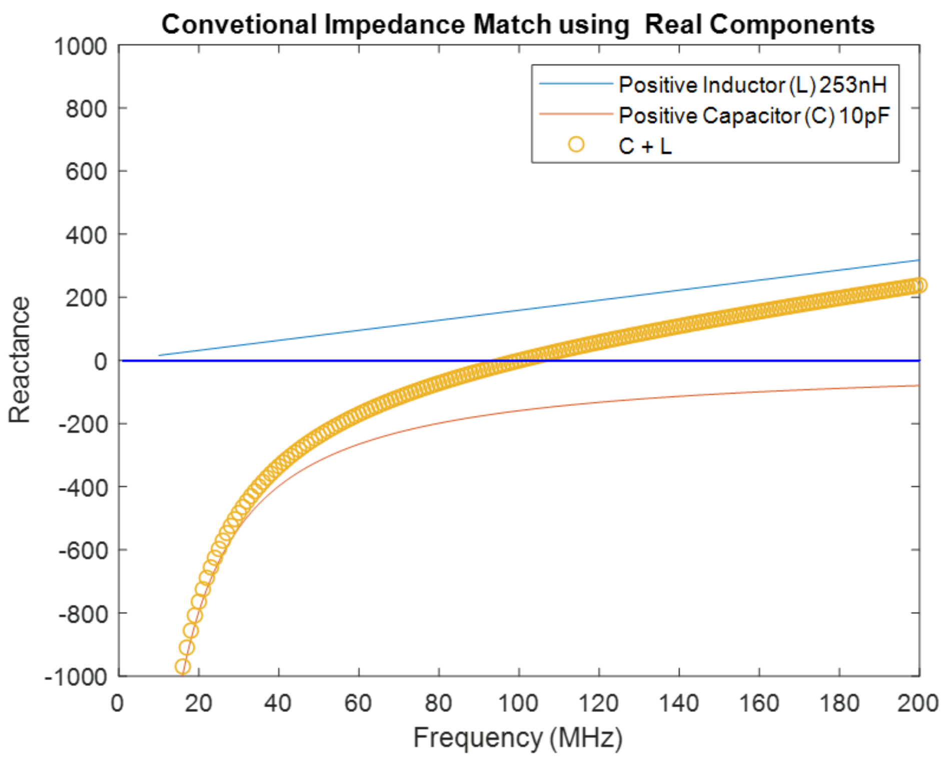

Figure 1.3: Conventional matching circuit

As shown in the yellow curve in Figure 1.3, the reactance is only minimized for frequencies around 100MHz, and becomes quickly unmatched as the bandwidth increases. The blue and red curves correspond to $Z_L(\omega) = j\omega L$ and $Z_C(\omega) = \frac{1}{j\omega C}$ respectively. The bandwidth of this match can be related back to the Bode-Fano criterion [6], where loads which have a higher initial reactance (or Q factor) will require a larger reactance matching circuit and typically be more limited in bandwidth. Bode [6] related the reflection coefficient to an impedance with the following equation:

$$\int_0^{\infty} ln\frac{1}{|\rho|}\, d\omega \leq \frac{\pi}{RC} \tag{1.4}$$

where ρ is the reflection coefficient and R and C are the shunt resistance and capacitance values of the network, respectively. Even a typical 10% bandwidth operation, which in this case would be 95-105MHz could be well suited for a conventional matching network. However, as mentioned earlier, more and more miniaturized electronic devices are demanding wider bandwidth and using higher-Q antennas. Section 1.4 will introduce a class of active matching networks under study to solve this problem.

1.4 Non-Foster Circuits

Conventional matching networks like the ones referred to in the previous section use passive elements to conjugately match load and source impedances. The reactance of all these elements monotonically increases with frequency according to Foster's reactance theorem [7]. Non-Foster circuits are those which do not follow Foster's reactance theorem, in that the curve of the reactance versus frequency is not necessarily monotonically increasing with frequency. The ability to use non-conventional impedance elements to match a network opens a new set of possibilities with matching networks. The Bode-Fano criterion may not be a limiting factor to the resulting bandwidth of a non-Foster matched circuit. In particular, the use of inverted impedance elements, or 'negative impedance' as they are sometimes referred to, allows the matching circuit to additively match a load circuit across a wider range of frequencies, unlike the result of Figure 1.3. The match (referred to here as zeroing the reactance) can be achieved in several different ways, but all of these require the use of active circuits or active metamaterials [8]. Using an active circuit with a seemingly 'negative element' or inverted impedance curve, a circuit can theoretically be matched perfectly over a wide band. This is illustrated in Figure 1.4, where the same mismatched load is now matched with a negative valued capacitor. Rather than applying the 253nH inductance to cancel the 10pF capacitance for the example provided in Figure 1.3, an ideal -10pF capacitor is placed instead of the matching circuit. In this scenario, the circuit would theoretically be matched across all frequencies.

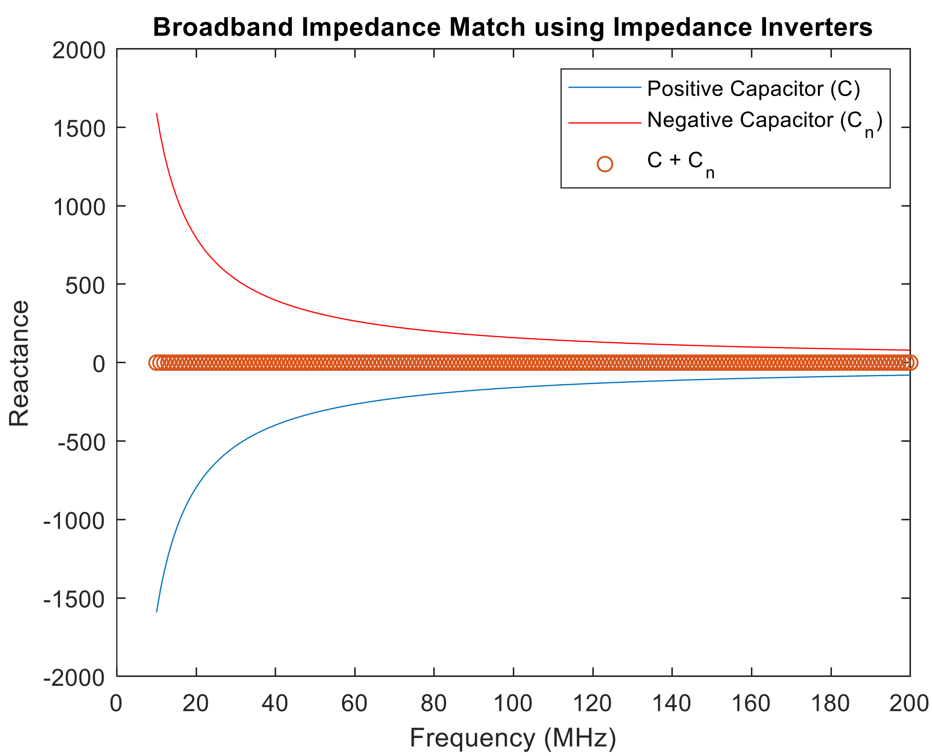

Figure 1.4: Inverted impedance matching network

As mentioned earlier, an active circuit is required to achieve this inverted impedance. There are several different ways this is made possible. Some use transistors, others use Op Amps, and there are other even less common ways of achieving inverted impedance such as the use of metamaterials [8, 9]. Note that the example shown in Figure 1.4 is for an ideal case, and practical cases differ from this ideal scenario for two main reasons. The first reason is that the inverted impedance circuit does not perfectly imitate a negative-valued impedance element. As the impedance further varies from the ideal target, the bandwidth would then be limited by the fidelity of the inverted impedance circuit. This could be largely dependent on the topology used for achieving the inverted impedance, as each method could be subject to different bandwidth limitations, such as the bandwidth of transistors used in these circuits. The second reason for a less than optimum match is that the original load impedance does not perfectly match that of a passive lumped element component. The case of the original load impedance being similar to a lumped component is not necessarily a reasonable assumption. In addition to these two reasons, it is important to understand that in practice there are many other constraints that cause active circuits to be less than ideal.

Thus, even with the assumption that the active circuit matches the ideal inverted impedance closely, and that the load is largely representative of a positive-slope reactance element, there are other bandwidth limiting considerations. In practice, an active circuit will have bandwidth limitations of its own. Whether that be a transistor based non-Foster circuit, or an op-amp based circuit, these are non-linear devices that have operational frequency curves specific to each device. For example, this could become evident at higher frequencies that are outside the transistor's finite operating band. Additionally, the stability considerations of the active circuit are of particular concern. Once this active circuit is placed in a matching network which has a resulting total zero sum reactance, it could induce a resonance which will render the active circuit unstable [10, 11]. Therefore, it is more practical to design for low reactance than to achieve zero reactance. Although not being a perfect match for a circuit, this could produce a much lower Q and broader bandwidth than that of a conventional matching circuit which uses passive components. A reoccurring theme when experimenting with active circuits is that any decrease in the magnitude of the combined reactance will result in a wider bandwidth match. That is, any reduction in reactance will lower the Q value of the circuit pair and enable improved power transfer between the two parts.

Another consideration is the potential for added noise arising from the active elements. Any additional noise presented by the active circuits will have a direct impact on signal to noise ratio. Active circuits will always contribute additional noise resulting from the injected power of the active circuit. For non-Foster matching networks, the primary concern is improving gain at the lower frequencies to mitigate the poor gain of an electrically small antenna. At the higher frequencies, when the antenna becomes self-resonant, the need for gain improvement diminishes. An active non-Foster matching network will inject unwanted noise across much of the frequency spectrum. This is based on the active components themselves as well as the characteristics of the supplied "DC" voltage. The supplied DC voltage is frequently generated from another DC voltage or even an "AC" voltage, and modern DC-DC or AC-DC converters have ripple on the output voltages which directly effects the amplitude and frequency of the added noise. For an active non-Foster circuit to provide benefit, the added noise at the frequencies of interest needs to be less than the gain improvement. For the higher frequencies, the added noise will indeed provide a degradation of performance which is traded for the improvements of the lower frequency region. An in-depth noise analysis is discussed in Section 5.2.

1.5 Related Work

Non-Foster networks are not limited to antenna matching networks, rather they have a variety of applications. The first types of applications for non-Foster networks pre-dated the need for wideband antennas and were for improving telephone communications [12]. Not only do the applications of non-Foster networks vary, but also the method of which networks achieve non-Foster behavior is also application specific. One of the first practical applications of non-Foster circuits was developed by Linvill in 1953 [12]. Linvill used transistors to achieve an impedance inversion, and in common knowledge, he is one of the first to publish such a circuit that was put into practice. Linvill presented two circuits, one of which was deemed open circuit stable (OCS), and the other was short circuit stable (SCS), as shown in Figure 1.5. He discusses in detail the properties of the transistors he chose and how they can achieve an 'impedance inversion'. The principal idea behind this circuit is the use of a pair of common emitter amplifiers which sample the voltages across a load. The outputs are crossed to provide an inverted voltage at the input terminals. Since the current through the load is unchanged, the input impedance then becomes the load current divided by the crossed voltage, and thus the inverse of the load impedance. The OCS and SCS cases vary by the input and load locations, where the OCS has the input at the emitter of the transistors and the SCS case has the input at the collector of the transistors.

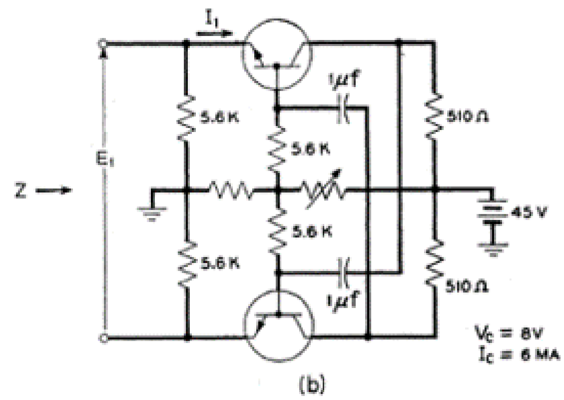

Fig. 1(a) and (b)—Open circuit stable negative-impedance converter and locus of its input impedance.

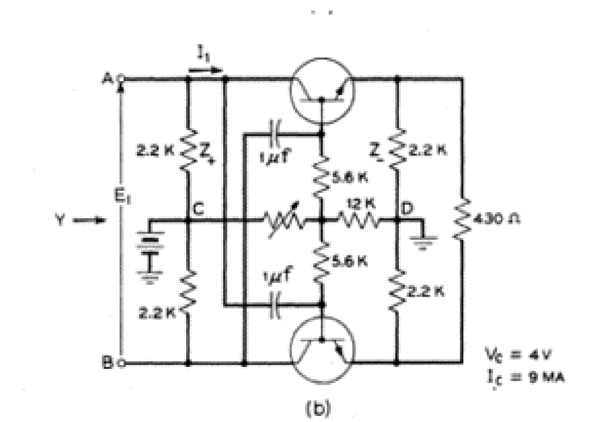

Fig. 3(a) and (b)—Short circuit stable negative-impedance converter and locus of its input impedance.

Figure 1.5: OCS and SCS negative impedance inverter circuits, reprinted from Linvill [12].

Linvill's work primarily focused on the transistor-based negative impedance converter (NIC). His example target applications were for transmission lines for utility companies, however his paper mainly focused on the transistor implementation and didn't mention antennas as a specific application. Steven Sussman-Fort and Ronald Rudish were some of the more recent advocates of matching ESAs using non-Foster matching networks to achieve a broadband match [10, 11]. They have published a multitude of research articles on different aspects of using NIC circuits to improve antenna performance, only some of which are cited here. However, shortly after Linvill's paper was published in the 50s, there were some early pioneers of this technology as applied to antennas versus the typical application of telephone transmission lines. Sussman-Fort and Rudish give a nice summary of the early work done with non-Foster matching circuits for ESAs [10]. In 1968 Harries and Meyers [13] were some of the first people to develop a negative impedance inverter used for improving the performance of a miniaturized antenna. They compared the results of an NIC actively matched short monopole with a top hat to that of a conventional 16' whip antenna and showed an improvement in receive gain over the long, non-matched case. They acknowledged that noise needed to be measured, but in their particular application at low frequency, they were not signal-to-noise ratio (SNR) limited, but interference limited, so they didn't perform a noise analysis [13]. A year later, in 1969, Poggio and Mayes [14] used an inverted impedance circuit to broadband match a dipole antenna. Many articles have since described the use of non-Foster networks for antenna matching, even until recent times. Reference [15] is a recent work where a non-Foster circuit was used in conjunction with an ESA to improve performance. In this example, they required a resonant 900MHz antenna. They created a miniaturized version of a 900MHz antenna that is naturally resonant at 2.2GHz by using a non-Foster matching circuit.

Our work presents a non-Foster circuit with a unique means of "tuning" (adjusting the input impedance of the non-Foster circuit to better match the impedances of a variety of devices in a variety of environments) as described in Chapter 4. Non-Foster circuits which have adjustable input impedance have been previously studied. These circuits are very useful for applications where the impedance of the load might not be constant or for the general ease of having a one-circuit-fits-all solution. Most of the previous solutions utilize a varactor or a vericap which changes the load impedance that the circuit will invert at its input [16, 17]. Others used the approach of a switched capacitor bank to change the load impedance of the NIC and therefore create a tunable negative impedance [18].

There also have been more creative solutions to achieving a tunable inverted impedance. Tian et al. [17] show a nice comparison of changing the load impedance of an NIC to achieve a tunable inverted impedance to changing the bias voltage of the transistors in the NIC circuit to have the same effect. They used a graphene-based transistor which might have contributed to the positive results. While the simulations where the bias voltage of the transistors in the NIC was varied to achieve an impedance inversion worked, however, the curve only very loosely fit that of an ideal impedance inversion. Thus, broadband matching could be difficult.

Other approaches to achieving a tunable impedance inversion circuit utilized different types of non-Foster circuits. Fan [19] experimented with leaky-wave antennas using metamaterials as non-Foster components and was able to achieve a tunable bandwidth with a varying biasing voltage. This is a type of active matching circuit which can be classified as a non-Foster impedance match. White et al. [20] used a proprietary IBM process to achieve a tunable non-Foster circuit using transistors in an operational transconductance amplifier (OTA) fashion. Unfortunately, there is little information about the IBM concept. It looks like they used multiple stages of non-Foster networks to create something tunable. However, their only published paper provided very little explanation of the circuit. Dhar et al. [21] followed this work by using this model to improve the performance of an electrically small antenna.

A general summary of the previous work in non-Foster circuits which are tunable is relevant here. A detailed summary of these tuning methods that discusses the advantages and limitations of each approach will be deferred to Section 3.3.4. Most literature focuses on using non-Foster circuits for particular applications. There are, however, only a few cases where the idea of a tunable non-Foster circuit has been discussed, and these cases are mentioned previously. In many of these instances, the tunability is not the general focus of the work, but merely a side issue. Therefore, most of these cases use the simple solution to create a tunable non-Foster network, and that is by the use of a varactor. A traditional, non-tunable non-Foster network uses a load element which has its impedance inverted to create a ‘negative impedance.’ The non-tunable circuits required changing the load element to achieve a different negative impedance. The simple solution to adjusting this ‘negative impedance’ value is to vary the load element’s impedance by using a varactor, which acts as a variable capacitor using a changing voltage. This is a valid solution for obtaining a changing impedance, and the solution presented in here in Chapter 4 serves as an alternative. The main disadvantage of the varactor method is that the parts for constructing the matching circuit become specialized and may not be commonly available depending on the application. A related approach is to use a switched capacitor bank rather than a varactor. The overarching idea is the same as a varactor, however this uses discrete components and a pair of switches that are all contained in a single package [18].

Another solution which uses specialized parts is the solution proposed by White et al. [20]. They use a proprietary IBM part that is not readily available to the general consumer and the inner operations of it is not explained in enough detail for the purpose of recreating it. As a result, simulation or measurement data based on this approach is limited to one article and has not been found elsewhere in the literature.

Other solutions which propose changing the bias voltage of the main transistors to achieve some tunability range suffer in their performance. Their advantage is simplicity. Their main deficiency is relying on bringing the transistor in and out of different operation regions to achieve a different impedance at the circuit input. This in turn makes the final result only very loosely match an ideal impedance curve. As mentioned above, a detailed summary of these tuning methods that discusses the advantages and limitations of each approach will be given in Section 3.3.4.

1.6 Format of the Dissertation

This dissertation considers the problem of wideband matching for electrically small antennas. Some of the motivating factors for wideband matching networks are discussed. With the push to miniaturize electronic devices and the physical limitations of wavelength, the need for electrically small antennas has grown. Furthermore, devices that incorporate multiple missions commonly need to span a large bandwidth. Therefore, the problem of utilizing an electrically small antenna is a difficult, but important, problem that this dissertation addresses.

Included earlier is a discussion of conventional matching networks and their associated limitations. The concept of non-Foster networks was introduced. A detailed literature search was given with all related works that have addressed non-Foster matching networks. Many different applications for non-Foster matching networks were discussed as well as different topologies. Previous work, both research in academia and papers produced from industry, was examined. Each scenario was evaluated for its respective benefits and limitations.

This dissertation focuses on one specific application of non-Foster networks: antenna matching. Therefore, Chapter 2 presents different methods of matching antennas. The goal is to maximize the power transfer to and from the antenna. This can be done in several different ways, some more common than others. A brief discussion on non-Foster networks specifically for the purpose of matching antennas is discussed. Chapter 2 concludes with a comparison between the different methods of matching antennas.

Chapter 3 gives a detailed discussion on negative impedance circuits. Linvill's transistor-based circuit is only one of many different types of non-Foster circuits. Within the category of non-Foster circuits, Linvill's circuit is also only one of many different types that can be used to match antennas. This chapter spells out the different types of negative impedance circuits such as the transistor-based circuits and the op-amp-based circuits. A discussion of the different types of tunable circuits is given. A literature search on many different works that have discussed the idea of tunable non-Foster circuits is indicative of the significant interest in the tunable aspect of non-Foster circuits.

Chapter 4 introduces the major contributions of this dissertation. A new type of tunable non-Foster circuit, which is based on Linvill's transistor circuit is given. A "tuning transistor" is added for tunability. This novel approach is recently patented by the author [22]. A thorough analysis is given. A rigorous and in-depth stability analysis is performed, and this circuit is shown to be stable under all conditions. This is critical as many active non-Foster circuits suffer from instabilities in some region of their operation. One of the leading limitations of non-Foster circuits is their stability considerations and introducing an unconditionally stable circuit relieves many of those concerns. A small signal circuit model is applied in evaluating the stability of the circuit, in addition to other parameters such as noise and transient characteristics. Modeling and simulation results are shown. Also included are measured results of a physically realized circuit that have shown to produce improved performance.

Chapter 5 provides practical applications of tunable non-Foster networks. We confirm that wideband matching of antennas is realized. First, a wideband resistively loaded antenna is discussed including simulation results that demonstrate wideband performance. Following that, the novel tunable non-Foster network is used to match the resistive antenna over a wide bandwidth, and both simulation and measured results are shown. Additionally, other practical applications of the tunable non-Foster circuit such as tunable filters are presented. The simulation results demonstrate that the novel tunable non-Foster network introduced here has applications beyond antenna matching.

Chapter 6 discusses some of the limitations of non-Foster circuits. Regardless of the positive and negative contributions of non-Foster circuits, there are scenarios where non-Foster circuits simply cannot be used. Further research or a change in fundamental operation would be required to use a non-Foster matching network in these applications.

Chapter 6 continues with practical considerations of using non-Foster matching networks for wideband antennas. Issues that were experienced firsthand in fabricating, building, and measuring the combination of both the resistively loaded dipole and non-Foster matching network combination are detailed in Section 6.4. Although many applications can imply good results with simulation data, there are often factors on physical implementation that serve as roadblocks to achieving the simulated results. Finally, in the Conclusion, a summary is provided of the accomplishments given in this dissertation, followed by a discussion of future work.

2 Antenna Matching Methods

In any system that transmits wirelessly, the antenna and the electronics must be impedance matched to ensure the most efficient transfer of power. In DC circuitry, the amount of power delivered to a component is a ratio of resistances. For higher frequency operation, reflections between components which are not perfectly matched becomes a concern. The resulting difference in impedances of components is proportional to the reflected power. The reflection coefficient, Γ, is described as:

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{2-1}$$

where Z_L is the load impedance, and Z_0 is the characteristic impedance. Despite best matching efforts, all systems will experience some reflection and the objective would be to minimize that reflection to ensure maximum power transfer. Any reduction in power transfer requires an increase in power to reach target performance. Likewise, this may require more amplification thus increasing noise with consequential reduction of signal to noise ratio upon receive. Antenna matching has been studied for many years. Different methods for antenna matching have been introduced based on the specific constraints of each system. Chapter 1 briefly mentions the conventional matching network and the use of discrete components for matching two impedances. Other methods for optimizing performance have been developed and some will be discussed in the following sections.

2.1 Other Approaches of Wideband Operation

Wideband operation of antennas drives innovative methods for antenna matching. Active matching networks offer a convenient solution for the circumstances discussed earlier, in particular, low frequency, electrically small, multi-octave bandwidth operation. In the interest of providing a comprehensive picture of available tools, alternative approaches to wideband operation will be discussed in this chapter. There are several reasons for using active circuits, and these will be discussed in a later chapter. However, some scenarios are better suited using passive circuits. This could be because of the requirements not needing active matching networks, or even because of some of the limitations of active circuits. Nevertheless, other approaches for wideband operation have long been studied and each offers its own advantages and disadvantages.

2.1.1 Antenna Geometry

One of the earliest stages of antenna design is constructing the best physical geometry to make it usable for practical circuits. Antennas can be considered impedance transformers in themselves, transferring from the impedance of free space (around 377 Ohms) to that of a source or generator. Because most RF receiver circuitry is set at either a 50-Ohm or 75-Ohm characteristic impedance, transforming to these impedances is a good target. Antenna input impedance rarely matches this impedance exactly, but the closer the value to the characteristic impedance of the receiver, the lower the Q value of the matching network needed which results in wider band operation. Recall Bode’s criterion (equation 1-4), which relates bandwidth to the reflection coefficient. As such, innovative techniques to constructing the geometry of the antenna can be utilized to force the antenna input impedance to be closer to that of the receiver. One example of a way to optimize antenna geometry is to carefully choose conductive “trace” lengths and widths of the antenna to match that of a particular characteristic impedance. A great example of choosing a clever antenna geometry for wideband operation is that of a Vivaldi antenna [23]. Gibson created this antenna in 1979 and demonstrated vast bandwidth gains compared to other antennas of that era. A Vivaldi antenna is a form of a dipole with flared arms that transform the impedance from the source to free space. Figure 2.1 depicts a simple Vivaldi antenna, as demonstrated by Gibson in his 1979 publication [23]. His work includes a comprehensive guide on altering the shape of the flare in accordance with frequency requirements, and beamwidth (or gain). Since its introduction, the Vivaldi antenna has become a staple for wideband operation. However, a key disadvantage of utilizing an antenna geometry method for wideband operation is that it will need to have an appropriately sized aperture dictated by the lowest frequency of operation. This means that for very low frequencies, these antennas are going to be prohibitively large.

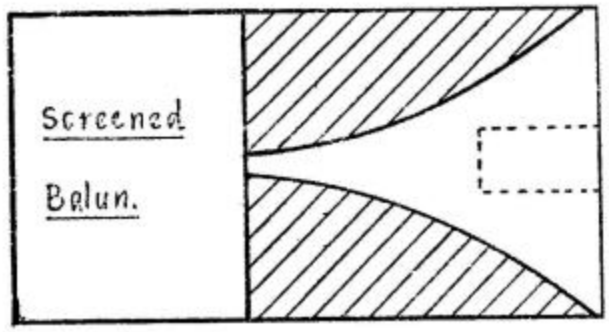

Figure 2.1: Vivaldi antenna, reprinted from [23].

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When considering antenna geometry for wideband operation, special care must be given to the feed network of the antenna. A poorly designed feed network can limit wideband operation. Feed networks come in many different forms. An abundance of literature has been published on wideband feeds of different kinds. Certain trade-offs are well understood among antenna designers, and one is size vs bandwidth. One method is to use a transmission line impedance transformer. A quarter wave transformer is unique in that the characteristic impedance of the transformer can be chosen to perfectly match the antenna and source impedances at a single frequency point. The input impedance of a transmission line can be described by

$$Z_{in} = Z_O \frac{Z_L + iZ_O \tan(\beta L)}{Z_O + iZ_L \tan(\beta L)} \qquad (2\text{-}2)$$

where Z_{in} is the input impedance looking into a transmission line, Z_O is the characteristic impedance of the transmission line, and Z_L is the impedance of the load (at the other end of the transmission line), L is the length of the transmission line and $\beta = \frac{2\pi}{\lambda}$ where λ is the wavelength.

After evaluating the limit of the tangent function as it approaches $\frac{\pi}{2}$, it is evident that, for a quarter wave transmission line, where $L = \frac{\lambda}{4}$, this equation reduces to

$$Z_{in\,(quarter\,wave)} = \frac{{Z_o}^2}{Z_L} \tag{2-3}$$

Using equation (2-3), a ¼ wave section of transmission line with a carefully chosen characteristic impedance can provide a reflectionless transition between two different impedances at one frequency. For an antenna with an input impedance different than that of a source, this could be a good candidate for a feed. Continuing the trade-off considerations, one can trade size for bandwidth. For high Q scenarios where the impedances are vastly different, rather than using a single stage impedance transformer, one can utilize multiple sections to transform the impedance in steps and realize a wider bandwidth match. Different intermediate impedance steps can be used for different requirements. Figure 2.2 shows the schematic representation of two different ¼ wave transformer configurations. Figure 2.3 shows the simulated return loss results that compare the two different transformers. In Figure 2.3, the red curve is the return loss of a single section ¼ wave transformer and the blue curve is that of a two section ¼ wave transformer. The source and load impedances are 50 ohms and 30 ohms, respectively. This demonstrates the trade-offs between size, bandwidth, and performance. The 2-section transformer clearly spans a larger bandwidth but will require twice the space as that of the single section. Also, the 2-section transformer's performance is better across a wide bandwidth, but at the target frequency it does not provide any advantage over the single section. This simulation was done using ADS [24].

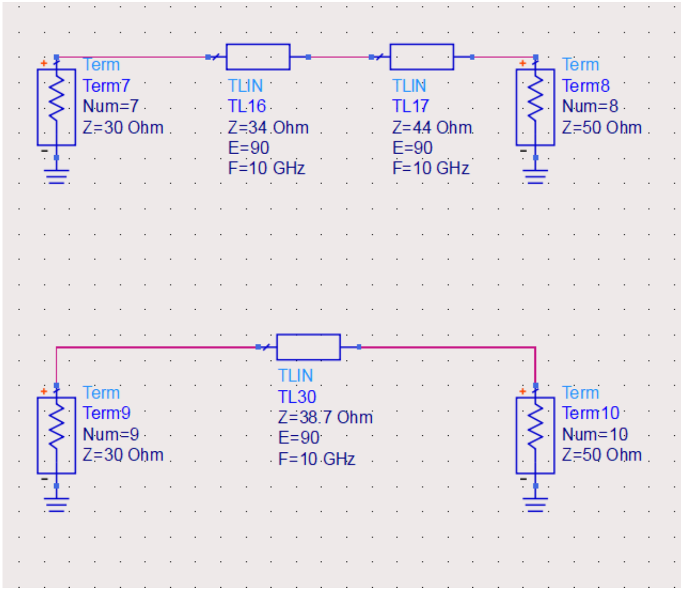


Figure 2.2: ADS schematic of quarter wave transformers

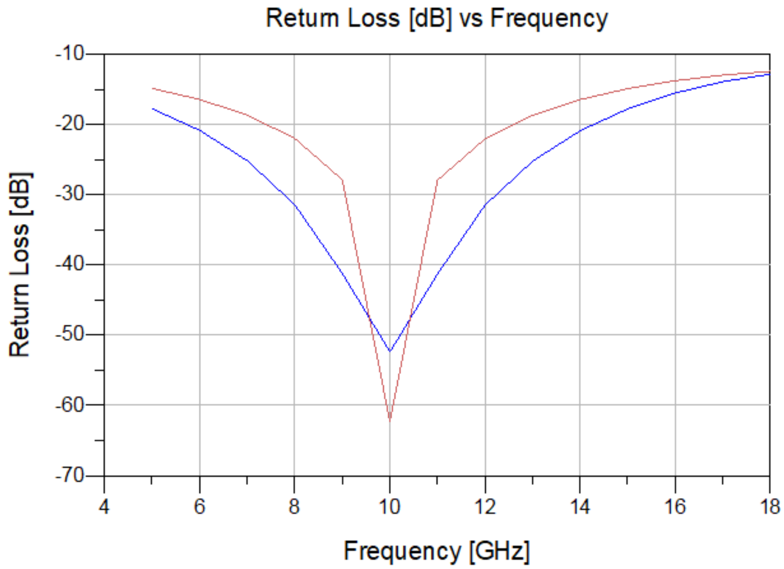


Figure 2.3: Return loss [dB] of one and two section transformers

2.1.2 Changing the Source Impedance

Antenna power transfer is a two-sided problem. The antenna and the associated circuitry need to be matched in impedance to maximize power transfer. Because of the common re-use of transmitter and receiver circuitry, antenna and microwave designers typically gravitate to altering the antenna impedance. However, there is an alternative. Rather than changing the antenna impedance, the source impedance can be selected to match that of the antenna. In some cases, an antenna designer might not be given freedom to shape the antenna for optimum performance.

In a cellular telephone, for example, varying antenna geometry for optimization can be difficult. Considering the compact nature of the phones and the surrounding components around the antennas, in addition to the variability of the user environment, this may be a scenario where the design revolves around the antenna. In general, if a specific input impedance is not required and the antenna geometry is not conducive to optimizing impedance, the source impedance can be chosen to optimize the entire assembly. This would present challenges with the transmitter and receiver design since most available components are already designed for a pre-determined 50-Ohm or 75-Ohm impedances. This limitation should not be underestimated. The number of available components that differ from these impedances is very small. Any common components or components which are already pre-matched would not be applicable.

Given this, one option is to design custom components for interfacing with the antenna. In some scenarios, commonly available components are not sufficient and custom component design is needed anyway. One example is a high power transistor for use on an RF amplifier. These transistors are not typically matched at 50 ohms. Part of designing a high power RF amplifier is to appropriately match them for the frequency of operation. Amplifier designers have historically tried to match them to 50-Ohms at the input and output. Having an antenna with a different input impedance than 50-Ohms can be matched using a custom design RF transistor. Similarly, receivers which might be used without commonly matched components such as low noise amplifiers (LNA) may benefit from this approach. One example of that would be a direct sampling receiver which would feed the antenna directly into an analog to digital converter. Analog to digital converters are generally not matched to 50-Ohms and require an impedance transformer, usually in the form of primary and secondary windings around a ferrite core at their input. The ratio of primary to secondary windings can be varied to create the impedance match. It is imperative to note that a receiver with an antenna connected directly to an analog to digital converter would suffer from very high noise-figure and other problems and is not considered as a good design in general. For the purposes of wideband antenna matching, it can however serve this purpose quite well.

2.1.3 Matched LNA Networks

The impedance matching methods discussed in the previous section had significant limitations because of the limitations of available components and of specialized parts. Since an antenna design is typically custom based on the application, it becomes natural to alter the antenna to match the circuitry. Standardized impedance across circuitry makes a designer’s job easier because the designer is not limited to a particular manufacturer or a specific category of parts. Mixing and matching parts from any source that they are impedance matched allows for more robust circuitry and optimized design. The designer loses these advantages using specialized circuitry. One approach some designers take is to implement matched low noise amplifier (LNA) networks. This approach is a nice compromise between both scenarios. No longer do the transmitter and the receiver need to be limited to specifically designed parts for a particular antenna.

The matched LNA network involves altering one component but allows the remainder of the receiver circuitry used to be common parts that are easily obtained. An example in Reference [25] shows a way to design an input and output matching network just for the LNA which allows for optimized performance from an antenna. This work uses a radial stub to match the input of the LNA and optimizes noise figure. This draws similarities to the non-Foster circuit which was introduced in Chapter 1, with more detail to follow. The non-Foster matching network is an active circuit which overcomes some of the limitations of passive circuits. The matched LNA network, being an active circuit, also enjoys relief from some of these limitations but introduces other problems unique to active circuits. One of the main shortcomings of active circuits in the antenna matching scenarios is their lack of reciprocity, implying that the input and outputs need to be clearly identified and that the circuit is not bi-directional. This means that the circuit must only be used for a receiver or transmitter, but not both. Conversely, passive matching networks can be utilized in both directions once implemented. Once again, the matched LNA network provides one method of wideband operation but does not serve to be an uncompromising solution. Normal bandwidth operation is considered to be about 10%, not adequate for the proposed work. Figure 2.4 shows the return loss of the matched LNA network when plotted across frequency. The matched LNA network of [25], shows a 3dB return loss bandwidth of around 20% and therefore can be classified as a wideband LNA. However, components utilized for wideband antenna operation have an order of magnitude better bandwidth as described in the following section.

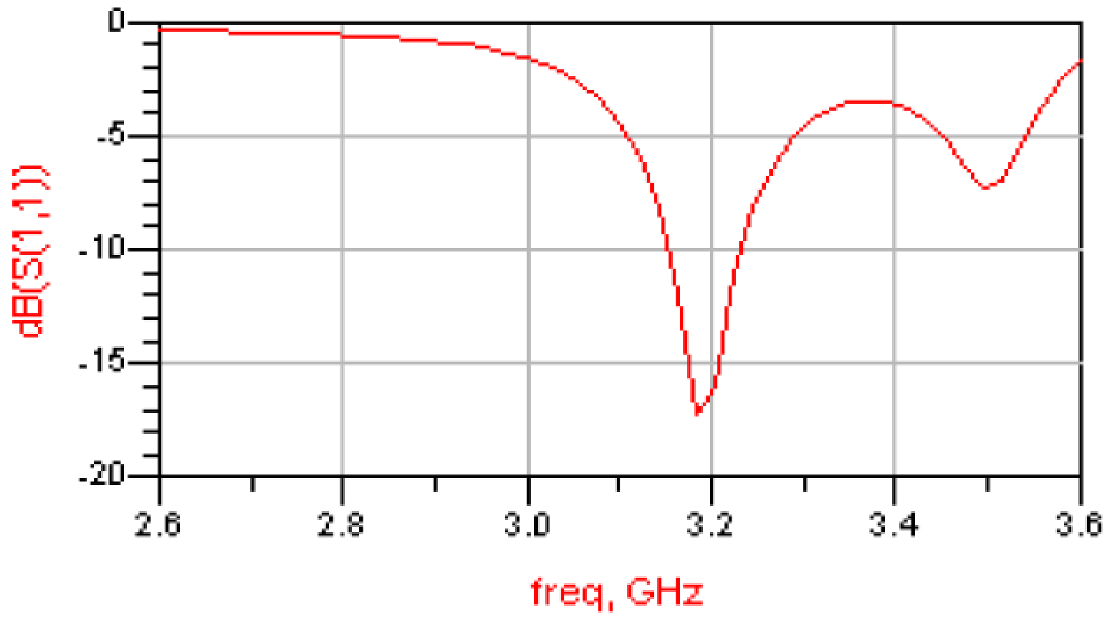


Figure 2.4: Bandwidth of radial stub matched LNA network, reprinted from [25].

2.2 Non-Foster Matching Networks

Section 1.3 and Section 1.4 described conventional matching networks, their bandwidth limitations, and non-Foster networks and the ability to circumvent these bandwidth limitations. This section will provide more detail on the use of non-Foster networks for antenna matching. For completeness, the advantages and disadvantages are spelled out here to serve as a comparison between the other antenna matching methods addressed. Even more detail on the specific implementation of non-Foster networks for antenna matching will be given in Chapter 3 and Chapter 4.

The discussion in Section 1.4 spells out some of the motivations for using non-Foster networks. Recall that the monotonically increasing slope of the reactance of passing matching networks is a limiting factor for wideband operation. When passive devices, such as antennas, are loaded with non-Foster matching networks, an improved match due to their conjugate reactance is observed. They can provide increased bandwidth especially regarding electrically small antennas operating at very low frequencies. However, real devices do not necessarily show a perfect conjugate reactance across all frequencies, and therefore gains can only be observed when the resulting total reactance is minimized. In fact, a major advantage of using non-Foster networks is improved performance at the lower frequencies. As with any active circuit, non-Foster circuits add noise when utilized for antenna matching. Understanding the noise contributions is critical to determining performance gains. Details and confirming measurements will be discussed in Chapter 5.

In the higher frequency region, however, the non-Foster circuit does not provide as significant an improvement, and in some cases can even degrade the match of the wideband antenna. Various reasons can explain this phenomenon. Most of these reasons are related to the practical implementation of a non-Foster circuit. There are several methods that can be used to implement a non-Foster circuit, some more common than others. Two of the most common methods, transistors and op-amps, are generally lower frequency devices that do not operate well at high frequencies. Of course, with careful selection of specialized devices, higher frequency operation *can* be achieved [26, 27]. Reference [27] targets other non-Foster applications, and while not necessarily relevant to antenna matching, these other applications can provide valuable techniques to mitigating some of the shortcomings of non-Foster circuits in antenna matching applications.

Alternative applications for non-Foster circuits are discussed in Chapter 5. They include negative group delay filters [27,28]. Therein, the active circuits can be adjusted to set the desired group delay in the filter. Likewise, non-Foster circuits have been experimented with in the use of wideband phase shifters [29]. Reference [29] also discusses the need for selecting transistors which suit the desired operating frequency. A more detailed discussion on alternative applications for tunable non-Foster circuits is continued in Chapter 5.

There are several different ways to implement a non-Foster circuit. The most common ways, which were alluded to earlier, are using transistors or Op-Amps. Other methods are mentioned in Section 1.5. This dissertation focuses on the transistor-based non-Foster circuits, originally introduced by Linvill [12]. However, other active devices have been used to implement a non-Foster circuit, like the op-amp circuit from [30] shown in Figure 2.5.

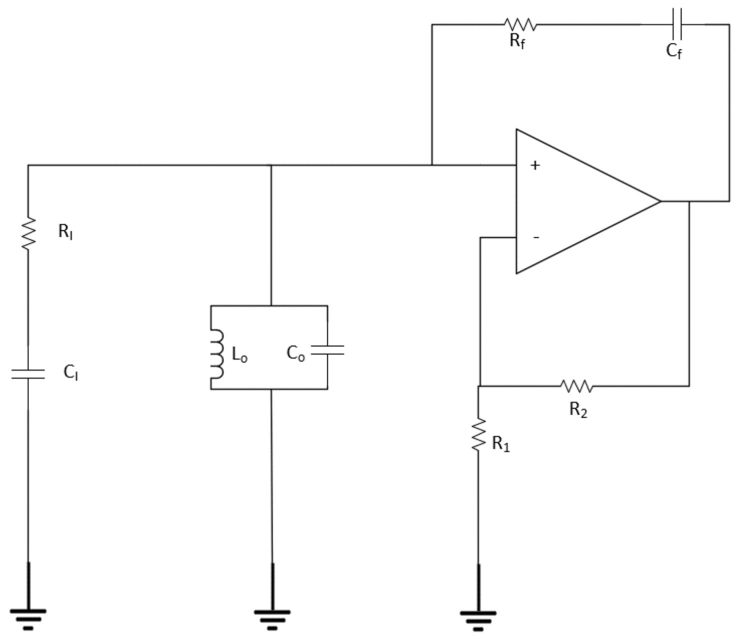


Figure 2.5: Hrabar's "antenna transmitter" circuit reproduced originally from [30]

Hrabar called this a "non-Foster antenna-transmitter" because it exhibited oscillations (oscillations which were intended) and therefore acted like an oscillator and technically classified as a non-Foster circuit because it violated the constraints of Foster's theorem. It is important to note that this circuit was only measured up to 10MHz and does not claim to be a wideband solution, only a negative impedance converter transmitter. Nagarkoti used resonant tunneling diodes to implement a non-Foster circuit [26, 31]. The negative resistance aspect of resonant tunneling diodes can be utilized to invert impedance. Nagarkoti shows operation at higher frequencies which is a distinct advantage over many traditional transistor-based circuits [31]. Figure 2.6 shows the impedance of Nagarkoti's circuit plotted across a wide frequency band including high frequencies. It also shows a varying impedance with bias voltage.

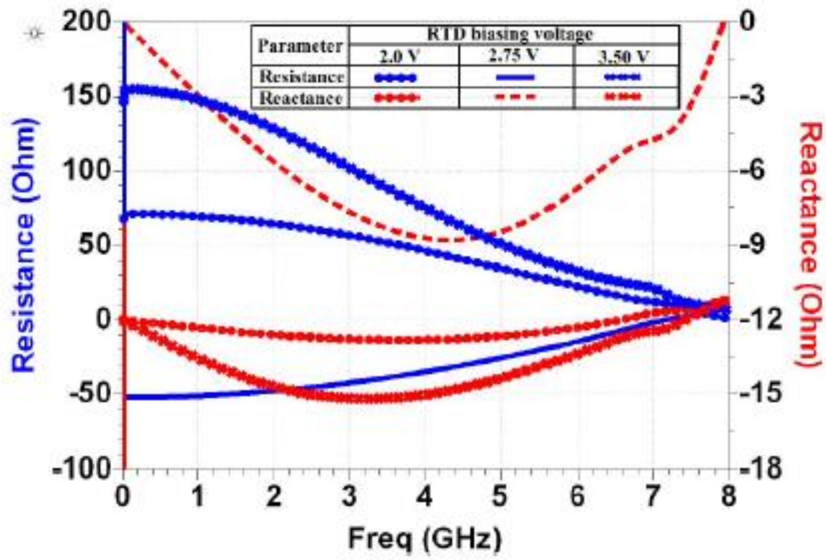


Figure 2.6: Impedance of Nagarkoti's circuit across different bias voltages; showing higher frequency operation. Reprinted from [31]. Copyright © 2016, IEEE

Metamaterials have also been used to implement non-Foster circuits. Hrabar et. al, [32] examines a recently introduced dispersionless metamaterial which demonstrates non-Foster properties in broadband antenna matching. While their conclusions indicate that more work is needed, it demonstrates the fact that Linvill's transistor-based non-Foster circuits are not the only method of achieving an inverted impedance slope across frequency.

Another consideration regarding active components is added noise. For the SNR of a non-Foster matching network to exceed that of an unmatched ESA, the increase in gain must exceed the increase in added noise due to the active components. This issue will be addressed in Section 5.2.

2.3 Comparison Between Different Matching Methods

The methods discussion in this chapter for wideband operation all have unique traits. Some methods are better suited for particular scenarios. There isn't a single method that can mitigate all issues with wideband operation. In fact, all have limitations. Therefore, it is important to understand each method's advantages and limitations to utilize the best option for each specific scenario. A brief description of each method's advantages and limitations is given below, followed by a comparison shown in Table 2.1. The first method discussed was varying the antenna geometry to optimize impedance. This translates into smart antenna design. Antenna designers typically customize the antenna to best suit the size, frequency, and bandwidth constraints. The advantage is the avoidance of a matching circuit, which simplifies the circuitry connected to the antenna. However, it is difficult in practice to design a wideband antenna that is already matched as many scenarios cannot afford an unconstrained antenna for optimization. Situations include antennas with maximum physical size constraints.

The second method is optimizing the circuitry that is designed to function with the antenna. This method could be used for scenarios with strict constraints on antenna construction. Unfortunately, this method utilizes a specifically designed receiver and transmitter circuitry which greatly complicates this design and requires special parts. The trade-off to not placing performance restrictions on the antenna comes in the form of unique parts that are purpose-built.

The third method is a compromise between the first two. A matched LNA network, for example, would allow a somewhat unrestricted antenna, and a somewhat unrestricted circuitry. Rather than designing the entire receiver to meet a custom impedance, the first receiver component could be custom designed to transform the impedance between the antenna and the remaining receiver circuitry and this would allow common parts to be used. Unfortunately, this type of impedance transformer, while still wideband, may not quite have the bandwidth required.

The fourth and last method which discussed earlier, and that which will be the focus of this dissertation, is a non-Foster matching network. A non-Foster network can provide a very large impedance matching bandwidth, especially including the lower frequencies. This involves active circuitry. With active circuits, there is the concern of stability and added noise. At present, non-Foster circuits have been predominantly used in receive-only applications.

Table 2.1: Comparison of different antenna matching methods

Index	Matching Method	Advantages	Limitations
1	Varying antenna geometry	Simplifies circuitry, wideband performance	Very restrictive antenna shapes, sizes, and patterns. Highly constrains the antenna design
2	Optimizing circuit impedance for antenna matching	No restrictions on antenna design	Uses non-common parts or purpose-built parts for each antenna
3	Matched LNA network	Good compromise between #1 and #2	Not as wideband as some other approaches (but can still be considered wideband)
4	Non-Foster matching circuit	Wideband, does not require constraining antenna or circuitry design	Receive only, stability and noise concerns

3 Negative Impedance Inverter Circuits

The previous chapters briefly mentioned the multitude of ways a non-Foster circuit can be implemented. A non-Foster circuit is any circuit that violates Foster's reactance theorem [7], which dictates the positive slope of the reactance curve with increasing frequency. There are many uses for non-Foster circuits, only some of which rely on a specific reactance curve. These other applications, some of which are discussed in Chapter 5, include the tunable filter and the phase shifter. Being non-Foster circuits, they will indeed have a non-monotonically increasing reactance with frequency, but other properties of the active circuit are exploited for each purpose.

Negative impedance inverter circuits (NIC) fall under the umbrella of non-Foster circuits. For the purposes of antenna matching, an inverted reactance is preferred as it creates a wideband match by minimizing the reactance across a wide frequency range. Linvill's transistor-based negative impedance converter circuit is the first known of its kind. There appeared other common implementations which surfaced later, such as the op-amp NICs. Some less-common implementations have also been found in the literature, such as those using metamaterials [19] or resonant tunneling diodes [31]. A discussion on the different types of non-Foster circuit implementations follows.

3.1 Linvill's Transistor-Based Negative Impedance Converter

Linvill's paper introduced the transistor-based non-Foster circuit [12]. Originally intended for applications in the communications industry, it has proven to have uses in other areas, such as antenna matching. Linvill introduced two types of circuits, the open circuit stable (OCS) and short circuit stable (SCS) variants. The original OCS circuit is shown in Figure 3.1. The classic circuits are composed of a pair of cross coupled transistors. A "load impedance" is provided as a reference for the impedance inversion. The transistor bases sample the voltage at the load on opposite sides and therefore the emitter shows a mirrored voltage of the load. However, the current through the load (reference) is largely unchanged, and therefore presents a constant current through the input. The resulting constant current (I) and an opposite voltage (-V) gives the negative impedance (-Z). The SCS variant is of similar construction of the OCS variant except that the load and the input are reversed. This transistor configuration is commonly referred to as a buffer amplifier, or an emitter-follower, because the emitter voltage follows the base voltage. As the name suggests, the buffer amplifier mirrors the voltage at the input and "buffers" the output from the input. Figure 3.2 shows an example of an emitter-follower transistor circuit.

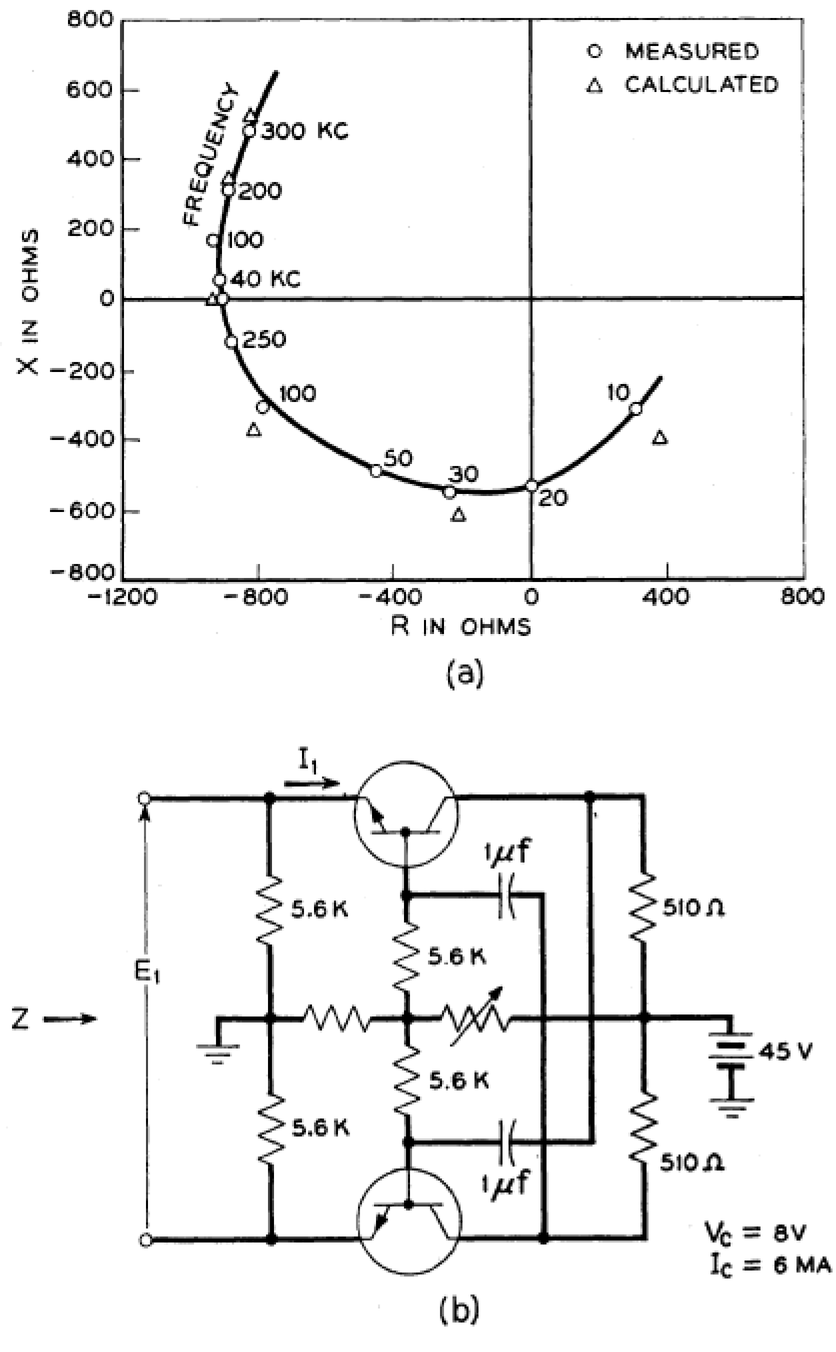


Figure 3.1: a) Impedance of Linvill's circuit and b) Schematic of Linvill's open circuit stable (OCS) negative impedance converter circuit, reprinted from [12]. Copyright © 1953, IEEE

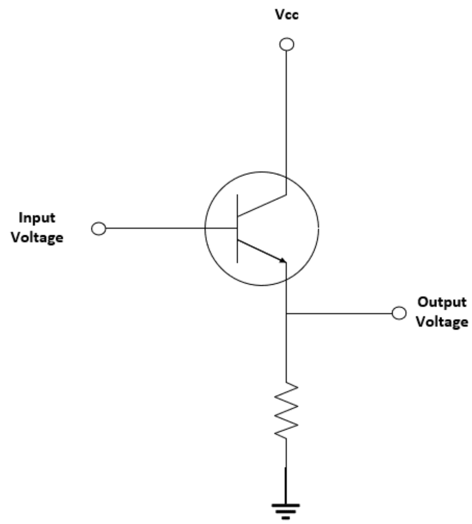


Figure 3.2: Emitter follower, or common collector circuit depiction, reproduced originally from [34]

The application cited in Linvill's original paper is for reducing the loss on telephone transmission lines. The OCS and SCS variants can be used depending on the impedances that are expected on each end of the transmission line. While generally referred to as a negative impedance circuit (NIC) or negative impedance inverter (NII), the antenna matching applications solely rely on the inverted reactance portion of the impedance. When the load of Linvill's circuit is a purely reactive element, such as a capacitor, an inverted reactance can be observed at the input terminals of the circuit. For antenna matching purposes, the "load" side of the circuit is generally a representative reactance element which approximates the reactance of the antenna. Therefore, the antenna is used as a reference for the inverted impedance. In the case of a capacitive load, the input of the circuit is then connected in parallel with the antenna to counteract the reactance, which will also be connected to the receiver circuitry. Non-Foster matching circuits are primarily used for receiving antennas.

A baseline investigation into Linville's circuits has been done using ADS simulations [24]. The first step is to simulate a non-Foster circuit and evaluate the results. An early ADS model, found on the internet [33] can be used. Unfortunately, the transistor used in this model was not a common or easily obtainable transistor, and this warranted performing the simulation with the commonly available transistor 2N222A. Using the early model [33] and the alternative 2N222A transistor, a simple version of the transistor-based non-Foster circuit is simulated in ADS. Figure 3.3 shows the schematic of this model, and the promising results shown in Figure 3.4 are obtained. The two curves in Figure 3.4 compare the reactance of the non-Foster circuit with an ideal reactance. The ideal reactance in this case is a -10pF capacitor, and the corresponding curve of $X_c = -\frac{1}{\omega C}$ is being used as a measure of fidelity. The reactance curve clearly shows a non-monotonically increasing nature with increasing frequency. Therefore, this qualifies as a non-Foster circuit. For the case of the Linvill's standard circuit, there is an excellent agreement between the ideal case and the non-Foster circuit. This can serve as an almost direct replacement for a "negative capacitor".

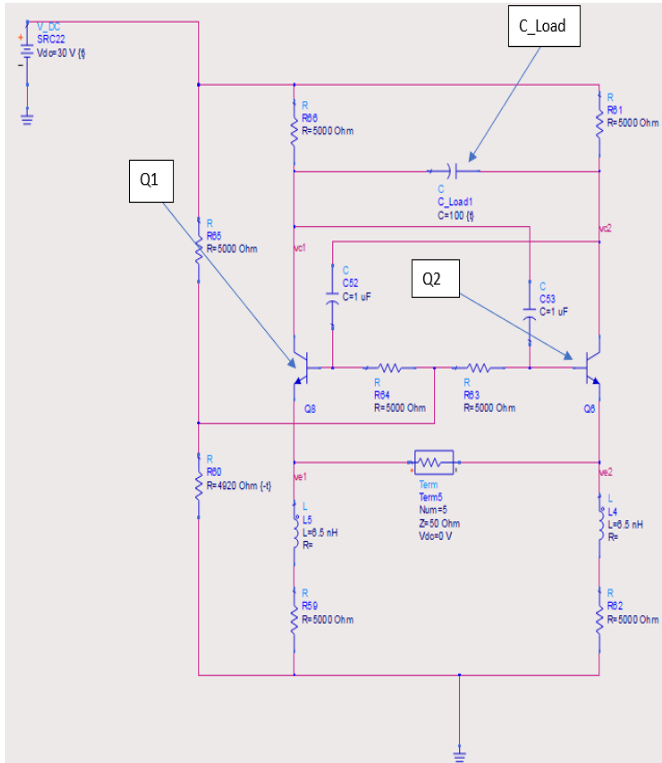


Figure 3.3: ADS schematic of Linvill's transistor-based non-Foster circuit

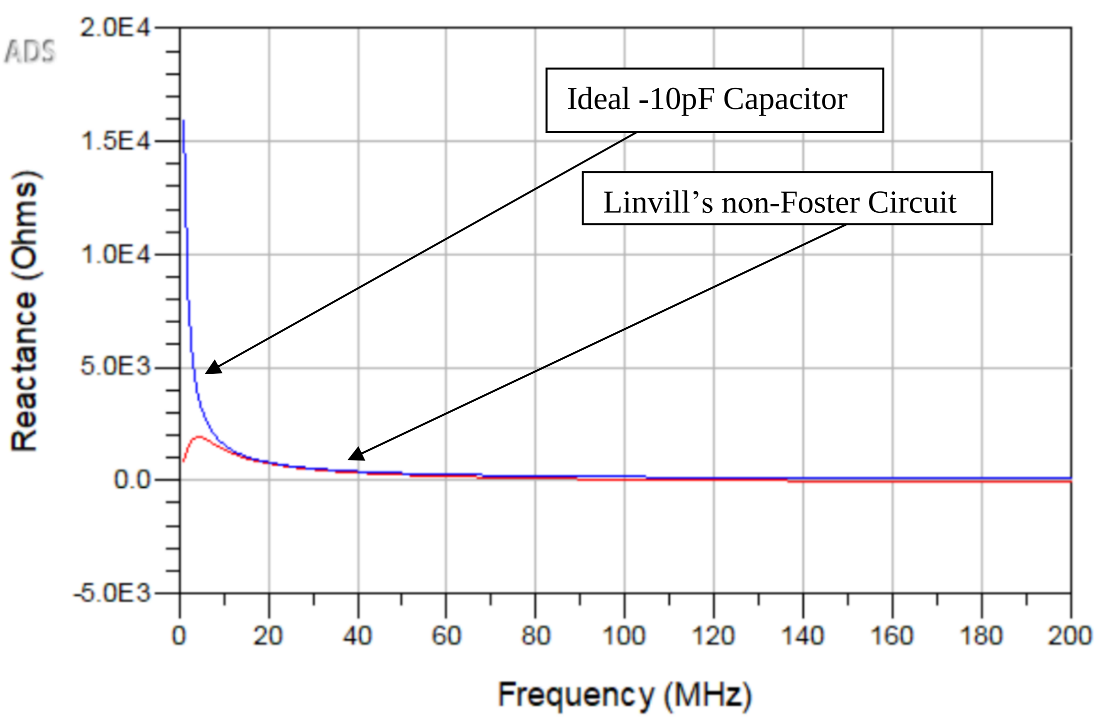


Figure 3.4: Simulated impedance of Linvill's transistor-based non-Foster circuit. Blue curve is an ideal -10pF capacitor

3.2 Operational Amplifier Negative Impedance Converter

Utilizing operational amplifiers (op amps) for non-Foster circuits is common. Because of the widely available parts, this is easily implementable. One of the earliest implementations of non-Foster circuits using op amps is from Perry [35]. The op amp-based non-Foster circuit is based on an inverting op amp configuration. At the two input terminals of an ideal op-amp, identified by the voltages V_1 (non-inverting) and V_2 (inverting), the current into either terminal is zero. Another property of an ideal op amp is the differential input voltage, meaning the voltage difference between the input terminals, is zero. Therefore, when the non-inverting input is grounded and when an impedance is connected between the inverting input and the output, the resulting output is an inverted voltage. This configuration is shown below in Figure 3.5.

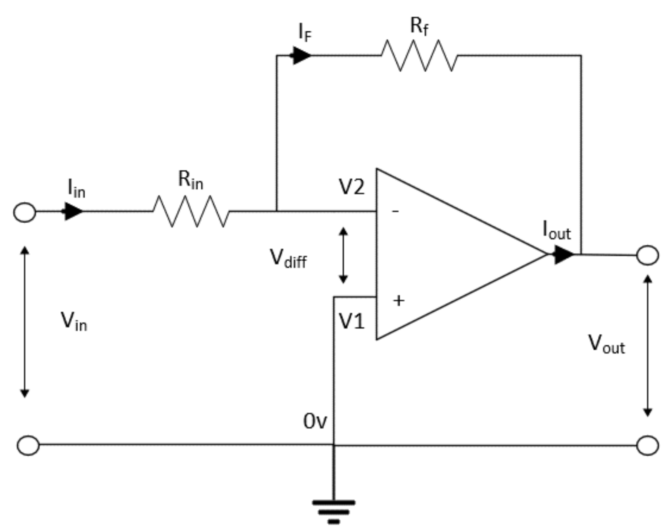


Figure 3.5: Inverting op amp circuit reproduced originally from [36]

Using the properties of an ideal op amp, it is easy to see why the output acts this way.

$$V_1 = V_2 = 0 \tag{3.1}$$

$$I_{V1} = I_{V2} = 0 \tag{3.2}$$

$$\frac{V_{OUT}}{R_F} = -I_F \tag{3.3}$$

$$\frac{V_{IN}}{R_{in}} = I_{in} \tag{3.4}$$

$$I_F = I_{in} \tag{3.5}$$

$$V_{OUT} = -\frac{R_F}{R_{in}} V_{in} \tag{3.6}$$

From Equations (3.1) – (3.6), an inverting op amp output gain can be adjusted by selecting the appropriate input and feedback resistors, R_{in} and R_F respectively. In the case that these resistors are set to the same value, the gain becomes unity with an inverted voltage output.

Fan [19] further clarified the ideal op amp based NIC with an ADS simulation. If the non-inverting input is connected to a load impedance that samples the output, then the resulting impedance into that port appears inverted. Figure 3.6 shows the ADS schematic and Figure 3.7 shows the simulated results, both courtesy of Fan [19].

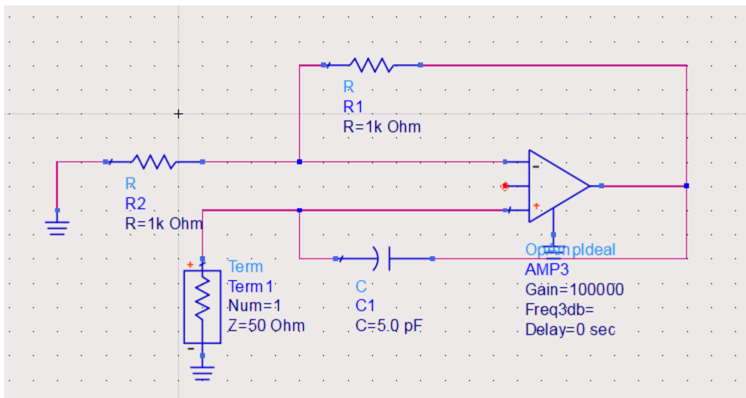


Figure 3.6: ADS schematic of ideal op amp based non-Foster circuit, reproduced originally from [19]

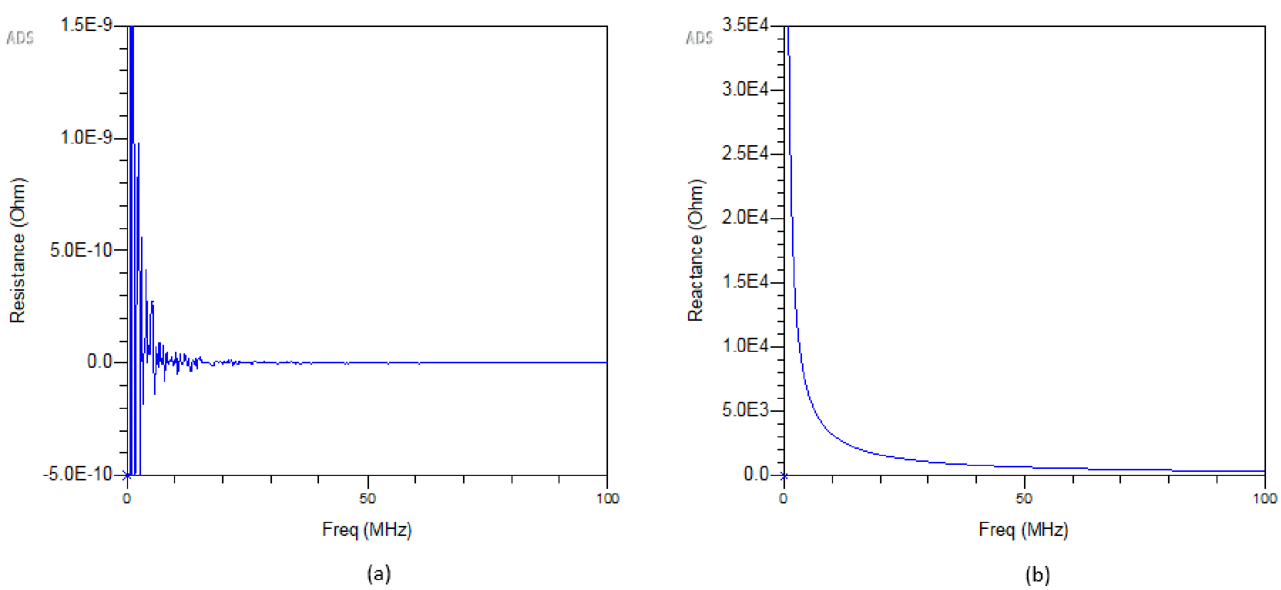


Figure 3.7: ADS simulation results of ideal op amp based non-Foster circuit, reproduced originally from [19]

Like the ideal transistor-based non-Foster circuits described earlier, the practical versions of the op amp based NICs also have their limitations. The defining circuit properties discussed in Equations (3.1) – (3.6) do not hold completely true for non-ideal op amps. Therefore, any variation from these ideal characteristics will degrade performance of the realized non-Foster circuits. Also, potentially even more so than transistors, op amps are severely frequency limited. While there is research into the use of transistors at ever higher frequencies, the use of op amps is generally limited to the lower frequencies and therefore these non-Foster circuits may be so limited. Further, an op amp is comprised of many transistors and therefore almost all the disadvantages of a transistor-based non-Foster circuit also apply to op amp versions. The stability considerations, which will be discussed in Chapter 4, along with the complete list of complications of active circuits are not unique to their transistor-based counterparts. Op amps also suffer from the added noise, are unidirectional, and suffer reduced performance from non-ideal models as do transistor-based non-Foster circuits. One key advantage that the op amp non-Foster circuits have over the transistor-based non-Foster circuits is their simplicity. Since the op amp typically comes in a single integrated package, the circuit is easier to implement. Also, the internal stability of the op amp is already controlled with appropriate terminations, and only the external feedback stability needs to be considered. Nevertheless, care is required to ensure that transistor-based circuits are stable before a load is applied to be inverted.

3.3 Tunable Negative Impedance Converters

Having a tunable negative impedance value provides many advantages for practical applications. Some of these are discussed in Chapter 5. An investigation into some of the different types of tunable circuits is warranted. The literature search discussed in Section 1.5 brings to light previous works that have discussed the tunability of non-Foster networks [18,19,20,21,37]. The tunable transistor-based non-Foster circuits presented previously in the literature fell under three general categories. The first is varying the load impedance, sometimes with the use of a varactor or a switched capacitor bank. The second is by varying the bias voltage of the cross coupled transistors. The last one, from Carson White et al., involves IBM's special process and uses unique parts to create tunability [20]. Each method carries its own set of advantages and disadvantages. The discussion that follows will provide more detail on each method.

3.3.1 Varying the Load Impedance

This method exploits the fundamental operation characteristics of a non-Foster circuit. In any configuration, the non-Foster circuit has a load impedance that is "inverted" to create a desired impedance at the input of the circuit. Non-tunable non-Foster circuits are typically designed ahead of time for the intended application, where a particular load impedance is chosen to best conjugately match the antenna, or for whichever other application is needed. In this method of creating a tunable non-Foster circuit, a variable load impedance replaces the static load impedance. Several devices can be used to achieve this. A switched capacitor bank, or inductor bank, involves an integrated circuit package which houses several different reactance values and a pair of multi-throw switches with which to select the desired impedance. An example is shown in Figure 3.8. The user selects which path through the switch is desired by a control voltage on the devices. When this is connected in place of the load impedance on a non-Foster circuit, tunable operation is achieved. A non-Foster circuit using this topology is shown in Figure 3.9.

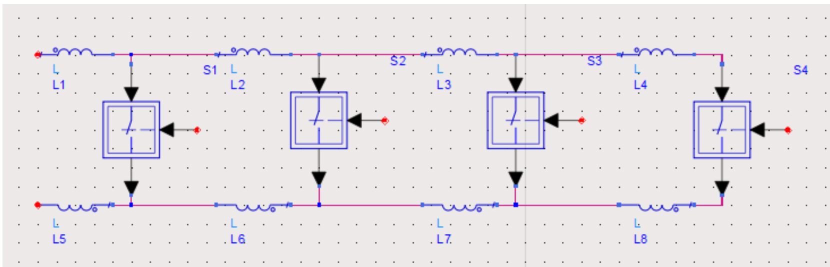


Figure 3.8: Switched inductor schematic reproduced originally from [18]

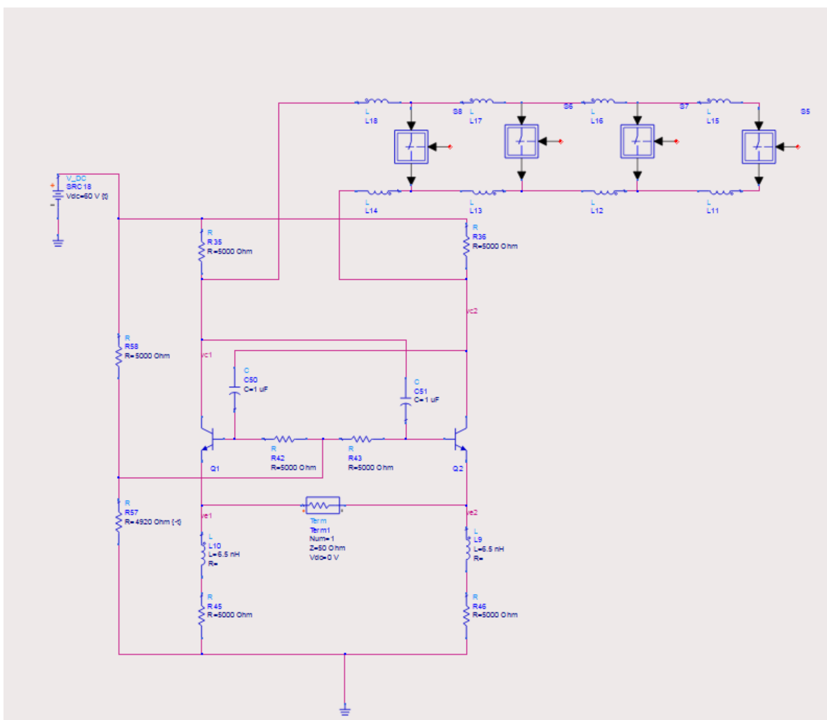


Figure 3.9: Schematic of a tunable non-Foster circuit using a switched inductor with circuit reproduced originally from [18]

There are other devices that can achieve a similar goal, such as varactors. Varactors can change their impedance with a bias voltage. Varactors are typically diodes or semiconductors that change their electron doping concentration based on an applied voltage [38]. Figure 3.10 shows a block diagram depiction of a varactor PN junction. As the doping concentration is changed across the PN junction, the observed capacitance value of the varactor is changed.

Both the switched capacitor bank and varactor provide a means of achieving a tunable non-Foster circuit. For the case of a switched capacitor bank, it will most closely resemble the performance of the non-tunable variant. The other methods of creating tunability, described in following sections, typically carry some performance reduction as a trade-off to achieving tunability.

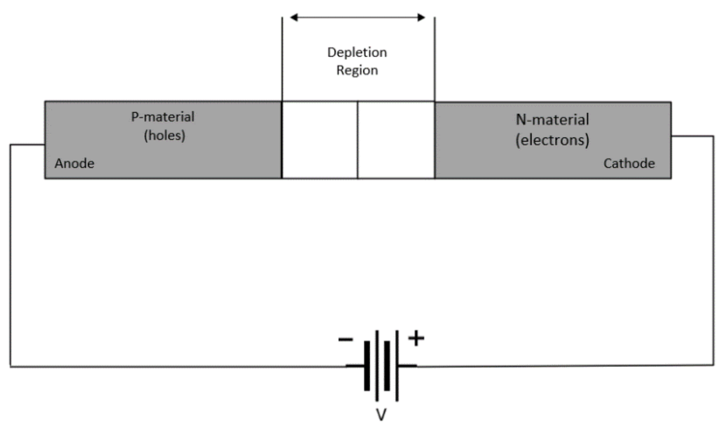


Figure 3.10: Block diagram of a varactor PN junction reproduced originally from [38]

A detailed comparison between the different tuning methods will be provided in Section 3.3.4. However, the main disadvantage of the two methods just discussed are the use of specialized parts. The switched impedance bank is also limited to tuning in discrete steps, while the other methods use a more continuous tuning range. Overall, from a performance perspective, the switched impedance bank stands to match most closely that of the non-tunable non-Foster circuit.

3.3.2 Transistor Bias Voltage Adjustment

In a transistor-based non-Foster circuit implementation, a pair of cross coupled transistors typically is required. The bias voltage of these transistors is largely dependent on the specific transistor chosen for implementation. In general, the transistors need to be operated in a designated region for the non-Foster circuit to exhibit the desirable properties. Transistors generally have three regions, pinch-off, linear/active, and saturation regions. However, even within a given region, as the bias voltage of the transistor varies, the transistor parameters vary, and these differences can be exploited in achieving tunability. Although it is generally not advisable to operate the transistor outside of its intended operating region, there is limited reporting of the tunability achieved by doing so. One example is that of a specialized graphene FET, where the bias voltage was varied via a tuning knob [17]. The graphene FET proves a viable candidate for tuning with varying bias voltage. Figure 3.11 shows a comparison between the traditional load tuning and voltage tuning a graphene type FET non-Foster circuit implementation.

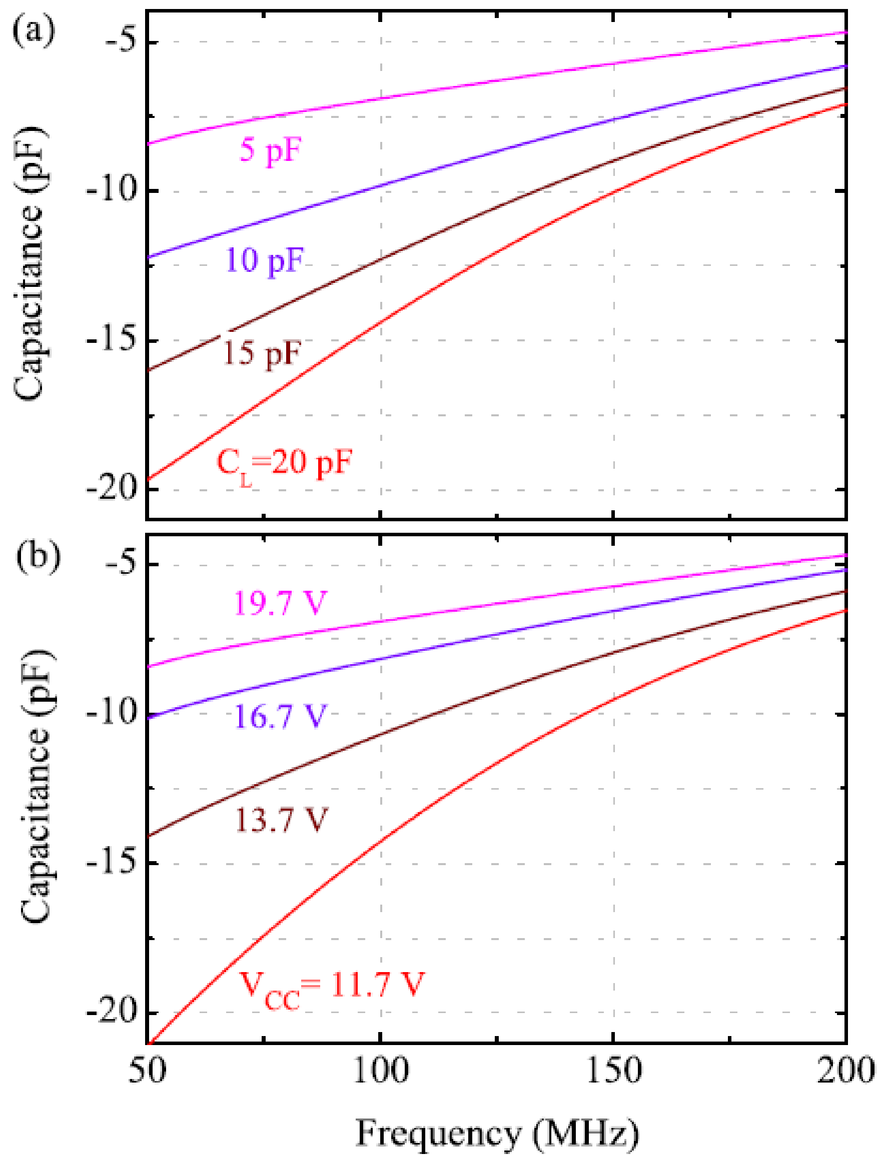


Figure 3.11: Tunable non-Foster network, a) Load tuning and b) voltage tuning plots, reprinted from [17]

Some variability is demonstrated, however, the plots are plotted in capacitance vs frequency, rather than impedance vs frequency. What this fails to demonstrate is the agreement of the curves between an ideal capacitor and the tunable non-Foster circuit. The plots of reactance and capacitance are indeed only related by a factor of $X = \frac{-1}{2\pi f C}$. In the plots shown in Figure 3.11, the trend with increasing frequency is the reactance of the non-Foster circuit further varies from the original capacitance at the lower frequencies. However, it is important to consider that transistor performance tends to decrease with increasing frequency anyway.

For the more general bipolar junction transistor (BJT), further investigation is warranted. ADS simulations have been carried out to demonstrate tunability with varying bias voltage. Building upon Linvill's transistor-based non-Foster circuit model, shown earlier in Figure 3.3, a tunable version is simulated. The tuning parameters are tweaked by adjusting the bias voltage of the main transistors (Q1 and Q2 in Figure 3.3). Changing the bias voltage in the ADS simulation of Linvill's circuit did indeed change the observed impedance. However, like the graphene FET variant shown in [17], the agreement between the curve of an ideal negative capacitance and the non-Foster circuit was poor. This is shown in Figure 3.12.

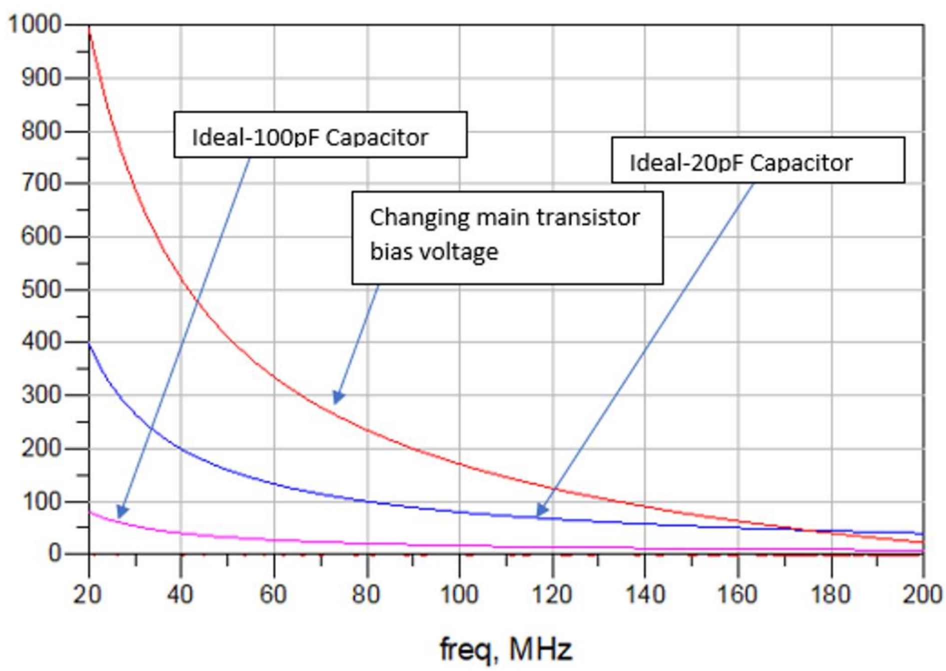


Figure 3.12: Demonstration of reactance with changing bias voltage on Linvill's transistor-based non-Foster circuit

The red curve in Figure 3.12 is the reactance of the transistor-based non-Foster circuit with the bias voltage changed for tunability. In this specific case, the bias voltage, set at 30V, was needed to match most closely the ideal -20pF capacitor. The blue and pink curves are ideal negative capacitance curves shown for comparison. It is evident that the red curve, while demonstrating a monotonically decreasing reactance with frequency, is a poor representation of the ideal negative capacitance. The variation from an ideal capacitance would limit bandwidth.

3.3.3 White et al.'s Method

Carson White et al. presented an innovative solution for a transistor-based, stable, negative inductance inverter with tuning capabilities [20]. The circuit they presented involves the same fundamental cross coupled transistors but adds multiple stages and other components which drive tunability. Figure 3.13 shows the circuit they presented [20]. The circuit was physically implemented on a proprietary IBM process which is not generally available to the public. However, despite the non-standard components being used, the circuit was shown to provide a varying negative inductance value and more importantly operate at frequencies higher than 1MHz.

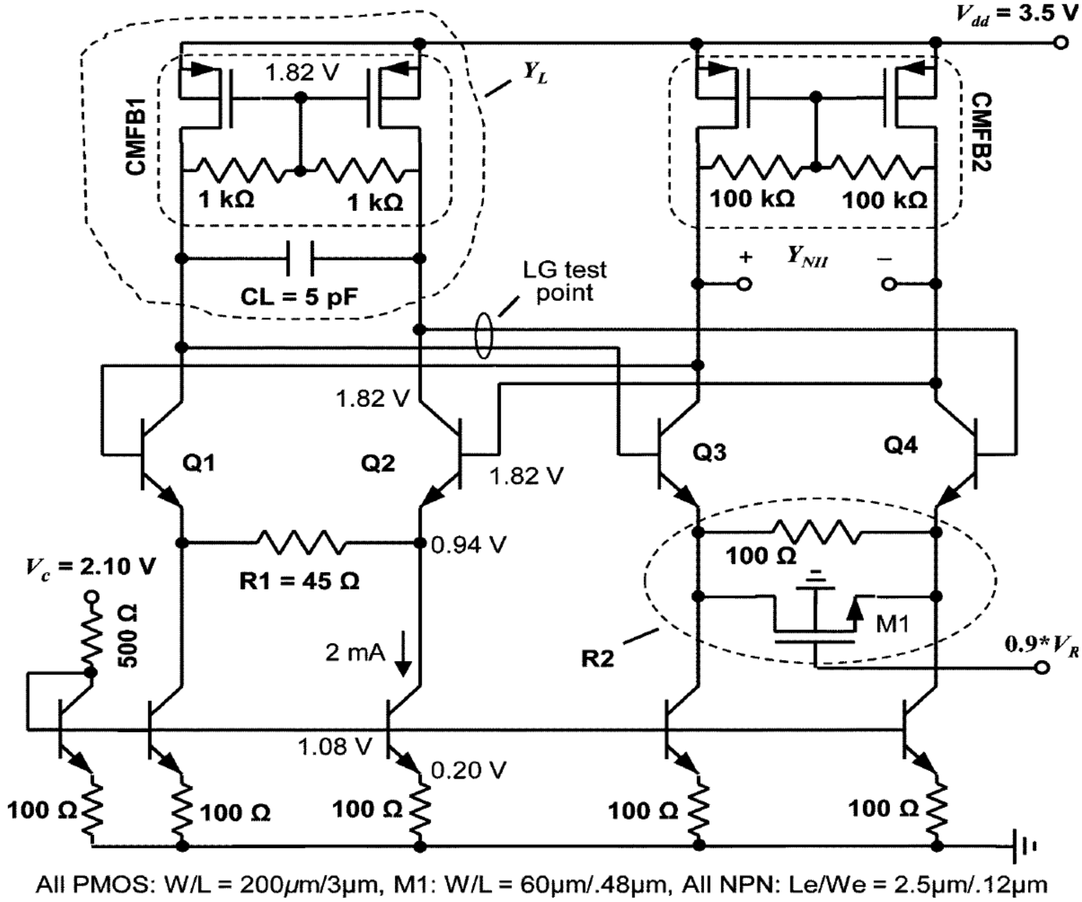


Figure 3.13: White, et. Al. tunable negative inductance circuit reprinted from [20]. Copyright © 2012, IEEE

As such, their measured results displayed excellent performance well above 1MHz and showed a tunable range of -40nH to -64nH. They addressed important stability considerations and used the loop gain method to characterize stability, which was determined via measurement by probing a portion of the circuit. (Later sections address stability of non-Foster circuits in detail.) A key stability finding of [20] was the importance of properly terminating the circuit. They indicated that a negative inductance has a negative time constant, a fictional phenomenon, and therefore cannot be measured in isolation. A “swamping inductor” was used to evaluate the circuit. To demonstrate, consider the equivalent circuit of Figure 3.14. Here, the total inductance measured would be 4nH, indicating that the non-Foster circuit reduced the total inductance by 1nH. This need for additional circuitry is an important aspect of the measurements of an unconditionally stable non-Foster circuit presented in Chapter 4.

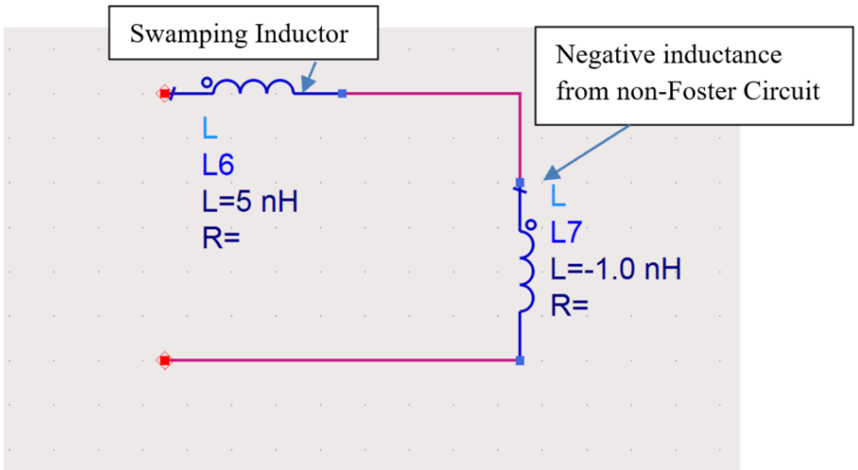


Figure 3.14: Measurement using a swamping inductor

3.3.4 Comparison of Different Tuning Methods

For adjusting the non-Foster output, tunable circuit methods offer versatility and new capabilities that traditional non-tunable circuit methods, such as changing the load impedance, lack. However, performance is assumed to be best with the non-tunable versions. Each of the tuning circuits bring unique advantages but generally trade off performance in exchange for tunability. Choosing the optimum tuning approach requires an understanding of each method. Like many aspects of engineering design, there rarely is a single approach that solves all problems. The following is a comparison between the different tuning methods.

Consider the first option of a switched capacitor or inductor integrated circuit. The electrical performance characteristics are nearly identical to their non-tunable counterparts. With a negligible addition of loss and impedance from the switch, the circuits are electrically identical. The chief downside of this approach is the specialized parts and size required to implement. When more variability is desired, or a better resolution in tuning, an increased number of components are required, including a switch with a higher number of paths.

A related option is using a varactor in place of the load. Rather than having a switched capacitor bank, this is an infinitely tunable capacitor that varies in capacitance with a changing voltage. Although the performance will not quite match that of the switched capacitor, the resolution is much finer with a smaller footprint.

The second approach is varying the bias voltage on the main transistors. This approach has not been reported extensively and understanding the implications of operating the transistors this way is needed before utilizing this method. Some transistors may be more accepting of this approach than others. Although the method simulated above did not perform well on Linvill's classic circuit implementation, the graphene-based FET variant was shown viable [17]. This method is simplest to implement, but performance suffers. The reactance curves of the non-Foster circuit with the transistors operating in unknown regions poorly match the ideal reactance curves.

Third, IBM's proprietary circuit introduced by Carson White et al. [20] provides a good option for non-Foster tunability. There is little information on this circuit, and it requires the use of specialized parts and a proprietary process which is not generally available. However, reference [20] shows good performance throughout a significant tuning range for an inductor.

Finally, a novel approach to a transistor-based tunable non-Foster circuit which will be introduced in Chapter 4 offers a good compromise between these approaches. There are no special parts needed, just an additional tuning transistor of the same type already used. Also, the required size is only slightly more than that of the non-tunable version. The performance will not be as good as the switched capacitor method, but it will be shown to be better than the varying bias voltage method. Table 3.1 shows a summary of the comparison between the different tuning approaches.

Table 3.1: Comparison between different tuning approaches

Circuit Type	Advantages	Limitations
Conventional non-Foster circuit	Simple, proven, effective	Cannot adapt to changing environments. Requires a physical circuit change to vary the matched load
Varactor or switched capacitor bank	Proven effective, closest to the non-tunable version	Uses non-common parts, increased space for capacitor bank
Varying the bias voltage	Simple, easy to build	Varies slightly from the ideal impedance curve; limited tuning range
White et al.'s method [20]	Compact, effective	Uses proprietary circuitry and substrate that is not readily available
Novel solution: adding a transistor	Effective, uses common parts that are already in use	Slightly narrower band than the non-tunable version

3.4 Circuit Implementation and Stability

In addition to ADS simulations, Linvill's original circuit shown in Figure 3.3 is physically realized with a PCB for testing. This process involves drawing up a physical layout of the circuitry and using an LPKF milling machine to cut the copper pattern on an FR4 substrate. The components used are mostly surface mount except for the transistors which are through-hole components. Because of the relatively low frequency of operation, features such as via transitions are realized by simply drilling holes and filling the holes with bus wire and solder. The traces between all the components are maintained at a 50 Ohm characteristic impedance.

The result is a version of Linvill's classical circuit, modified by replacing the transistors with the commonly available 2N222A transistors. Although being terminated in a larger positive capacitance, the measurement of the physical implementation did not align with the ADS simulations. The reason is that the circuit was operating in an unstable region. The resulting circuit had the transistors railing to their bias voltages, meaning the output was fixed at the absolute maximum value, which is a clear indication that the circuit is unstable, and that the transistors were in a positive feedback state. Figure 3.15 and Figure 3.16 show the physical circuit implementation and the impedance measured on an impedance analyzer, respectively. It should be noted that Figure 3.16, which shows the reactance of the unstable circuit, does not necessarily show the instability. The instability was evident by the DC voltages on the transistor pins being outside the operating region of the transistor.

With transistor-based circuits, stability is of particular concern. Any active circuit has the potential to fall into a positive feedback state depending on the external power source applied. In addition to the performance of each type of circuit, stability is a key factor in determining the feasibility. Stability is not guaranteed during all modes of operation. Different impedances, voltages, and temperatures can all have stability implications. For a circuit intended to be connected to an antenna, because the impedance could change with the environment, the stability should be guaranteed for a wide range of loads. Likewise, as different signal levels are incident upon the antenna, there could be the potential for unstable operation. Even temperature fluctuations can impact the transistor level properties such as gain and thus impact stability. An additional stability consideration applies to tunable circuits.

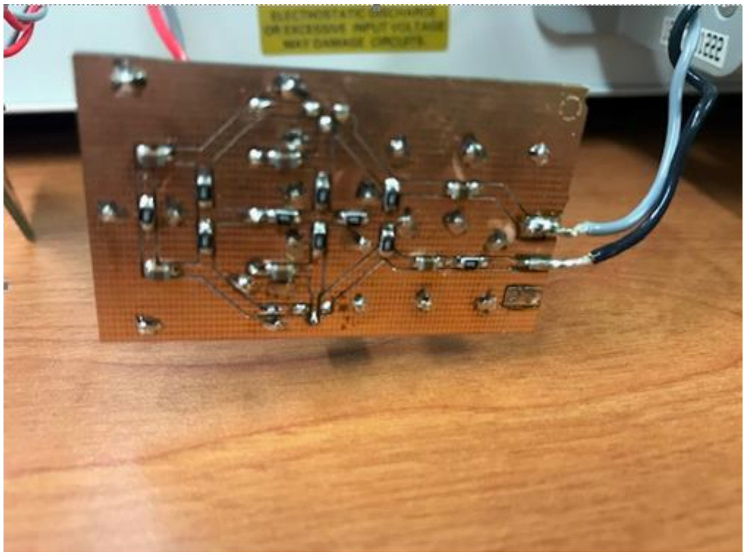


Figure 3.15: Physical implementation of Linvill's original circuit with 2N22A transistors

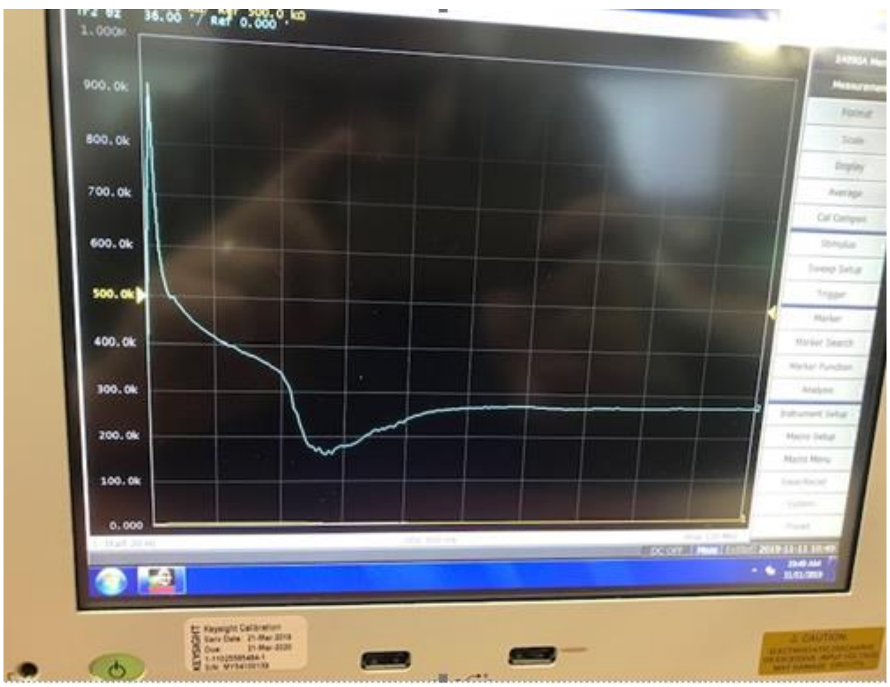


Figure 3.16: Reactance measurement of Linvill's implemented circuit

The tuning parameters can push the circuit in and out of stability regions. For example, one of the tuning methods cited earlier was varying the transistor bias voltage. With a changing bias voltage, the transistor is likely to cross the boundaries of the operating regions. As the transistor moves through these boundaries, each transversal can have a different effect on the resulting impedance and stability. This becomes especially important when the tuning voltage is being changed and it cannot be measured separately when the matching circuit is already coupled to the matched load.

There are numerous papers about stability of non-Foster circuits [39,40]. In Linvill's introduction of transistor-based non-Foster circuits, he included several rules of thumb [12]. These involve the open circuit stable (OCS) configuration and the short circuit stable (SCS) configuration, and both have stability implications. In each of these configurations, the input impedance must be either larger (OCS) or smaller (SCS) than the load impedance. At a minimum, if the input impedance is the same as the load impedance, this will cause the circuit to be operating on the edge of the stability region. As the input impedance increases, for example, in the OCS case, the circuit comes more stable. Of course, the downside to this is that as the input impedance increases, the bandwidth degrades because of impedance mismatch. Some optimal trade-offs usually will be required.

The other caveat to these calculations is that most non-Foster circuits cannot be measured in isolation, because under those conditions they are inherently unstable. The non-Foster circuit must be terminated in an antenna or some specified load that meets certain requirements and allows for an accurate and stable measurement. An example is the swamping method of Figure

3.14 shown in White et al. [20]. However, with simulation, the circuit at least can be analyzed in isolation. Most of the literature addressing non-Foster matching networks shows measurement results of the matching networks when combined with an antenna. The procedure for measuring the effective impedance of this circuit is to first measure the impedance of a passive “input” device, then measure the overall impedance of the input device connected to the non-Foster matching network. The overall ‘decrease’ in the imaginary part of the impedance will improve the Q value of the combination of the matching circuit and antenna.

The stability analysis of a non-Foster circuit is quite involved and is required to guarantee unconditional stability. Many methods exist for determining stability, some with unrealistic assumptions. Stability analysis will be addressed in depth in Section 4.2, where a general criteria for unconditional stability is discussed as well as a specific stability analysis for a proposed tunable non-Foster circuit.

4 A New Method for Tuning Transistor-Based non-Foster Circuits

Antenna design has become increasingly constrained with the advancement of technology and evolving applications. Systems are demanding wider bandwidths and more compact size. Using a traditional non-Foster matching network with an electrically small antenna often leaves more gain and overall system performance to be desired. A tunable non-Foster network unlocks an additional degree of freedom for meeting stringent system performance. The discussion in Section 3.3 detailed a comparison between various non-Foster tuning methods found in the literature. Each tuning method comes with its associated advantages and limitations. Specific applications could benefit from one method over another.

The proposed method offers several unique advantages of its own. Systems which operate over a large bandwidth can be divided into two categories: those which operate over many discrete narrow-band steps but encompass a large bandwidth, and those which require a large instantaneous bandwidth. For example, a communications system which has many narrow-band channels would only require the use of a narrow bandwidth at any given time, however, the minimum and maximum frequencies of the channels may span a large bandwidth and therefore be considered wideband. Methods utilizing discrete tuning transitions may not be suitable for these applications. Furthermore, the proposed method requires no special parts and can be considered an add-on to an existing non-Foster circuit, rather than being a standalone circuit.

This chapter will detail a novel method of adding tunability to a transistor-based non-Foster circuit. The inner workings, using a classic Linvill non-Foster circuit, will be explained in detail to show circuit operation. Simulation data of the added transistor will show the tunability of the non-Foster circuit and plots are given that compare the provided reactance with that of an ideal inverted passive element. With all active circuits, stability is of particular concern. An iteration of the proposed circuit which combines the Harris & Myers [13] circuit with the added transistor is shown to be unconditionally stable in Section 4.2. Simulations of the tunable non-Foster circuit are shown in Section 4.3 which show a changing reactance with varying the tuning transistor base-to-ground voltage. The resulting reactance is compared to that of an ideal inverted capacitor reactance. Physical implementation of the tunable non-Foster circuit on a printed circuit board (PCB) is shown in Section 4.4 where measurements of the circuit are taken with an impedance analyzer. Measurements are also given for the combination of the tunable non-Foster network and a wideband antenna in Section 5.2. Gain improvements are shown at the lower frequencies where electrically small antennas suffer from poor gain.

4.1 Method for Tuning Transistor-Based non-Foster Circuits

The proposed approach to tuning a non-Foster circuit maintains the same elements as the original non-Foster circuit. Only one additional tuning transistor is added for the purpose tuning. The added tuning transistor is connected to the load impedance, which is the reference load to be inverted for the traditional non-Foster circuit. Changing the base-to-ground voltage of the tuning transistor results in a tunable output reactance.

As discussed in Section 3.1, the original transistor-based non-Foster circuit is based on the principle of two emitter follower stages, which reference the voltage at their bases to their emitters, with the emitters being the output. The impedance inversion occurs because the base voltage of the transistors is crossed over the load, and therefore the output voltage at the emitters is the negative of the voltage across the load of the non-Foster circuit. Since the current in the load is mostly unchanged, this results in the same current (I), but an inverted voltage (V), which creates a negative impedance (Z).

A more detailed description of voltages and currents in the critical paths that effect circuit operation follows. The voltage provided to the non-Foster circuit, in this case is a DC voltage which provides several functions. First, it provides the bias voltage for the main transistors. A pair of resistors forming a voltage divider determine the base-to-emitter voltage (V_{BE}) on the cross coupled transistor pair. Additionally, on the base of the transistors Q1 and Q2 is also where the capacitor load voltage is sampled from each side of the capacitor. Note that the sampled load voltage is an AC voltage which is superimposed on the DC base voltage. To ensure that only the AC voltage is sampled, a pair of DC blocking capacitors are put in series with this sampling wire. Figure 4.1 depicts Linvill's non-tunable non-Foster circuit with these voltages annotated.

The novel method to render this circuit tunable is adding an additional tuning transistor. Figure 4.2 shows the modified version of Linvill's non-Foster circuit with the tuning transistor, Q3, highlighted. The main transistors, Q1 and Q2, operate in the saturation region, as determined by the base-to-emitter voltage (V_{BE}) which is greater than the transistor threshold.

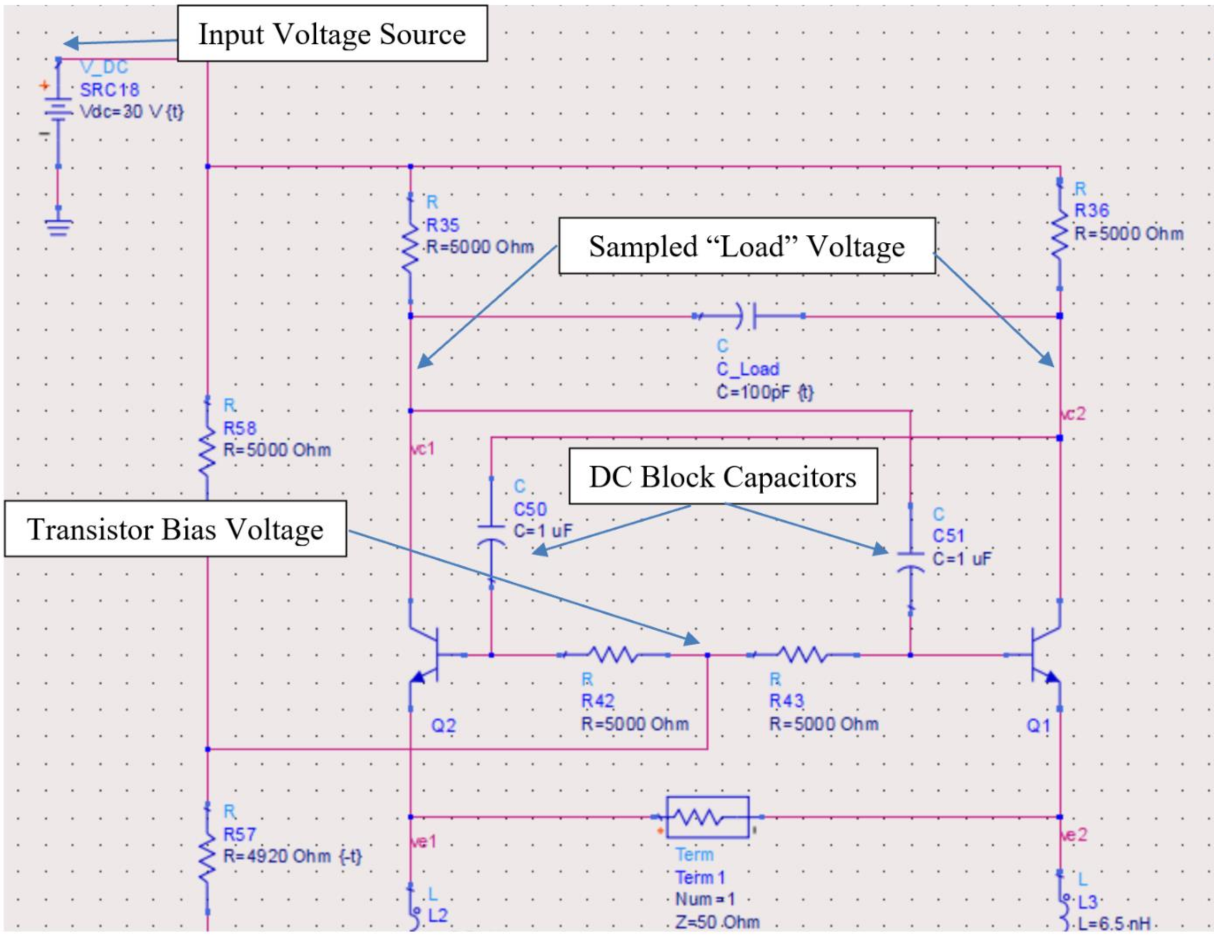


Figure 4.1: Detailed description of Linvill's transistor-based non-Foster circuit voltages

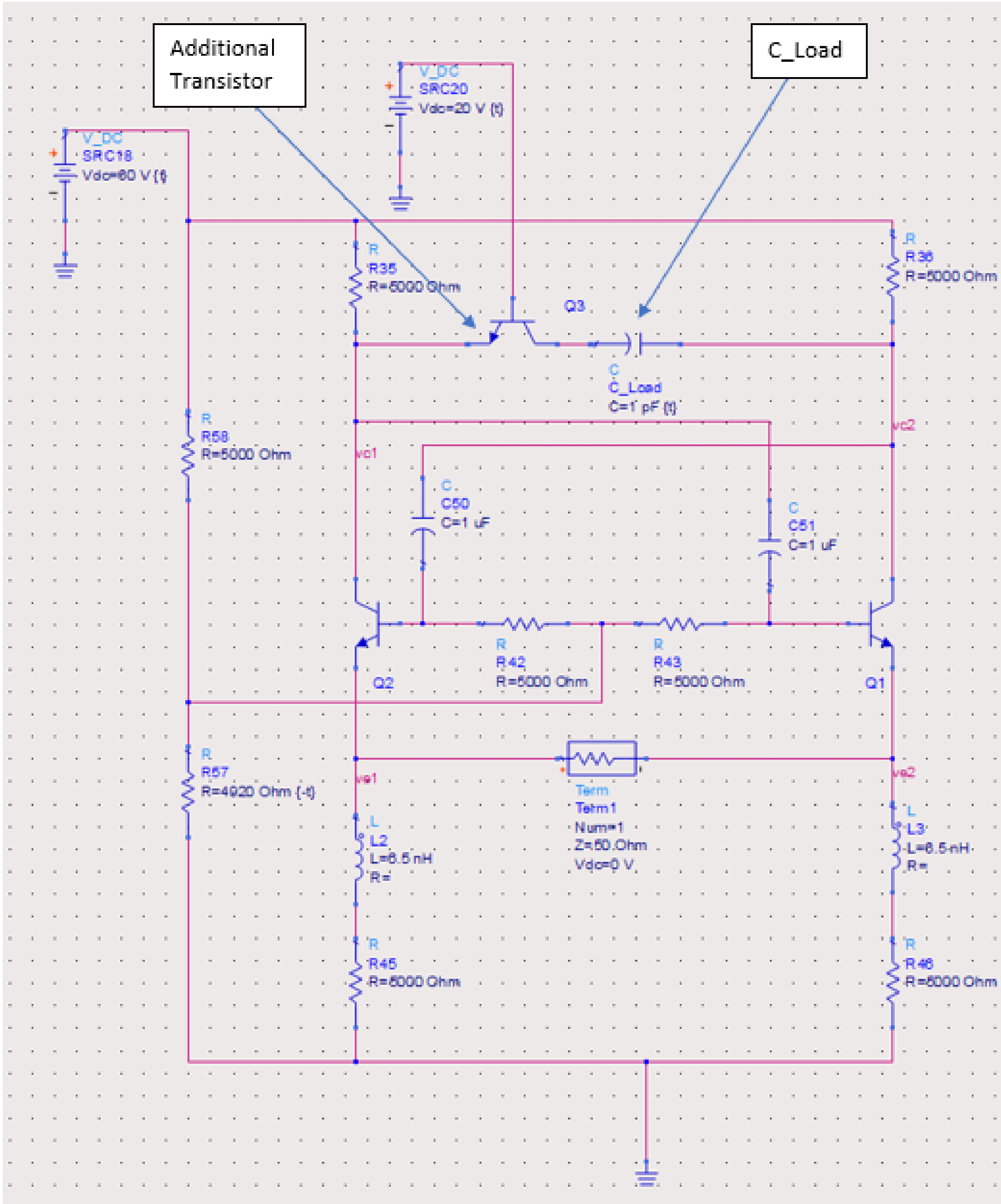


Figure 4.2: "Added Transistor" tunable non-Foster circuit approach

The fundamental concept behind the proposed method of adding tunability is to alter the voltage sampled by these transistors without affecting the current through the load capacitor. This is a key difference from a related method discussed earlier in Section 3.3.2. That method changed the bias voltage of Q1 and Q2 which could push the transistor out of the saturation region and into other undesirable regions. Not only would the voltage sampled by the transistor change, but the current would also change. However, if the voltage and current change in a non-proportional manner, a changing impedance will still be observed.

Unfortunately, the transistor exiting the saturation region renders this no longer an emitter-follower and deviates from the fundamental operation of Linvill's circuit. The proposed method, which adds an additional transistor to change the voltage ensures that the transistor always remains in the saturation region. The altered voltage provided by the added transistor does not allow the base-to-emitter voltage (V_{BE}) of the main transistors to exit the saturation region.

Simulations for this circuit were performed using ADS, showing good results. With this specific configuration, the circuit is able to achieve tunable capacitance values with a range of -20pF to -100pF. Figure 4.3 and Figure 4.4 show the resulting impedance curve of this circuit in green. The reactance closely follows the curve of the ideal negative capacitance curve. Adjusting the tuning voltage can make the circuit more closely match the ideal curve at a desired frequency, but getting a perfect match across all frequencies may not be feasible for any one tuning voltage.

As a comparison with other tuning methods, Figure 4.5 shows an overlayed plot of the proposed tunable non-Foster circuit and the previously mentioned method of varying the transistor bias voltage. While both result in a changing reactance, the proposed circuit matches the ideal reactance curves much more closely. This is especially important for wideband operation.

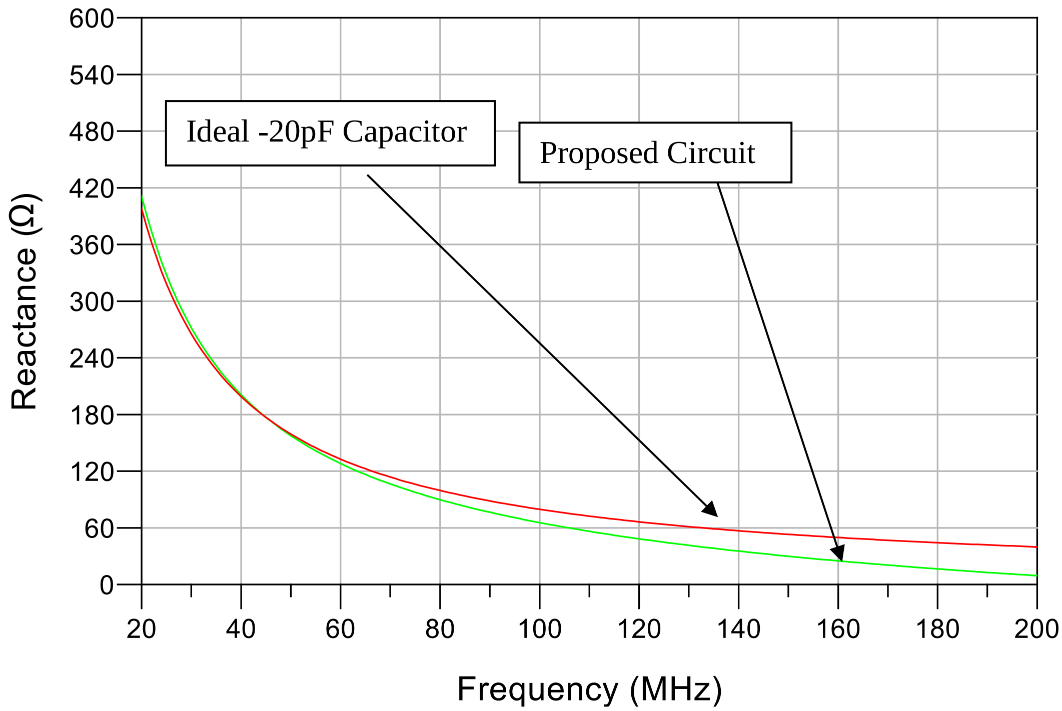


Figure 4.3: Tunable non-Foster circuit configured at -20pF

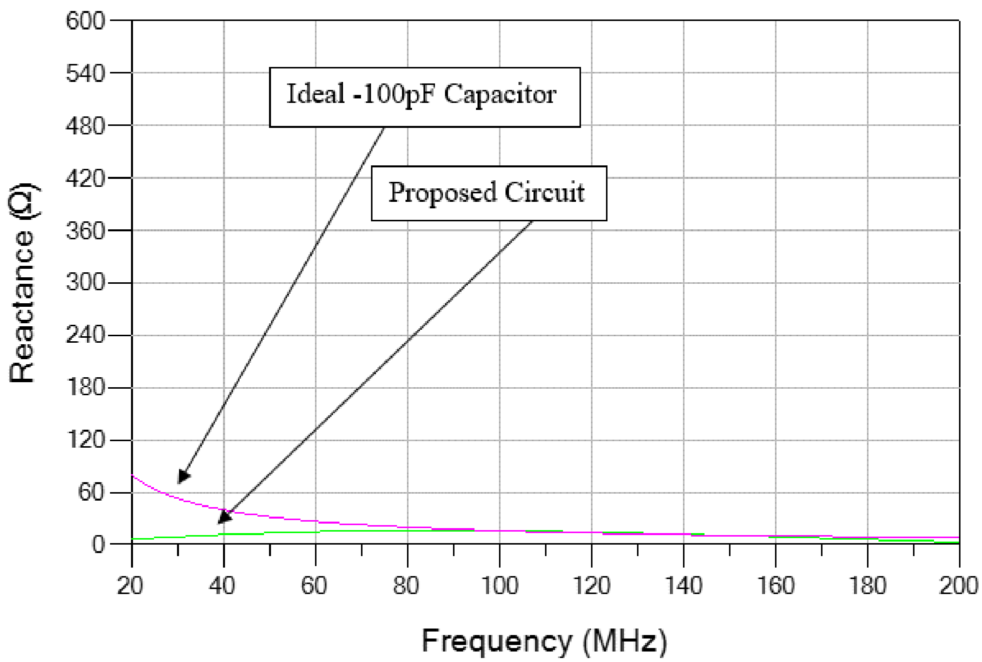


Figure 4.4: Tunable non-Foster circuit configured at -100pF

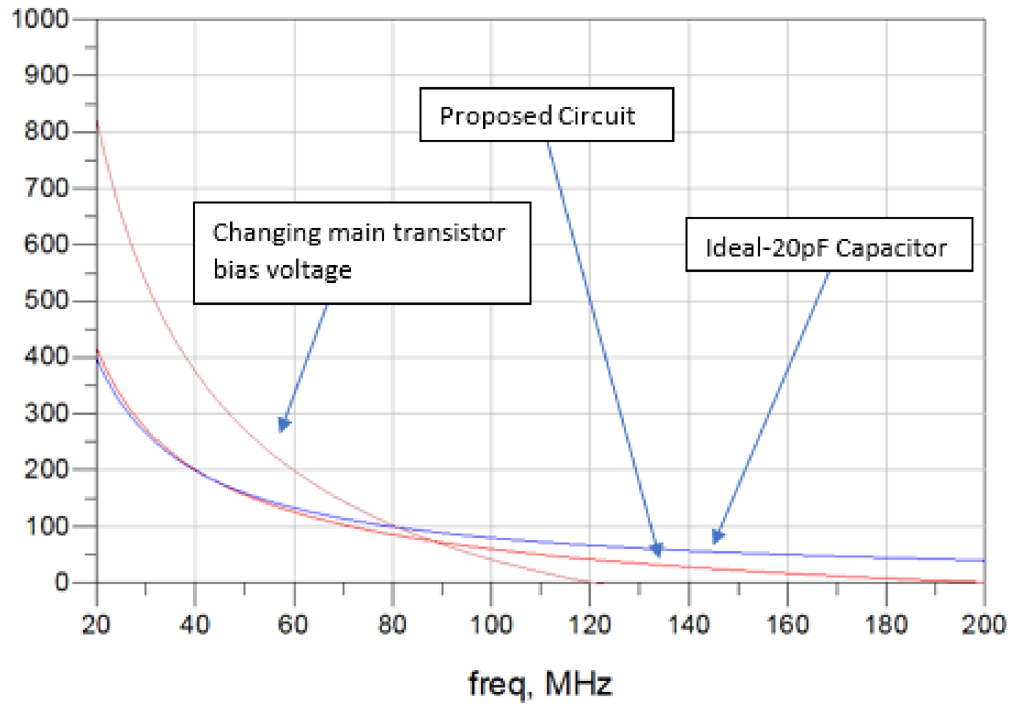


Figure 4.5: Comparison between proposed tunable non-Foster circuit and changing transistor bias voltage

4.2 Stability Analysis and Small Signal Circuit Model of the Tunable Non-Foster Circuit

4.2.1 Harris & Myers' Circuit with Improved Stability

As discussed in Chapter 3, Linvill presented two non-Foster circuits for inverting impedances, one of which was deemed open circuit stable (OCS), and the other was short circuit stable (SCS). In 1968, Harris and Meyers [13] developed a negative impedance converter used for improving the performance of a miniaturized antenna. They compared the results of a negative impedance converter (NIC) actively matched short monopole with a top hat to that of a conventional 16-foot whip antenna and showed an improvement in receive gain over the long (16 foot), non-matched case. Harris and Myers introduced a circuit which started with Linvill's baseline transistor-based non-Foster circuit and added additional transistor stages to improve stability. For Harris and Myer's application of antenna matching, the original circuit stability was not sufficient and additional work was needed to improve stability.

The tunable circuit shown in Figure 4.6 was derived from Harris and Myers' original creation [13]. While this circuit will be discussed in more detail later, the basic idea behind their creation was that they forced the input and output impedances to desired values by using transistors stages which had high input impedance and low output impedance. For an open circuit stable (OCS) NIC, where the input impedance must be greater than the output impedance, putting a stage at the beginning and the end of the non-Foster circuit was an easy (and ingenious) way to meet this requirement.

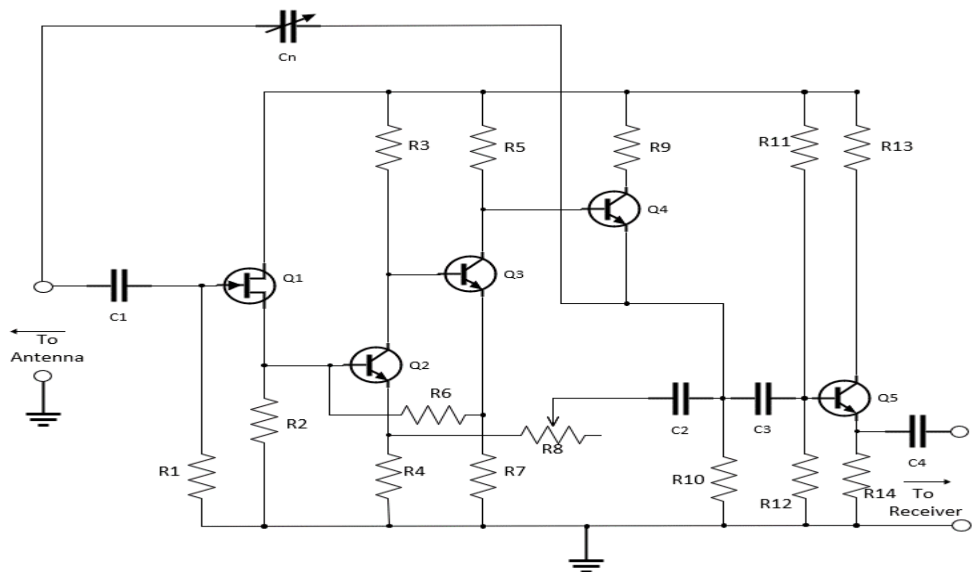


Figure 4.6: Harris and Myer's original circuit with improved stability reproduced from [13]

4.2.2 Stability Analysis of the Tunable Non-Foster Circuit

Modeling stability becomes a critical component to designing a usable non-Foster matching network. A literature review shows several methods for determining stability of a network. Critical assumptions need to be observed when choosing a method for stability analysis. Stability issues originated with the design and measurement of cascaded amplifiers and feedback networks. As individual components are combined into larger circuits, interactions between different circuit sections induce instabilities during operation [41,42]. These instabilities could be mapped to right-half-plane (RHP) poles of the system transfer function across frequency. For example, amplifiers which are designed and tuned to be stable independently can become potentially unstable when combined. Feedback constraints were examined to guarantee stability.

Early criteria for determining stability, such as Rollet's K-factor [41,42,43], were used to numerically enumerate stability. The K-factor can be written as

$$K = \frac{2\,Re\,(\gamma_{11})Re\,(\gamma_{22}) - Re\,(\gamma_{12}\gamma_{21})}{|\gamma_{12}\gamma_{21}|} \tag{4.1}$$

where γ is impedance, admittance, or other parameters [44]. A K factor greater than one indicates a stable network, and less than one indicates a potentially unstable network. Platzker and Struble make note of a key restriction of Rollet's K factor that the unloaded network transfer function can not contain any RHP poles [41,43]. Essentially, this means that the system must be inherently or "naturally" stable and the K-factor can determine if the application of feedback under certain conditions can potentially cause instability. However, if the system is not guaranteed stable under unloaded conditions without feedback, the K-factor cannot be used as a rigorous method for determining unconditional stability.

Many microwave components and circuits are characterized by their scattering parameter (S-parameter) matrix. Kurokawa, Bodway, and Hauri devised an alternate stability criterion based on S-parameters for a two-port network [45]. In their work,

$$K = \frac{1- |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2\,|S_{12}S_{21}|} \qquad (4.2)$$

where S_{11}, S_{12}, S_{21}, and S_{22} are components of the S-parameter matrix and $\Delta = S_{11}S_{22} - S_{12}S_{21}$. There are implications of using S-parameters to determine stability. Some assumptions must accompany the use of an S-parameter matrix. The first is that the system is terminated in the S-parameter normalizing impedance, usually 50 ohms. A more subtle assumption rises from the fact that a given S-parameter, whether derived from measurement or simulation, indicates that a network is stable without external feedback. Otherwise, the scattering parameters would not properly characterize the network if it had an inherit instability. While this may be relevant and accurate for a plethora of microwave circuits, neither of these assumptions apply to non-Foster matching networks. Therefore, this is another stability criteria that is widely used in microwave circuits but cannot be used as a rigorous determination for stability in non-Foster matching networks.

Platzker and Struble introduced the normalized determinant function (NDF) method as a rigorous method for determining unconditional stability [43,46]. The NDF method is based on the system transfer function equation. For linear systems, we can represent their input and output dependencies by their impedance or admittance matrices:

$$Y(s)V(s) = I(s) \tag{4.3}$$

and

$$s = \sigma + j\omega \tag{4.4}$$

Output I(s) can be represented by Y(s)V(s) where Y is the admittance matrix, V is the voltage, and I is the current. The characteristic equation of the system, or sometimes referred to as the system transfer function, is best described by the admittance matrix Y(s). The system response, represented by I(s), can be determined as a function of the input V(s) and the characteristic equation of the system Y(s).

The NDF is defined as

$$NDF = \frac{|Y|}{|Y'|} \tag{4.5}$$

where $|Y|$ is the determinant of the system admittance matrix, and $|Y'|$ is the determinant of the system admittance matrix of a known stable variant. The stable variant by definition needs to completely stable and contain neither zeros or poles in the RHP. The NDF method therefore does not rely on the assumption of an inherently stable system (i.e. no RHP poles) for Y. Conversely, the NDF method examines the system transfer function, in the form of the admittance matrix, for RHP poles that could potentially cause instabilities. The poles of the system transfer function become the zeros of the network determinant [41,43,46]. A determinant is zero when the matrix is not of full rank. A matrix which is not full rank indicates that one or more columns are linearly dependent. In the case of a determinant which is comprised of a system admittance matrix, this means that the (voltage) inputs could be dependent on one another. Recall, from (4.3), the equation describing a network with the system admittance matrix. If we consider the case where [Y] is not full rank, its determinant is zero and there will be some vector of voltage inputs [V], that is non-trivial (meaning non-zero voltage on all inputs), that results in the net current vector being zero. This indicates a scenario with positive power input to the system resulting in no power out. This is a discontinuity that can be recognized as a pole of the system transfer function.

In the time domain, the system transfer function can then be represented as a function of complex conjugate root pairs:

$$\sum_{k=1}^{p} a_k t^{(m_k-1)} e^{(\sigma_k \pm j\omega_k)t} \tag{4.6}$$

where t is time, m_k is the multiplicity factor, and $\sigma_k \pm j\omega_k$ is the kth root of the network, or the zero of the determinant [41]. By examination, it is apparent that the response will decay as t $\rightarrow\infty$ if $\sigma < 0$. For any RHP poles ($\sigma > 0$) this will result in an exponential growth of the function as t $\rightarrow\infty$. This conclusion applies to any system with a transfer function that can be expressed as a function of complex exponentials (which by extension of a Fourier series, any function can be expressed this way). Therefore, the lack of RHP poles is a rigorous determination of system stability.

To determine if a transfer function has any RHP poles, we use the Index Principle, or sometimes referred to as the Principle of Argument Theorem [47]. For a closed curve C in the s-plane, and for a function f(s) which has no singular points or zeros on C:

$$\oint_C \frac{f'(s)}{f(s)} ds = j2\pi\left(\sum_{p=1}^{N} n_p - \sum_{q=1}^{M} m_q\right) \tag{4.7}$$

where N is the number of zeros and M is the number of poles, and n_p and m_q are the individual zeroes and poles, respectively [47]. Any function that can be represented in this way is subject to this theorem. This allows easy identification of poles and zeros by observing the encirclements of the phase about the origin. Encirclements are when the angle (or phase) of the function, plotted on a polar plot, makes a complete circle about the origin. With a carefully chosen contour, and with the assumption that this function decays as $\omega \rightarrow \infty$, this can more specifically limit this to RHP poles and zeros. There is no guarantee that a system will decay with infinite frequency ($\omega \rightarrow \infty$,), in general. However, the NDF method forces this assumption by normalizing with a known stable alternative circuit that has the same number of determinant poles as that of the circuit under consideration. There will be more discussion on this later. By choosing to integrate over the entire RHP, the number of encirclements will reflect only the RHP poles and zeros. Platzker and Struble [41] also make a note of the travel direction and its relationship with the phase encirclements as follows:

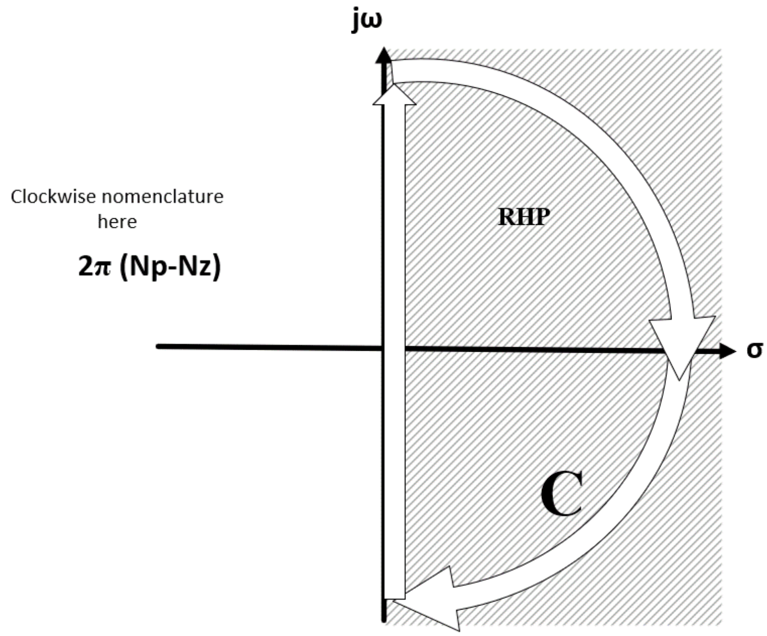


Figure 4.7: Contour of integration path (RHP) illustration, reproduced originally from [41]

Note, from equation (4.7), standard mathematical textbooks such as LePage [47] write the net encirclements as zeros minus poles. Figure 4.7 indicates the opposite, where Np and Nz are number of poles and zeroes, respectively. However, if the travel direction is reversed, then the total number of encirclements becomes poles minus zeros. This becomes important later because through utilizing the NDF, by design, we will ensure there are no zeros and the net encirclements will only be the RHP poles.

If there are no zeros in the function whose angle is plotted on the polar plot, then the origin encirclements will only describe the poles. Recall that the zeros of the determinant correspond to poles of the transfer function. Therefore, satisfying $N_z = 0$, which indicates that real functions decay as $\omega \rightarrow \infty$, is not necessarily guaranteed for potentially unstable circuits. Like the K-factor method, having this assumption would severely limit the applicability of this stability criterion. The NDF forces the overall argument to behave this way and therefore does not require the function in question (numerator) to decay with infinite frequency. This is done by normalizing the determinant with a known stable variant. A known stable variant must contain the same number of nodes, or rows in the network admittance matrix as the function in question. However, all active or questionable sources must be disabled to ensure that the known stable variant is "always stable." An "always stable" variant is such that this network will have no RHP zeros or poles [48]. Creating this stable variant for our purposes can be done by simply deactivating the sources (transistors) of the network and rendering it passive [48]. With only passive and therefore lossy components, the circuit cannot grow exponentially in any capacity.

The determinant of the admittance matrix of the full active network is then normalized by the determinant of the admittance matrix of the stable variant. This ratio, sometimes referred to as a "return ratio" similar to the bode plots [46] becomes the NDF. The resulting phase of this function, when plotted on a polar plot (or Nyquist plot), is a quick visual check to determine network stability. The number of times the NDF encircles the origin will determine the number of RHP poles in the system transfer function. For theoretical purposes, this needs to be done across infinite frequency (to encompass the entire RHP). However, for practical purposes, negative frequencies and frequencies that are much higher ($\omega \gg 10f$) than the frequency of operation (f) are not included. With modern simulation software, the Nyquist plot can provide additional information as to what frequency contributes to the instability for further analysis and troubleshooting.

From Figure 4.7, the phase encirclements are dependent on the number of poles and zeros of the function. By definition, the NDF has no poles since the normalizing function is totally stable. This just leaves zeros of the determinant which correspond to RHP poles of the network admittance matrix. Therefore, simply observing the phase of the NDF when plotted on a Nyquist plot, the number of encirclements about the origin is directly related to the number of RHP poles of the admittance matrix. This is a rigorous stability criterion which applies to all functions that can be represented in the equations mentioned earlier.

4.3 Modeling and Simulation Results

The modeling and simulation for this tunable non-Foster circuit is all done in Advanced Design System (ADS) [24]. ADS is a nice tool that manages the balance between a 2D modeling software and a 3D modeling software. The models aren't as primitive as ideal Spice circuit models in that they include S-parameters and matching considerations for cascading components. For example, many of the components used in ADS are from models available online from the component manufacturers. The software does not compute a full 3D model of the circuit board and all its components. Nevertheless, ADS is a relatively quick simulation tool that can provide design guidelines and give an approximate result.

The ADS modeling started with Harris and Myer's original circuit as derived from their publication [13]. First, it was modeled and evaluated for basic functionality of a negative impedance inverter circuit. Figure 4.8 shows the modeled circuit in ADS.

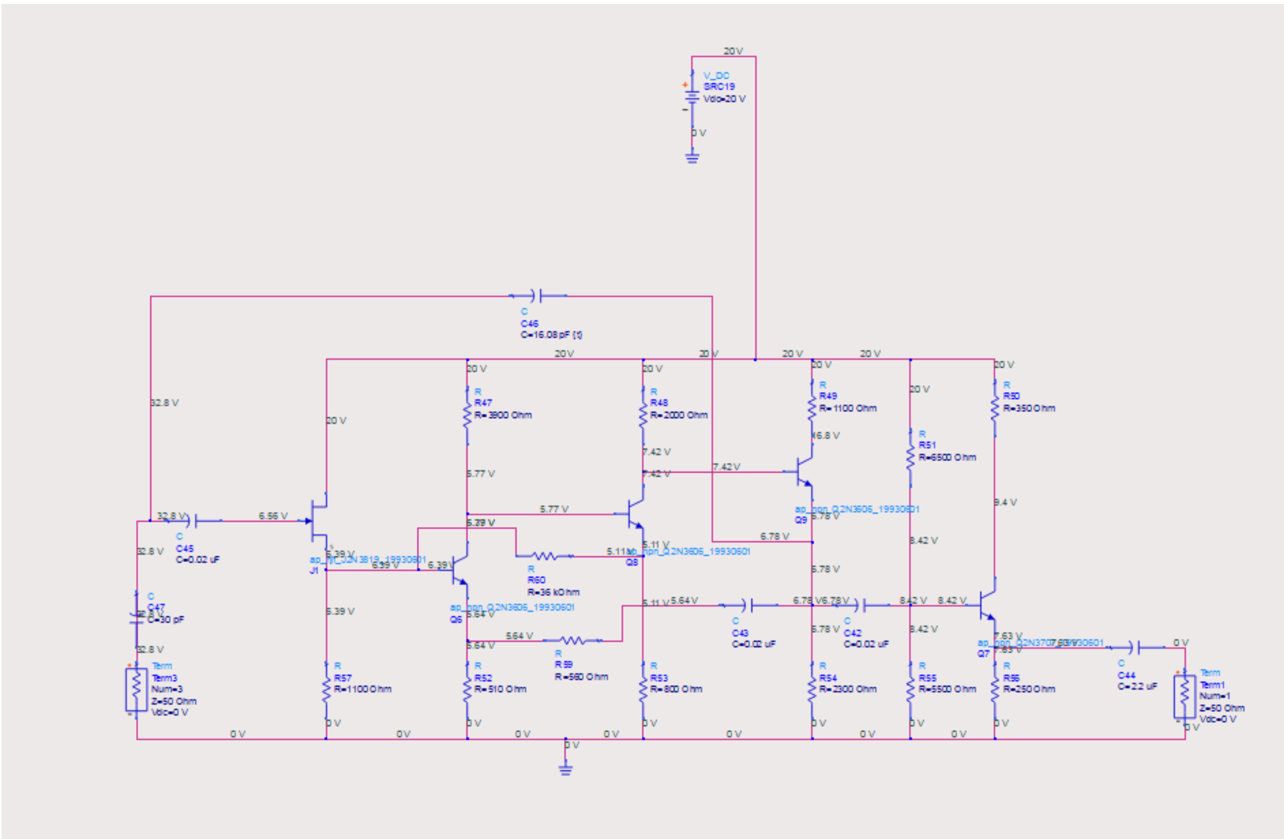


Figure 4.8: ADS schematic of Harris and Myers' circuit [13]

The simulation indicated an inverted impedance curve as expected, however there is a definite decrease in range of impedance values over the classical Linvill transistor-based circuit, which could be unstable. Figure 4.9 shows the reactance curves vs frequency from the ADS simulation. An ideal value of -150pF was chosen as a best fit curve to the reactance of the non-Foster circuit.

The blue curve represents an ideal inverted 150pF capacitor and the red curve represents the resulting output of the non-Foster circuit shown in Figure 4.8.

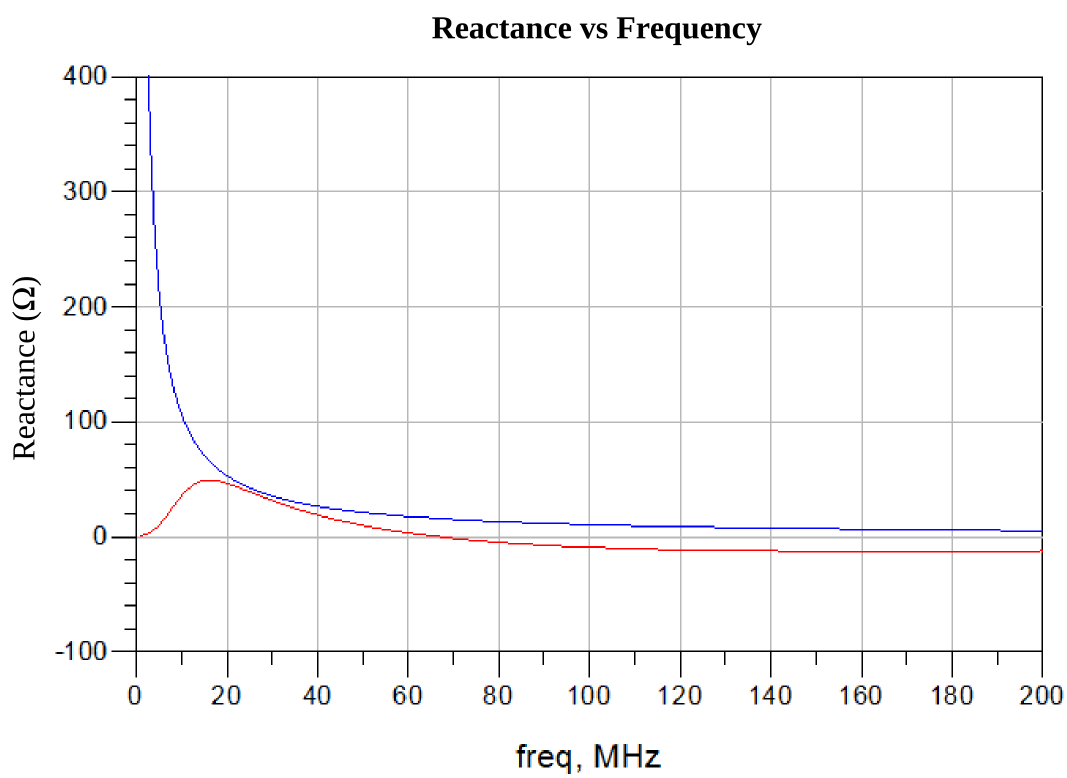


Figure 4.9: Reactance vs frequency of the schematic from Figure 4.8

Overall, the agreement between the ideal inverted capacitance reactance and the non-Foster circuit reactance is good above about 20MHz. There is a deviation below 20MHz as a result of the physically realizable nature of the circuit. Since an infinite reactance is not physically realizable, the deviation below 20MHz is to be expected. The non-Foster circuit improves performance at the lower frequencies where the reactance of the antenna is large. At the higher frequencies, where the antenna reactance is negligible, the non-Foster circuit contribution diminishes. This can be seen in Figure 4.9 at 70MHz where the reactance dips below zero. The tunability feature, shown later, can shift this zero-crossing to provide optimized performance. Next, the additional transistor is added to the circuit for tunability. The transistor is added in the same manner as for the classic Linvill circuit, in series with the load impedance. This is shown in Figure 4.10.

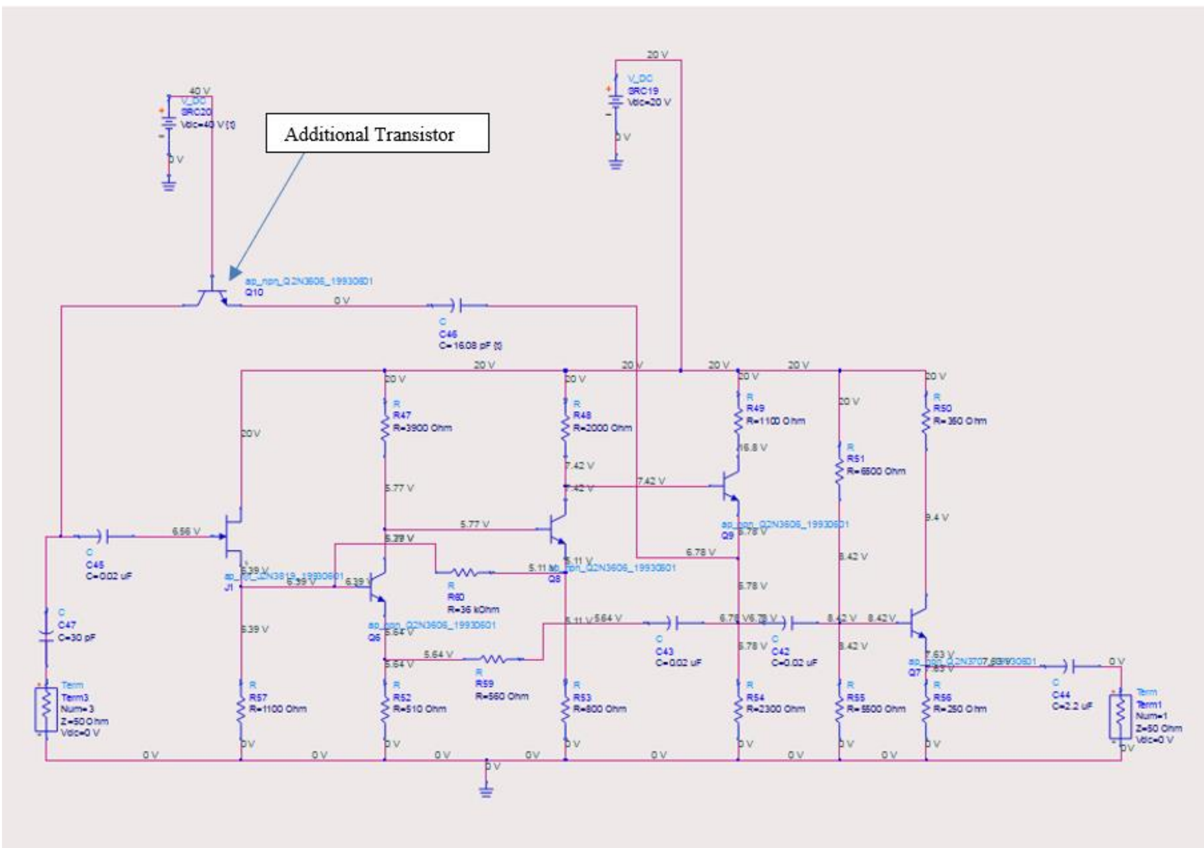


Figure 4.10: Harris and Myer's circuit with additional transistor for tunability

The simulated results for the tunable version of Harris and Myer’s circuit are shown in Figure 4.11. The blue curve represents an ideal inverted 240pF capacitor and the red curve represents the output of the non-Foster circuit shown in Figure 4.10 with the additional transistor added for tunability. Notice that the previous reactance curve, in Figure 4.9, was compared to that of a 150pF inverted capacitor. This shift in capacitance is due to the modifying tuning range of the added transistor.

Reactance vs Frequency

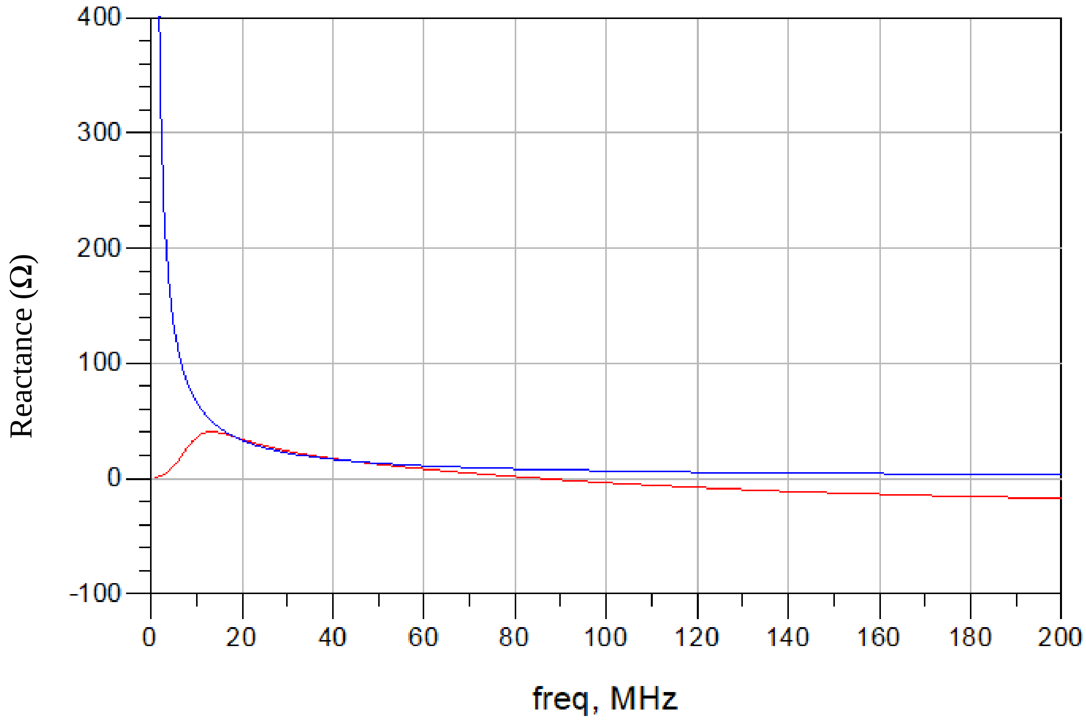


Figure 4.11: Reactance of the tunable improved-stability non-Foster circuit

With every additional circuit modification, there is a slight degradation in overall circuit performance. Performance degradation is traded for unconditional circuit stability. The added stability limited the range of inverted reactance that can be used. This observation agrees with that of the original theory of an inverted impedance and circuit stability. As the overall circuit edges closer to perfect agreement between un-inverted and inverted capacitances, resonances can occur, and the circuit approaches instability (like a marble on top of an inverted bowl). With the additional transistor, the agreement between the ideal inverted reactance and the simulated results slightly worsened. Note, however, that the zero-crossing has now shifted to 80MHz with the tuning transistor in this configuration, rather than the 70MHz zero-crossing as shown in Figure 4.9. It is important to note that in a physical circuit, reduction in reactance of the matched antenna match improves performance. At higher frequencies below resonance, the magnitude of the reactance of a capacitor is small, and a matching circuit may not be needed. In the case of a non-Foster matching network, the desired improvement is focused on the lower frequency region. For the figures shown above, there is actually a degradation in performance at the higher frequencies (above 80MHz), which is traded for improvement at the lower frequencies. A slight increase in reactance magnitude, perhaps by a tiny amount is an acceptable trade for substantial decrease in reactance magnitude at the lower frequencies. Also, with practical antennas in real environments, it is unlikely that their realized impedance will represent that of an ideal lumped element component.

The NDF function was applied to quantifying stability of the tunable version of Harris and Myers' circuit. The admittance matrix as an output of the ADS simulation is used for the full active network determinant. For the normalizing term, a determinant of a known stable variant circuit is used. A hybrid-Pi model replaces BJT transistor [49] as shown in Figure 4.12. This model uses a voltage controlled current source, valued at the base-emitter voltage (V_{be}) times the transistor transconductance (g_m). To disable the transistors, the current source is set to zero. The FET transistor model is shown in Figure 4.13 [50], with the voltage controlled current source valued at the gate-source voltage (V_{gs}) times the transistor transconductance (g_m).

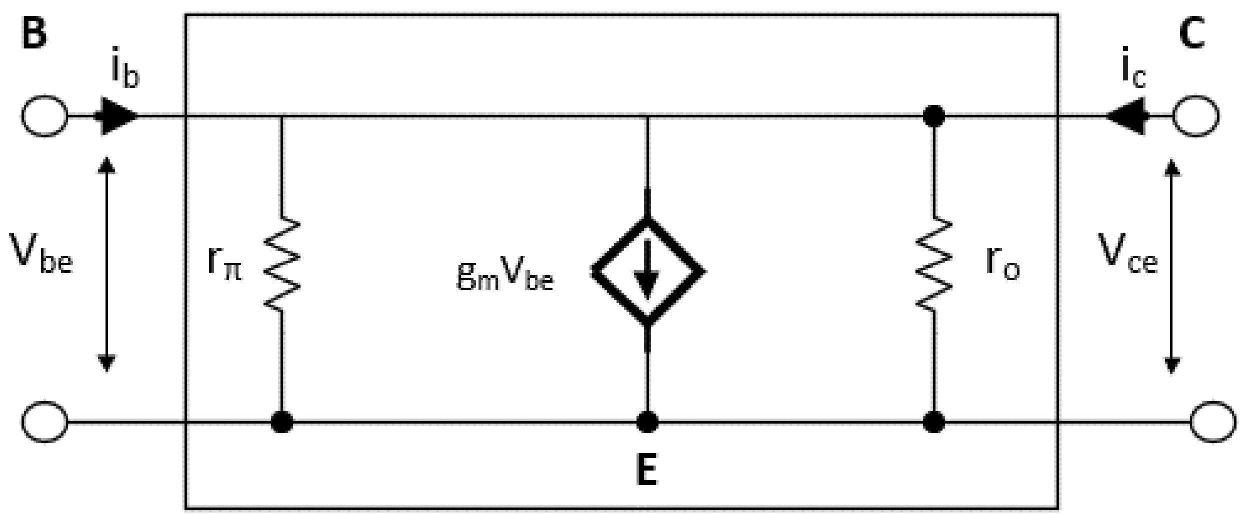


Figure 4.12: BJT small signal transistor model reproduced from [49]

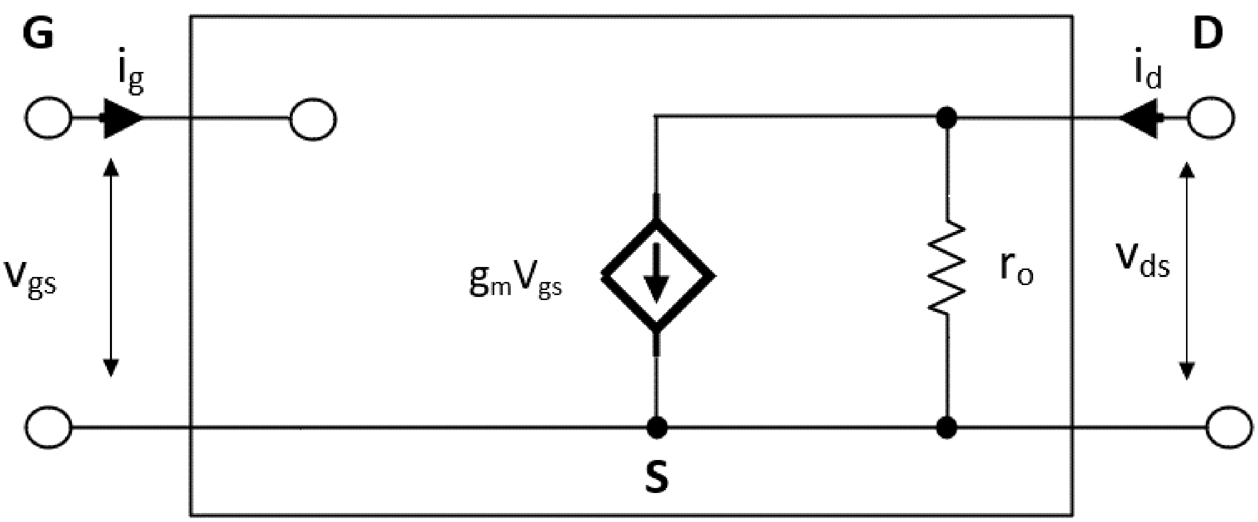


Figure 4.13: FET small signal transistor model reproduced from [50]

Figure 4.14 shows the resulting Nyquist plot of the NDF for frequencies between 1kHz and 100GHz. The large frequency variation impedes determining the number of encirclements at the lower frequencies. Figure 4.15 further clarifies this determination by plotting the encirclements as a function of frequency where it clearly shows that the circuit is stable since the encirclements are less than unity. Recall that a NDF origin encirclement occurs when the locus of the NDF makes a complete revolution about the origin in the complex plane. The phase of the NDF is required to change by more than 2π for an encirclement. The plot breaking out encirclements by value in Figure 4.15 is simply the phase of the NDF divided by 2π.

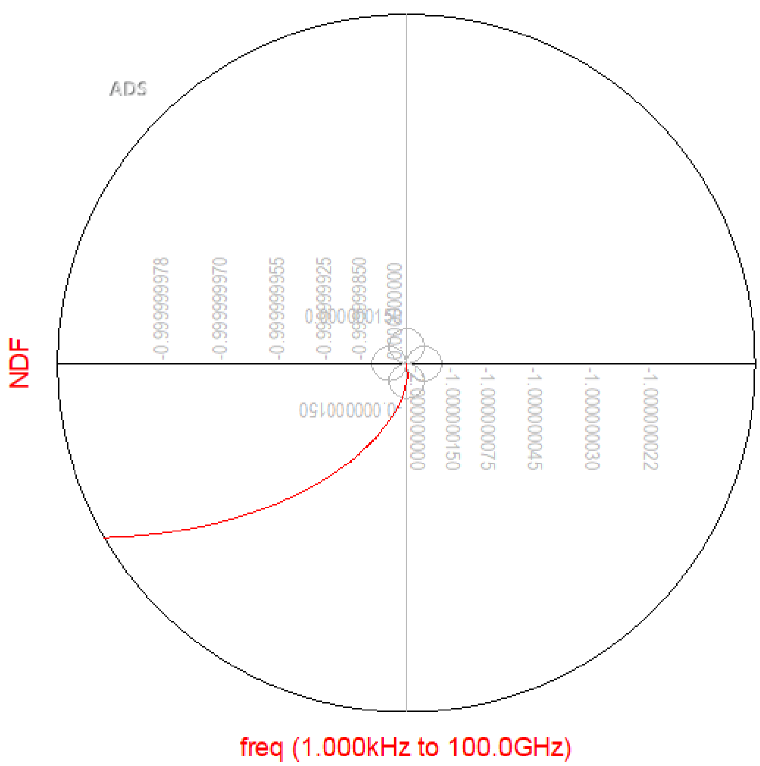


Figure 4.14: NDF Nyquist plot

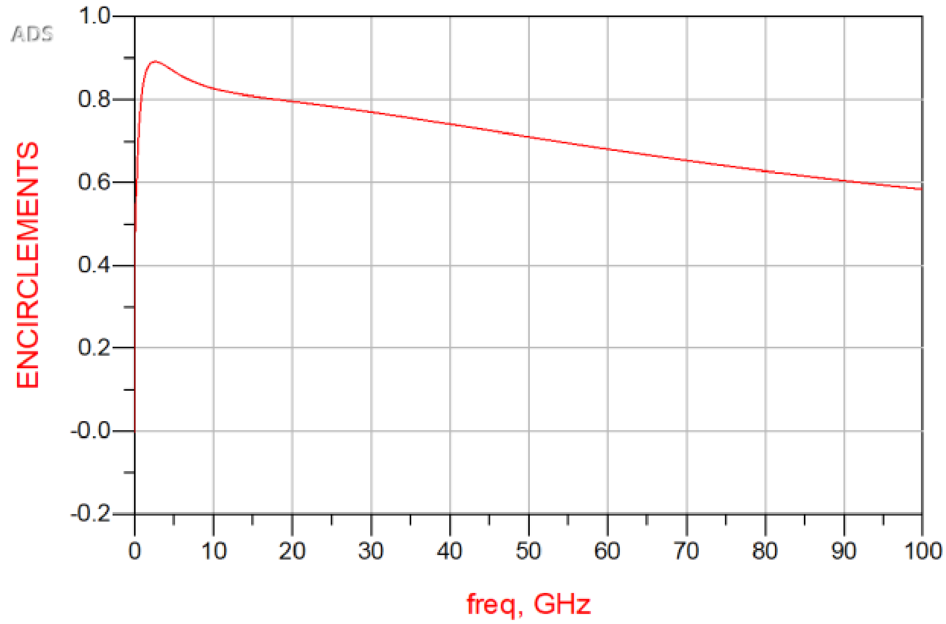


Figure 4.15: Number of encirclements of NDF

4.4 Physical Implementation Results

The first experiments are carried out with a circuit fabricated according to the design of Harris and Myers [13]. The goal is to achieve a stable circuit that measures an inverted reactance curve. If the stability and performance are verified, the plan is then to fabricate the modified circuit and observe tunability. The second circuit is then to be fabricated and measured as shown in Figure 4.16 and Figure 4.17 respectively. This is the modified-stability but non-tunable version of the non-Foster circuit. The measured results in Figure 4.17 appear to be unconditionally stable, and while not perfect, exhibit an inverted reactance curve for a large portion of the simulated frequency band. If the circuit is unstable (like the classic Linvill version), the circuit would not have demonstrated an inverted impedance slope. The unstable circuit would react in a positive feedback fashion and the transistors would output their maximum voltage which is the supply voltage and unrelated to the input. Note the discontinuity of the curve at the lower edge of the frequency band and this was evident to a lesser degree in the simulated results in Figure 4.11 (the bump in the red curve at the lower frequencies). This discrepancy likely resulted from applying transistors in the circuit that differ from those simulated. They are replacements for the now obsolete transistors. The original circuit transistors were available in the late 1960s and many of them are no longer available but have recommended replacement parts. This non-Foster circuit is designed to provide improved impedance match of an antenna designed to operate from 20 MHz to 6 GHz (as discussed in Section 5.1.1). Because this antenna is electrically small at 20MHz, and resonant at around 150MHz, the main focus for extracting improvement from the non-Foster circuit is in the 20MHz to 150MHz range. It is understandable that there might be a degradation of performance above 150MHz due to the added noise of the non-Foster circuit. However, the gain improvement at the lower frequencies is generally desired despite the degradation at higher frequencies due to the antenna having sufficient gain in that region.

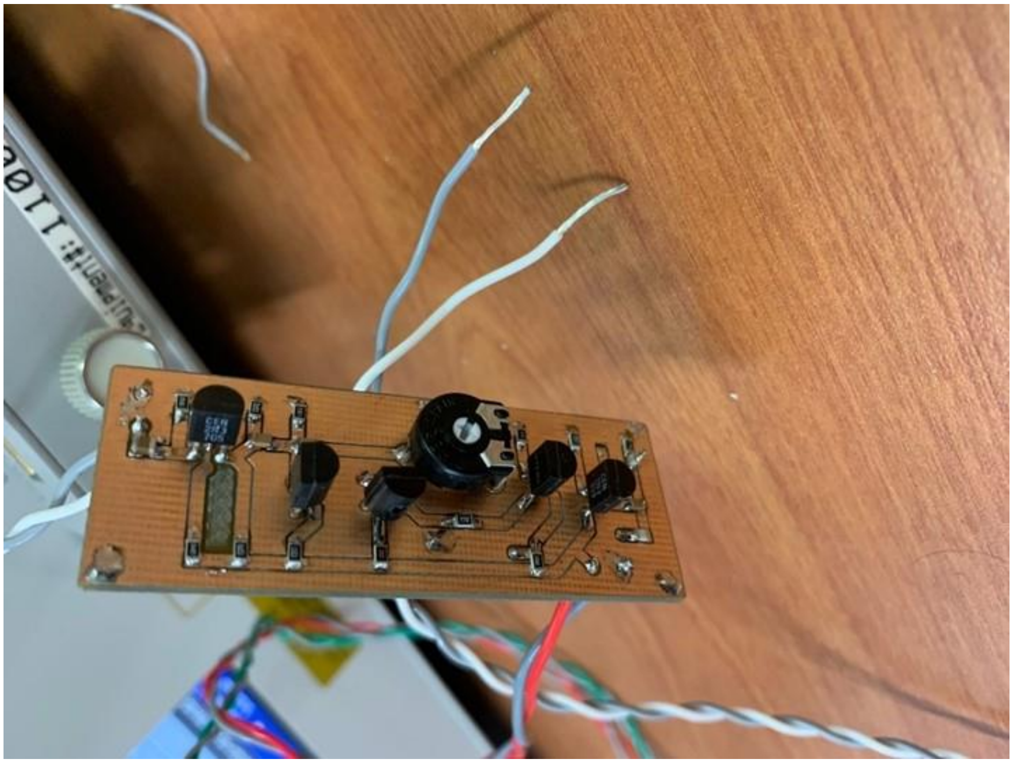


Figure 4.16: Non-Foster circuit with improved stability [13]

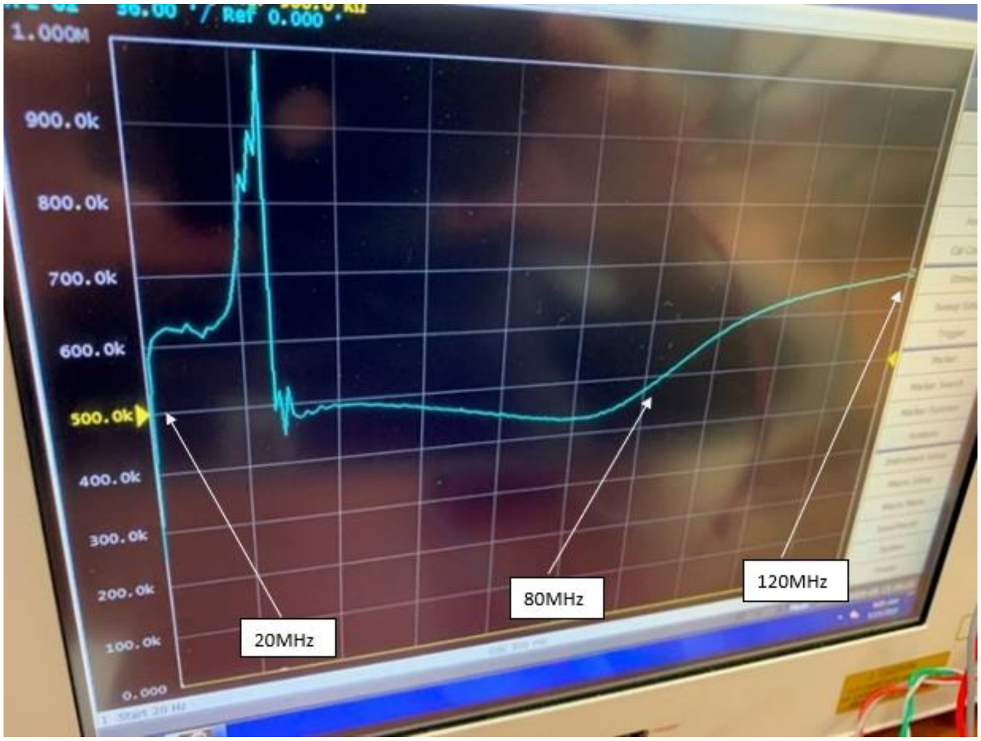


Figure 4.17: Measured reactance of non-Foster circuit with improved stability [13]

Figure 4.18 shows the circuit modified for tunability by the addition of a tuning transistor. The measured results are shown in Figure 4.19. The goal is to observe a changing reactance as the bias voltage on the third transistor is adjusted. Changing reactance is observed, with a general trend of higher reactance magnitude with an increased bias voltage. The measured results show that when the bias voltage on the added transistor is changed from 3V to 20V, the reactance of the output changes as well. Because the reactance can be changed during circuit operation, this is considered a rapidly tunable non-Foster circuit. As the tuning transistor moves out of the linear region, either to the off region or the fully saturated region, there is little change in reactance. Although the tuning range is slightly shifted in the reactance scale from that simulated and more limited than that of the original circuit, this is to be expected due to differences in the simulated models and the physical transistors which were applied.

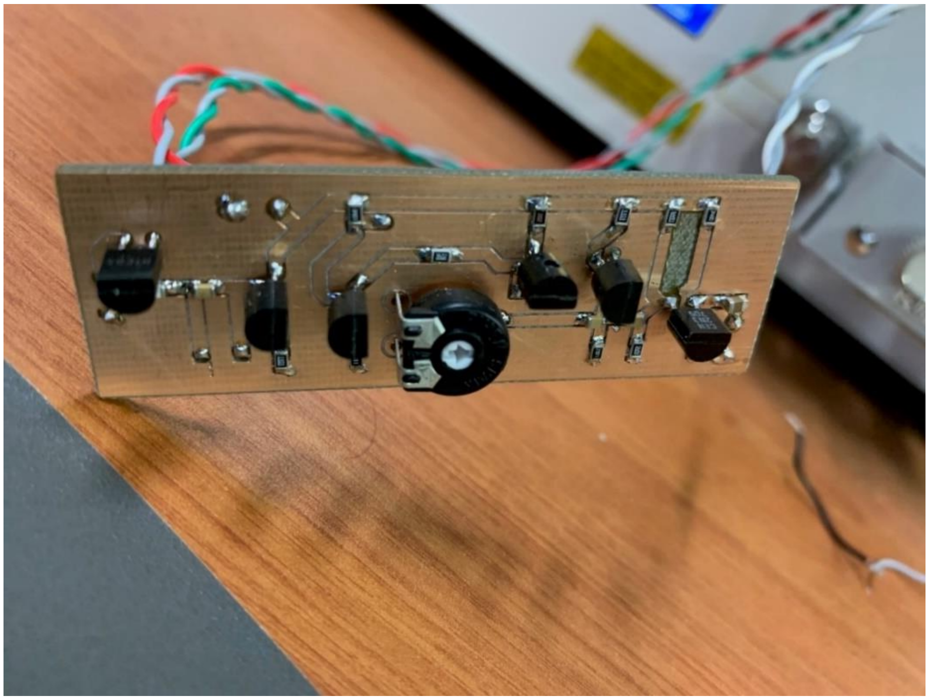


Figure 4.18: Modified (for tunability) non-Foster circuit with improved stability

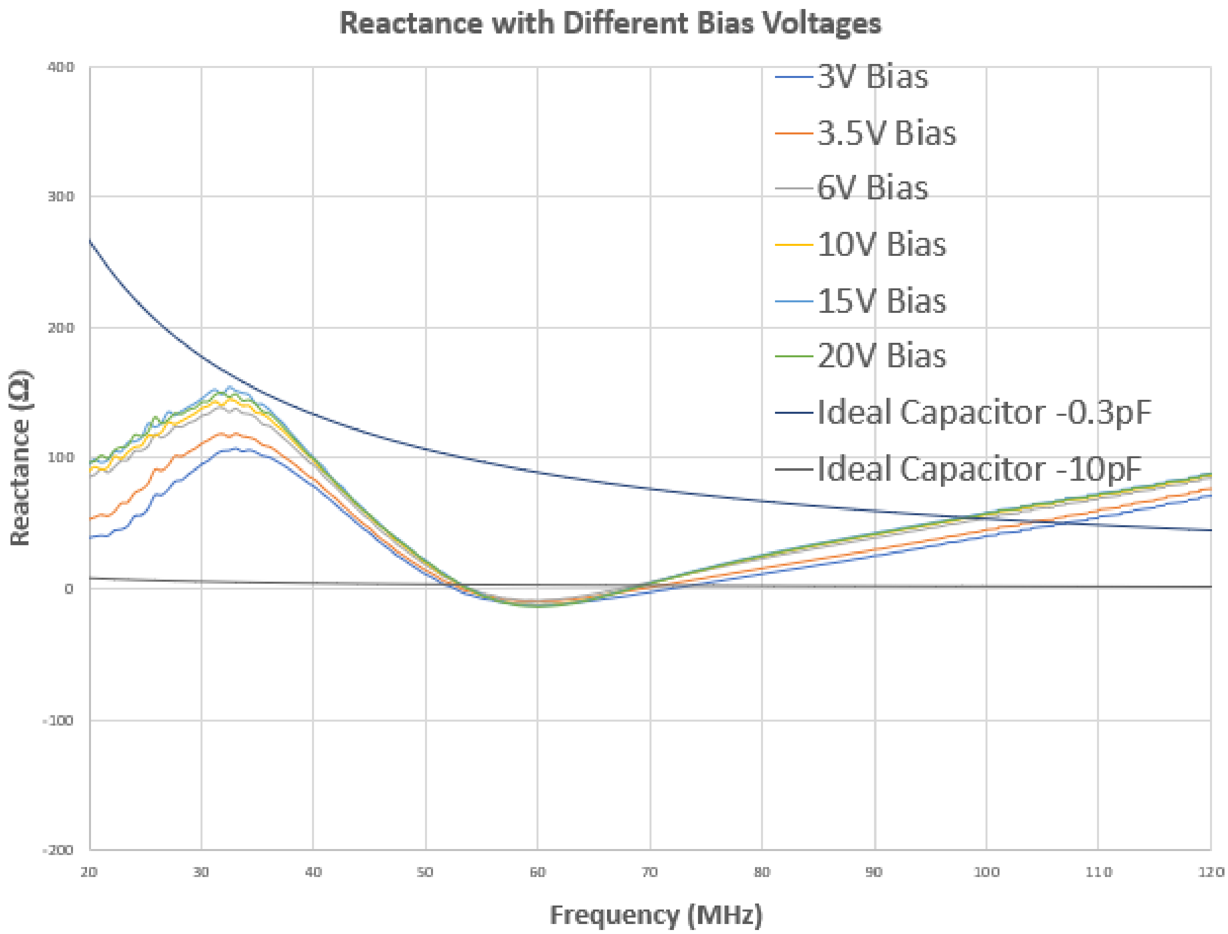


Figure 4.19: Measured reactance of tunable non-Foster circuit

At first glance, it seems that there is poor agreement between the ideal curves and the measured curves. However, there are a few things to note with this measurement. First, the non-Foster circuit reactance is very dependent on the given load. In fact, the load which terminates the non-Foster circuit will change the effective reactance of the non-Foster circuit. This can be seen in simulations by varying the termination impedance. Traditional transistor-based non-Foster circuits will move into the unstable region when measured unloaded, and all literature that presents measurements of non-Foster circuits describes behavior when paired with a proper load (such as an antenna). Simulation results of a "non-loaded" non-Foster circuit can be given to help with design parameters, but standalone measurements of a non-Foster circuit are seldom seen in the literature. Harris and Myer's [13] addition to Linvill's classic transistor-based non-Foster circuit has forced the circuit into a stable region despite being unloaded. Hence, the measurement here is indicative of the improved stability circuit and it does exhibit a negative reactance slope for part of the frequency curve.

Second, it should be noted that any inverted reactance helps improve the wideband match of a reactive load even if the inverted reactance slope does not match that of an ideal lumped element. The antenna discussed in Section 5.1.1 happens to have a capacitive reactance curve. However, unless this curve perfectly matches an ideal capacitor, the agreement between the non-Foster circuit and an ideal capacitor becomes a moot measurement. The curves of Figure 4.19 showing the ideal negative capacitance values are for reference to indicate the tuning range observed on the unloaded circuit. The loaded circuit will likely be more well behaved and have a more consistent slope across a wider frequency range.

5 Practical Applications of a Tunable non-Foster Circuit

Non-Foster matching networks have been used historically for a variety of different applications. The original application for the transistor-based non-Foster circuit helped improve the efficiency of telephone lines [12] and many applications afterwards were used for antenna matching. Tunable non-Foster circuits bring an additional degree of freedom for performance optimization. The discussion that follows will detail some specific applications that can benefit from the added tunability of the non-Foster network. Achieving on-the-fly tunability opens new possibilities for applications where non-Foster networks might have been previously limited.

The tunability feature adds an additional class of applications of non-Foster circuits. The ability to rapidly change the group delay opens opportunities for the use of non-Foster circuits in phase shifter devices. These applications require a constant response across all frequencies, which differs from the typical antenna matching case. With the careful selection of transistors, Buyantev and Kholodnyak have demonstrated non-Foster phase shifter circuits which can operate at higher frequencies [29]. A caution when utilizing such a phase shifter is that a conventional phase shifter which might use transmission line sections will have an ideal phase dependence across frequency (based on the speed of light in that medium), and this is critical for many applications, as operation across multiple frequencies is assumed. However, for phase inverters which do not use typical physical transmission line sections, this phase slope (whether positive or negative) is not guaranteed. Reference [29] also addresses this issue and how one can ensure the phase slope is maintained over a broadband frequency range. Other uses are possible for non-Foster circuits and negative impedance inverters in general, as they provide similar functionality to a passive capacitive or inductive element but with the ability to affect a negative capacitance or inductance. The inverted impedance slope across frequency is beneficial to many applications.

Additionally, traditional applications of non-tunable non-Foster circuits can benefit from the added tunability. Wideband antenna matching, one of the most common uses for non-Foster circuits, can be expanded with the added tunability. The allowable variance from the original circuit parameters can help match antennas over a wider bandwidth when compared to traditional non-Foster circuits. The next sections will describe specific applications where the tunable non-Foster circuit presented in Chapter 4 can provide benefit.

5.1 Wideband Matching of an Electrically Small Antenna

Chapter 2 introduces the use of non-Foster networks as a means of achieving a wideband antenna match. Non-Foster matching networks are useful for wideband antenna matching, mostly benefitting electrically small antennas. For antennas which operate at higher frequencies, an antenna of resonant length is not difficult to realize. However, systems with low operating frequencies frequently utilize electrically small antennas. Other problems could manifest themselves with wideband operation of resonant length antennas which will be discussed later. For antennas which operate at very low frequencies, conventional matching networks have long been utilized as a practical means for operation. For example, an AM radio antenna, which if built at proper resonant size of ½ λ at 570kHz, the required length would be 263 meters. This is a very common frequency of automobiles and portable, etc. radios. At this frequency, an electrically small antenna is an inevitable requirement due to the large wavelength. Conventional matching networks can reduce these antenna sizes to the practical length for consumer devices that utilize this application. Since these radios do not exhibit wideband operation, this is a sufficient solution. A radio channel, being an audio application, can completely sample the human sound spectrum of 20kHz with a 40kHz sampling rate, because of the Nyquist sampling criterion. For a 40kHz bandwidth at 570kHz, this is less than ten percent bandwidth and is not considered to be wideband. Practically speaking, the bandwidth of a radio channel is even smaller as it does not need to encompass the entire human hearing spectrum. A Smith chart derived RLC matching network suffices as a solution for this application despite the relatively high Q of the required matching network.

The issue of wideband operation at low frequencies suggests non-Foster networks as a solution for impedance matching. Low frequency operation forces the use of an electrically small antenna for practical size constraints. Some applications are discussed in the following subsections.

5.1.1 Resistively Loaded Dipole

Dipole antenna gain patterns and impedances vary rapidly with respect to frequency. One way to suppress the associated resonances is to apply a distributed resistive loading to the dipole. Originally introduced by Wu, King & Shen in 1965 [51,52], this loading is designed to suppress reflections from the ends of the dipole that result in standing waves. The current distribution along the dipole antenna can form many local minima and maxima. Using a restively loaded architecture can help reduce these. The resulting resistively loaded antenna pattern can be well behaved across a large bandwidth, but still requires a matching network to utilize the antenna with a practical RF receiver at low frequencies. This makes it an ideal candidate to be paired with a non-Foster matching network. Direction finding systems in particular are wideband can greatly benefit from this technology.

Here we present the combined use of a resistively loaded dipole and non-Foster matching network combination. The figures below show the difference in beam patterns of a resistively loaded dipole antenna and a conventional unloaded counterpart. The antennas are otherwise identical, i.e., they have the same length and geometry with the only difference being the resistive loading. Figure 5.1 shows the comparison of return loss between resistively loaded and non-loaded dipoles across the frequency band. Figure 5.2 shows the comparison of the antenna pattern at one frequency (1.523GHz). These results were obtained using the commercial full-wave computational EM code WIPL-D [53].

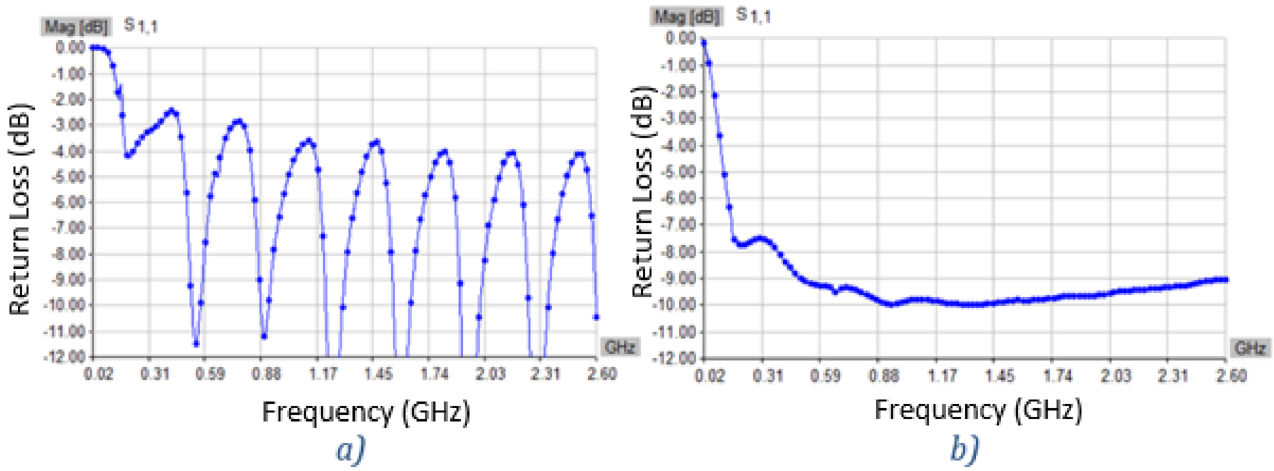


Figure 5.1: Return loss of (a) conventional dipole and (b) resistively loaded dipole

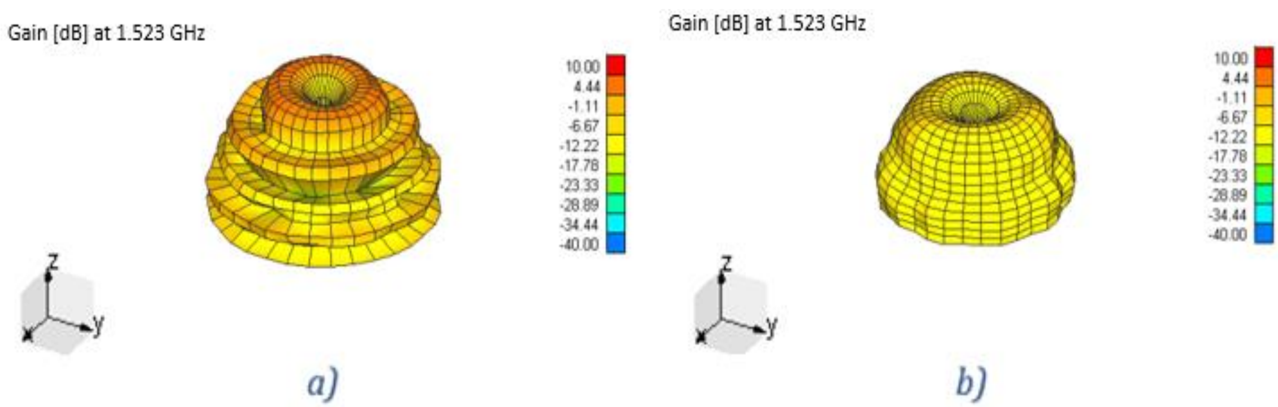


Figure 5.2: Pattern of (a) conventional dipole and (b) resistively loaded dipole

As a dipole operates across a wide bandwidth, the length of the dipole becomes multiple wavelengths as the frequency increases. The traditional sinusoidal current distribution across the dipole remains consistent, but now involves many cycles resulting in several zero crossings and min/max repetitions as the current traverses the length of the dipole. The effect of the standing wave current distribution is evident in the rapid pattern variation for the unloaded dipole. The nulls corresponding to specific θ angles are a direct result of the irregularities present without resistive loading. Resistive loading helps reduce these local minima and maxima current fluctuations. Although the resistively loaded dipole impedance and pattern are generally better behaved, performance is not ideal and could be improved with a proper matching network. The antenna, only 36” or 0.91m in length, was designed to operate over a wide frequency band, from 20 MHz to 6 GHz. The wide bandwidth makes a conventional passive matching network unsuitable especially with regards to the lowest frequencies. Note that the gain peaks and dips of the unloaded antenna are suppressed in the resistively loaded variant resulting in broader patterns as well as wider bandwidths.

Dipoles with resistive loading are one example of antennas operating across a large bandwidth. Should the resistive loading prove unsuitable for a particular applications, other methods and antenna types exist for wideband operation. Dagefu, Choi, Sadler, and Sarabandi [54] detail a survey of different types of small, low frequency antennas that span multiple antenna structures and applications for the HF-UHF (few MHz to several hundred MHz) frequency bands. Small antennas in this frequency range are common due to the large size of resonant antennas. Reference [54] also includes a discussion on several types of matching networks for these small, low frequency antennas.

5.1.2 Design of a Matching Network for Electrically Short Resistively Loaded Dipole

Despite being a tunable circuit, the non-Foster network introduced here still needs a baseline starting point for the design. The tunability can be used to optimize the impedance match. To begin, the simulated impedance and S-parameters of the resistively loaded dipole antenna were used as a baseline for setting the impedance of the non-Foster matching circuit. Since the non-Foster matching circuit emulates an inverted impedance, the real and imaginary impedance components of the resistively loaded dipole are examined. Figure 5.3 shows the simulated antenna impedance split into resistance (Ω) and reactance (Ω) of the resistively loaded dipole antenna discussed earlier.

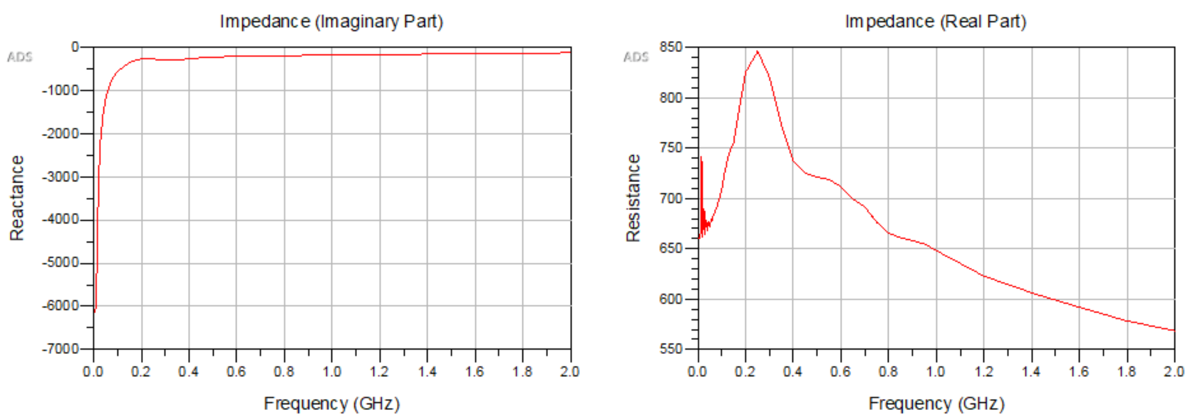


Figure 5.3: Simulated impedance of the resistively loaded dipole antenna

The reactance curve of the resistively loaded dipole alone represents that of a passive capacitance element. This makes it an ideal candidate for matching using a non-Foster circuit which resembles a negative-capacitance reactance curve. It should be noted that the designed non-Foster circuit might not exactly match the capacitance value given with the resistively loaded dipole. However, any inverted, or "negative-capacitance" value will help reduce the absolute value of the resulting reactance and increase power transfer.

Next, examine the impedance curves of the non-Foster network. The simulation of the stand-alone non-Foster circuit can be misleading in terms of determining the input and output impedance. The non-Foster is made to be paired with a load which is not 50 ohms on the antenna side and an impedance transformer is likely required. The simulation software, ADS, however, is applied here with an inherent 50 Ohm termination with a transformer in calculating S-parameters. The general performance of the non-Foster circuit can be assessed by examining the resulting output reactance of the circuit when terminated with 50 Ohm terminations at each port. Figure 5.4 shows the reactance curves of the input and output ports of the non-Foster circuit. Here, they are loaded with an ideal 50 Ohm termination on the ports.

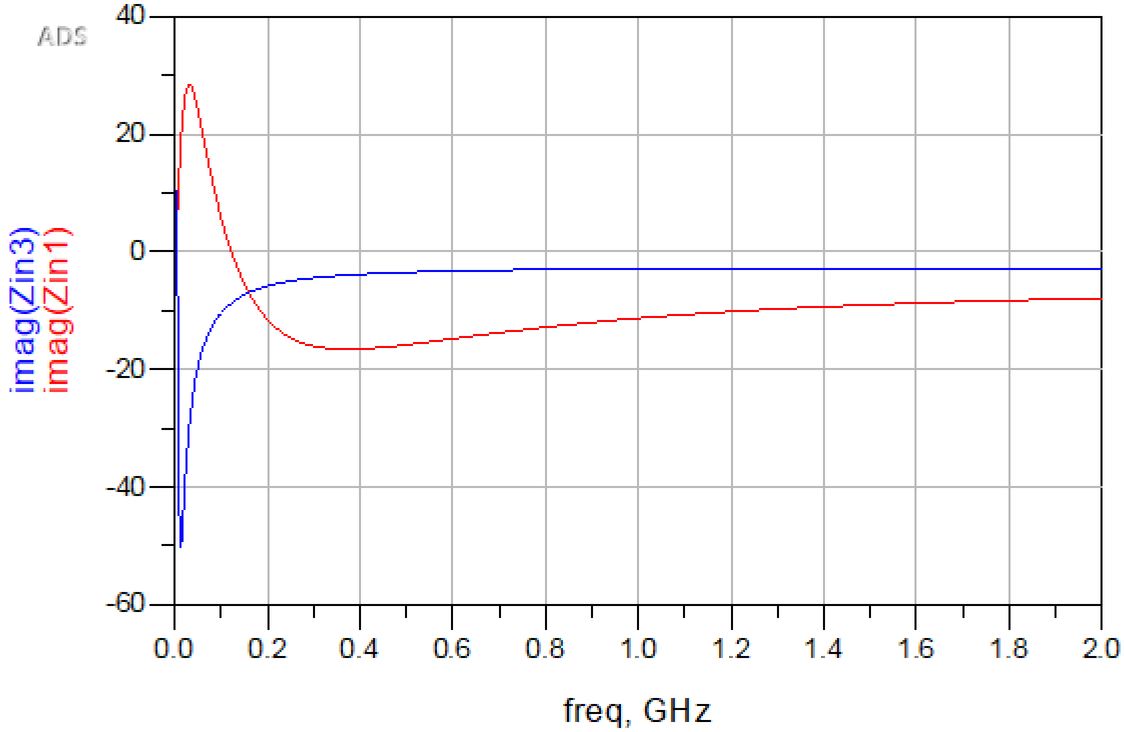


Figure 5.4: Reactance curves of the input and output of a non-Foster matching network

This verifies the operation of the non-Foster circuit as an impedance inverter. The input reactance, colored blue, on one end, looks like a positive valued capacitance, and on the other end, colored red, it looks like a negative valued capacitance. Note that the reactance value, while sloped like an inverted capacitor, is still below zero. This demonstrates the concept behind the non-Foster circuit but reiterates the need for a practical non-Foster circuit to be terminated in a proper impedance for operation. This non-Foster circuit, by nature of being open-circuit-stable (OCS), exhibits a higher input impedance and a lower output impedance. In this case, the "input" of the circuit is to be applied in parallel with the antenna and the "output" is applied to the receiver circuitry. At the input side of the matching network, the relatively high input impedance of the antenna at low frequencies becomes less of a matching problem. Similarly, the lower output impedance of the non-Foster circuit makes it a better candidate for attaching the receiver circuitry.

The matching network utilized here is a tunable non-Foster network. The tunability is done by the addition of a tuning transistor as discussed in preceding sections. Varying the base voltage of this tuning transistor changes the reactance curve of the matching network allowing for rapid tunability. This has many useful applications, such as adjusting for antennas deployed in various environments, but also, as shown later, for maximizing gain across the frequency spectrum. Figure 4.19 shows the measured reactance curves across frequency of the tunable non-Foster network. The different curves represent different tuning parameters which modify the impedance. Note that this circuit is the improved stability circuit which was discussed in Chapter 4 and therefore has been "degraded" to account for the stability measures guaranteeing unconditional stability.

The resistance curve is also of concern for antenna matching. Conventional matching networks need to suppress the non-zero reactance as well as transform the impedance of the circuitry to that of the antenna. Non-Foster matching networks generally have low output resistance and exhibit a flat resistance curve when plotted across frequency. Most antennas have a higher resistance value than the typical 50-ohm impedance of a transmitter or receiver. A transformer can be utilized to transform the impedance from 50 ohms to a different value. The issue arises when the resistance varies with frequencies. This makes wideband impedance transformation difficult. Except at the lowest frequencies, the return loss of the resistive dipole is reasonable and fairly flat throughout the band as shown in Figure 5.1 when matched to a high impedance. A wideband impedance transformer would leave only the low frequency capacitive portion of the resistively loaded dipole to be matched. The non-Foster circuit would enhance this low frequency match.

5.2 Physical Implementation Results & Noise Analysis

This section details the results of combining the physically implemented antenna and circuits as well as discussion on the noise of the system. The simulations conducted in the previous section were built on two-layer FR4 dielectric PCBs. There were several steps of iteration on physical circuits because of battling stability issues as discussed in Chapter 4. The final version, which is unconditionally stable, is that discussed here. The circuit is optimized for the simulated antenna values shown in Figure 5.3. Figure 5.5 and Figure 5.6 show the resulting measurements of the transmission coefficient. Both figures have the same measured data, with Figure 5.6 showing how tunability can be utilized to improve performance across bandwidth. The non-Foster matched antenna is compared to a non-matched antenna and evaluated for performance across the frequency band. "ANT2" is the non-matched antenna, and the other curves are those with different tuning parameters of the non-Foster matching network. The plots show the performance below 100MHz where the non-Foster network will add the most value. As previously discussed, at these low frequencies, resonant antennas may be too large for practical use. Above these frequencies, the non-Foster network will provide diminishing gains, and eventually only be a noise contributor without providing for additional gain.

Figure 5.5 shows the antenna alone as compared to utilizing the tunable non-Foster network with two different tuning voltages. For the 6V base voltage, improvements are seen throughout the frequency range, but are more focused in the lower portion (30-65MHz). The 8.4V base voltage shows the opposite, where the most improvement is seen above 65Mhz over the unmatched antenna. Other voltages in between will give different curves, but these two are shown as examples of the tunable concept.

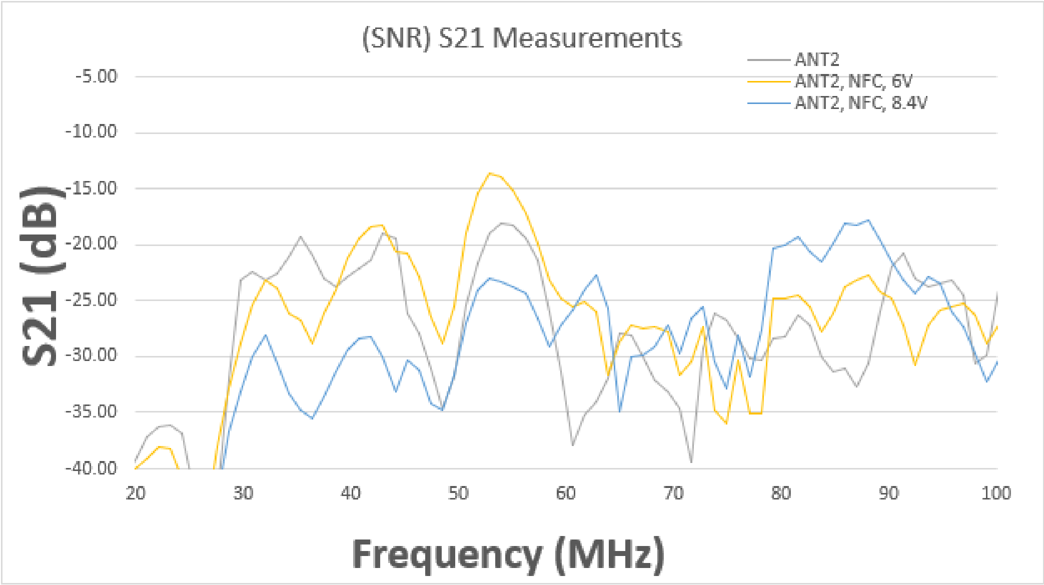


Figure 5.5: Measurement of resistively loaded dipole and tunable non-Foster network combination

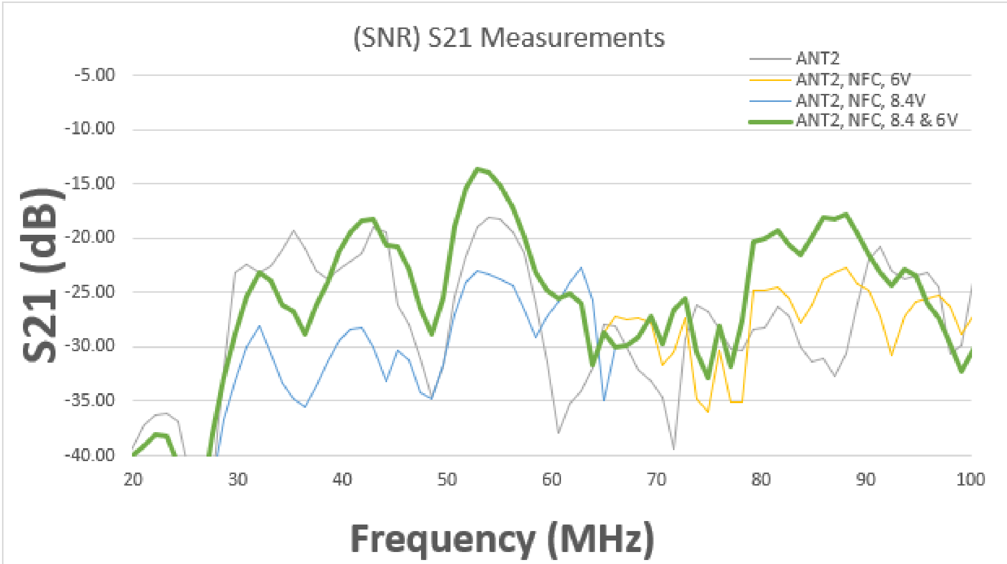


Figure 5.6: Measurement of resistively loaded dipole and tunable non-Foster network combination

Figure 5.6 shows the advantage of utilizing the tunable non-Foster network with the combination of the resistively loaded dipole. Because of the on-the-fly tunability, this tuning voltage can be easily adjusted based on the frequency of measurement. Figure 5.6 combines advantages from both tuning voltages, the 6V setting from 30-65MHz, and the 8.4V setting from 65-100MHz. This result demonstrates improved SNR over much of the bandwidth. Perhaps, a better metric to show improvements could be an integrated SNR over frequency. It should also be noted that this is simply a conceptual proof. In practical terms, there is nothing that limits the tuning from splitting the frequency range into more than two sections for tuning voltage. Further extension of this concept would be an infinitely variable algorithm to maximize the gain at every frequency and even bypass the matching network at the higher frequencies where it becomes less effective.

Several factors need to be considered when evaluating performance improvement. The SNR achieved with the non-Foster network includes both the additional signal received from the improved match, and the added noise which is resultant from the active circuitry in the non-Foster matching network. To observe improved performance, the added noise needs to be less than the additional signal strength. It is quite difficult to separately quantify the added noise since these circuits are only made to be measured together.

A network analyzer can be used for determining the performance of this non-Foster matching network. The network analyzer displays S-parameters as measured data, which is a signal level measurement at face value. However, understanding the conditions of the measurement is critical. The network analyzer will excite all ports with a known power level and measure the power received at all ports. It takes these measurements and then calculates the S-parameter matrix. Being an active device, the network analyzer has noise characteristics of its own, but this portion is taken out of the measurement. This is done by a coupler on the output port (e.g. Port 1), which samples the actual signal power and noise power which was transmitted. Dividing this sampled transmitted power by the power received at the opposite port (e.g. Port 2) will result in the displayed S-parameter measurement. While each of the measurements includes the inherent noise of the circuitry of the measurement device, the noise component is the common reference for both Port 1 and Port 2 and therefore it is divided out. However, additional noise, which may be added externally, such as contributions from the device under test, are not divided out. This is accurate for devices which only add noise to one port, such as our non-Foster matching network.

Figure 5.7 shows a block diagram of the measurement setup of the resistively loaded dipole and non-Foster matching network combination.

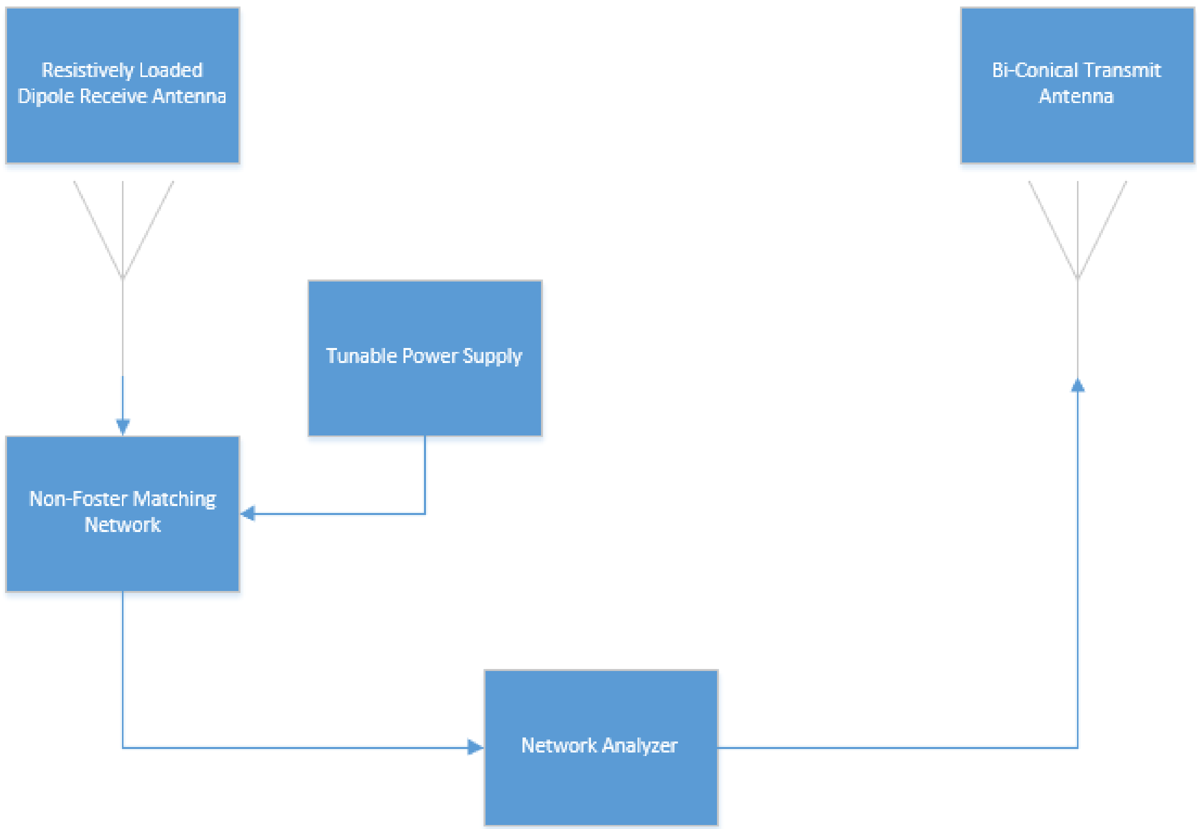


Figure 5.7: Measurement setup of resistively loaded dipole and non-Foster circuit combination

The measurement setup is as follows: Port 1 is connected to a transmitting antenna that does not change between any of the measurements. This serves as a baseline "transmitter" for evaluating our non-Foster network/antenna combination. Port 2 is connected to the "device under test." For the baseline case, rather than going through the non-Foster matching network, it is bypassed.

The resistively loaded dipole is connected directly to Port 2 on the network analyzer. Being a passive antenna, no noise is contributed, and we are simply measuring a “Signal to Noise ratio” with the noise being only that of the receiver circuitry in the network analyzer. The second measurement is taken under the same conditions but with a non-Foster network attached to the resistively loaded dipole. This is now an active network which adds both signal power because of the improved match, as well as some noise contributions because of the active transistors. Now, the “signal to noise ratio” that is measured includes both the additive noise of the non-Foster network, and the noise of the receiver circuitry in the network analyzer. Since the latter noise is constant between the two measurements, the difference between the two “S21” plots is the difference in SNR. This will serve as our performance metric of the non-Foster matching network.

5.3 Investigation of a Tunable non-Foster Circuit for Tunable Filters

Using non-Foster circuits to create active filters has been done previously [55]. With non-Foster filters, either transistors or Op-Amps can be used to achieve the negative impedance conversion. One benefit of an active filter is a broadband match with features that might not be possible with conventional passive filters. For example, non-Foster circuits can create negative group delay [57] as discussed below. Further, the tunable non-Foster circuit could replace bulky conventional tunable filters. Further discussion follows.

The group delay application requires a constant response across all frequencies, which differs from the typical antenna matching case. With the careful selection of transistors, Buyantev and Kholodnyak have demonstrated non-Foster phase shifter circuits which can operate at higher frequencies [29]. A caution when utilizing such a phase shifter is that a conventional phase shifter which might use transmission line sections will have an ideal phase dependence across frequency (based on the speed of light in that medium), and this is critical for many applications, as operation across multiple frequencies is assumed. However, for phase inverters which do not use typical physical transmission line sections, this phase slope (whether positive or negative) is not guaranteed. Reference [29] also addresses this issue and how one can ensure the phase slope is maintained over a broadband frequency range. Other uses are possible for non-Foster circuits and negative impedance inverters in general, as they provide similar functionality to a passive capacitive or inductive element but with the ability to affect a negative capacitance or inductance. The inverted impedance slope across frequency is beneficial to many applications.

A preliminary investigation in the use of non-Foster circuits for filters shows a category of negative group delay (NGD) filters which utilize non-Foster elements [55,56]. Group delay is defined as the derivative of the phase across frequency [55]. Negative group delay filters are those whose phase slope across frequency is positive. Traditional filters have a decreasing phase slope with frequency. This can be especially useful where insertion phase across a receiver circuitry needs to be limited as each component will likely have its own positive group delay. Nevertheless, a typical RLC filter topology can be used with the lumped elements being replaced by non-Foster ‘negative’ equivalents.

Using the tunable non-Foster circuit introduced earlier, a tunable filter can be achieved. In fact, tunable filters go hand-in-hand with wideband systems. Wideband systems, while having a great advantage of being agile and able to accommodate many different frequencies, also suffer from being vulnerable to all emitters across the wideband frequency. Traditional frequency focused systems include a filter which limits out of band interferers to limit both the noise contributions and the signal contributions of out-of-band sources. A device operating at a narrow frequency band with proper filtering is not susceptible to any devices which do not share a common frequency, whether they are intentional or un-intentional interferers.

Many commercial or industrial devices emit energy in parts of the frequency spectrum and can cause significant issues despite being for a different mission and frequency band altogether. For example, aircraft tracking radars in the USA, which are tasked with tracking planes from the moment they enter US airspace, have very high transmit power and very high gain antennas to cover the area of interest. For a system which is narrow band and does not share the same frequency band as these airport radars, they do not present an issue. However, for a wideband system, even when not operating in the same frequency band, a very high power emitter with a high gain antenna will surely cause significant issues. If this signal emitted from the aircraft radar were to be incident on the wideband antenna and receiver, it would likely saturate the receiver and cause a complete blindness across the entire operating bandwidth. Furthermore, a saturated receiver makes it difficult to determine the frequency of the offending signal to properly mitigate the issue. That means the operator must have apriori knowledge of the strong emitters and sufficiently filter them out before operation.

For wideband systems, one way to deal with filtering unwanted signals is the use of a tunable filter. Similar to the applications of a tunable antenna matching network, this allows agility in frequency and can be changed on-the-fly without disrupting system operation. Also, like some of the previously discussed approaches to tunable non-Foster circuits, there are several methods that have been used to create tunable filters that are widely available in the commercial world. The first is the use of a switched filter bank, where the user can select between any number of preset filters to suit their frequency of operation. Each individual filter will behave identically to a traditional filter and will provide the flexibility to choose any of the preselected filter configurations. However, as more and more filters are added to this switched filter bank, to provide maximum flexibility, the associated loss will likely increase. Another common approach to tunable filters is the use of varactors which was used as some of the early forms of tunable non-Foster circuits. Varactors provide an “infinitely” tunable range which give maximum flexibility in terms of small increments but limit the total range of frequencies which can be serviced with the varactor method of a tunable filter. Using a non-Foster circuit to create a tunable filter falls within a similar category of maximum flexibility within a small tuning range.

Figure 5.8 shows an ADS schematic circuit of a tunable filter designed with a tunable non-Foster circuit. The filter was modeled after a simple “LC” filter with just two components, a 1.5nH inductor and a tunable negative capacitor that is provided by the non-Foster circuit. The next 3 figures show the frequency response of the filter at different tuning voltages, and it is evident that the cut-off frequency can be changed rapidly by varying the third transistor voltage. Figure 5.9 shows the return loss of the tunable filter with a 1V tuning voltage, and similarly, Figure 5.10 shows the return loss with a 5V voltage and Figure 5.11 shows the return loss with a 10V voltage. If we evaluate the "1dB bandwidth," meaning where the insertion loss of the filter reaches -1dB, we can see that the filter cut off frequency can vary between 800MHz and 1200MHz for this parameter set. The classical filter, without the non-Foster circuit has a cut-off frequency of about 920MHz. Of course, this can be adjusted by changing the inductor value from 1.5nH as well as changing the inverted capacitance value in the non-Foster circuit. Similarly, as with traditional filters, one could add more stages to improve the filter cut off slope, etc. While the range of 800MHz to 1200MHz is a respectable tuning range, it is limited by the operating regions of the transistors. In one configuration, the third transistor is essentially "off" and in the other extreme it is saturated. Increasing or decreasing the tuning voltage beyond these parameters will not yield any additional change in frequency. As mentioned previously, if a different cut off frequency point is desired, the physical values of the circuit would need to be changed.

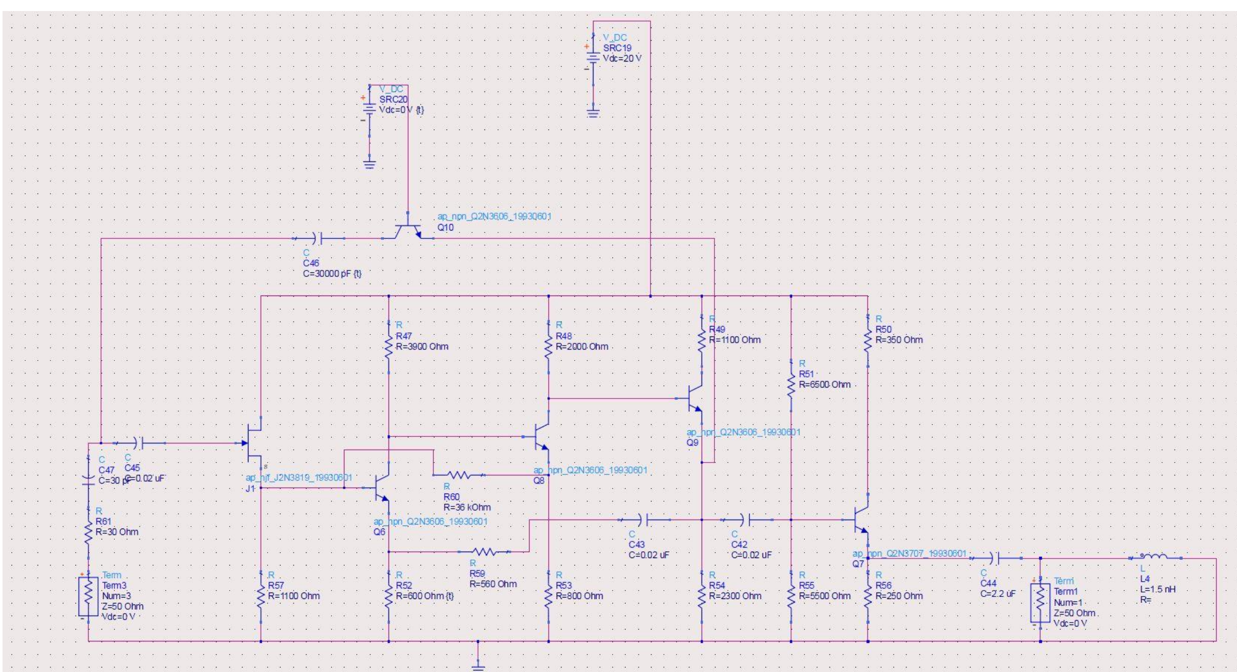


Figure 5.8: ADS schematic of a tunable filter using a non-Foster circuit

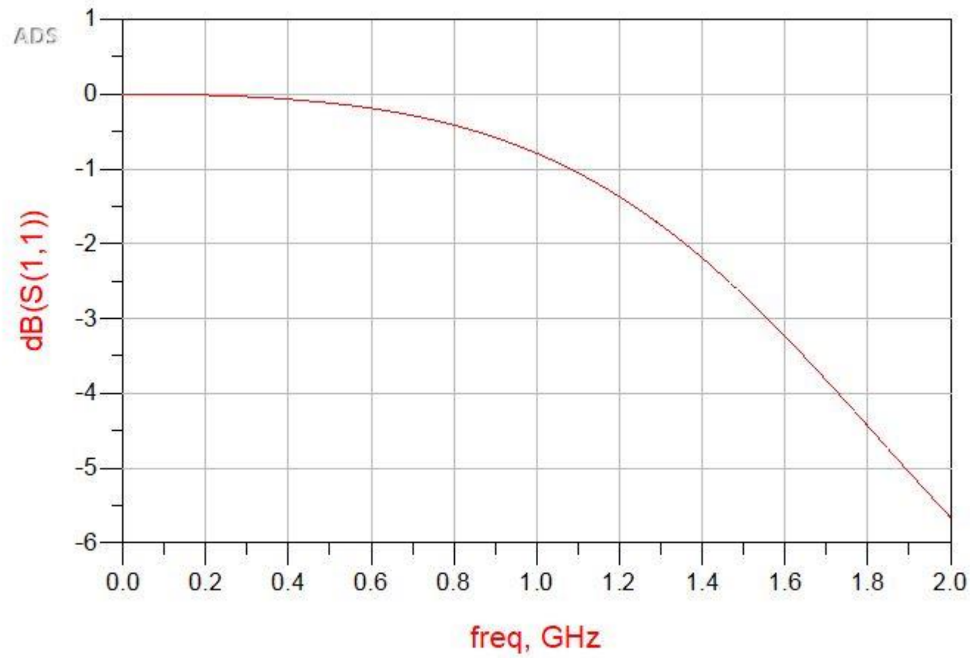


Figure 5.9: Tunable filter frequency response with a 1V tuning voltage

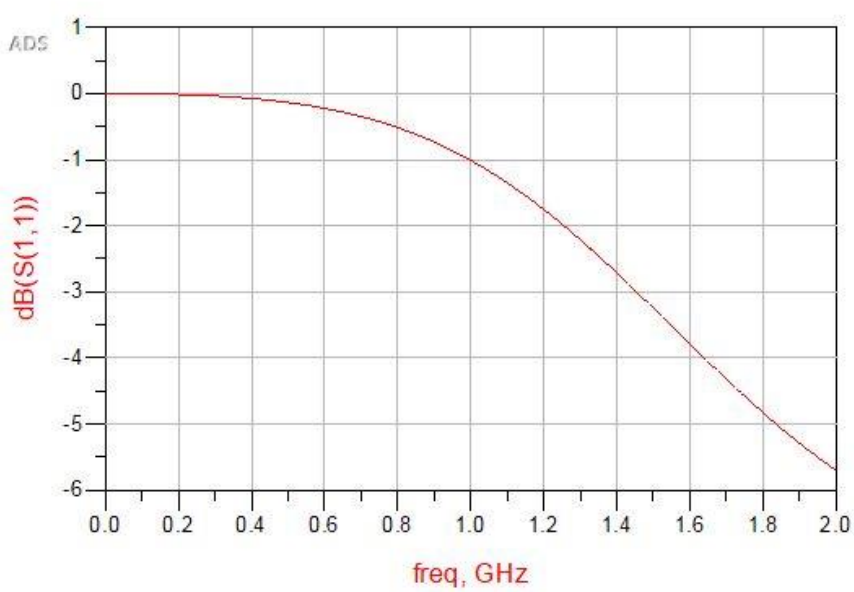


Figure 5.10: Tunable filter frequency response with a 5V tuning voltage

Figure 5.11: Tunable filter frequency response with a 10V tuning voltage

6 Limitations of Using Non-Foster Circuits for Antenna Matching

Non-Foster matching networks offer distinct advantages for matching electrically small antennas. Simply put, using conventional matching networks to match electrically small antennas will restrict operation to a narrow bandwidth. The advantages and disadvantages of each wideband operation method have been discussed in Chapter 1 and Chapter 2. Some practical issues with implementing non-Foster networks for antenna matching have also been discussed in Chapter 5. The non-Foster matching network is not a single solution to solve all wideband matching issues. In addition to limitations with added noise, stability, and flexibility, there are some fundamental scenarios where a non-Foster network has not yet been utilized.

6.1 Receive Only Antennas

A transistor-based non-Foster network transmits signals using active transistors. These active transistors sample the antenna voltage and relay that to a load, which might be a receiver circuit. When supplied with external DC power for transistor bias voltages, this non-Foster network can inject additional power and noise into the received signal. Therefore, as a fundamental rule, the power of the signal is not constant from the input through the output of a non-Foster matching network. For received signals, often very low in power, this is not an issue. Conversely, a conventional matching network transmits power already provided by an antenna into a load, or vice versa. The conventional matching network doesn't inject any external power which didn't originate from an antenna.

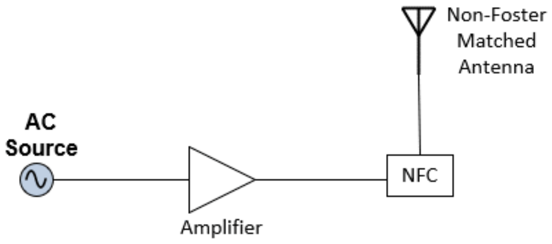
Recall that the transistor-based non-Foster circuit is composed of several emitter-followers, or "buffer amplifiers". Similarly, an operational amplifier based non-Foster circuit is also comprised of operational amplifiers in a buffer configuration. This is consistent with the capability to transfer power across different impedances over a wide bandwidth. As such, receiver circuitry is well suited for non-Foster matching network use.

On the other hand, non-Foster matching networks are not well suited for transmit circuitry. A transmitting antenna requires a well-matched input to allow the power to be transferred efficiently. The buffering nature of a non-Foster circuit which results in a non-constant power through the circuit is not suitable for transmit applications. A transmitter could include a power amplifier that determines the power input to the antenna. The transmitted power needs to be preserved for meeting performance requirements of each application. Any loss in transmit power reduces performance requirements. Further, if a non-Foster network is used in a transmitting application, the power level will not be preserved.

Constant power transmission is only one of the constraints restricting non-Foster circuits for receive only. Impedance matching characteristics play a role in practical applications of non-Foster matching networks. In a receive configuration, the non-Foster network impedance inversion characteristics allow the antenna to see a conjugately matched impedance. When the antenna is conjugately matched, the power is then buffered from the antenna, through the non-Foster network, and into the receiver. A carefully chosen “load impedance” which the non-Foster circuit uses to create an inverted impedance should match the impedance of the antenna to produce a wideband match. This way, the antenna sees a well-matched termination and the receiver sees the output impedance of the non-Foster circuit, a relatively constant impedance.

In a transmit configuration, the opposite must be true. Since the non-Foster matching network is designed to match an impedance at its input, the load impedance chosen for the impedance inversion should match the transmit circuitry. The output impedance of the non-Foster circuit is defined by the transistor properties and not by the carefully chosen “load impedance”. This impedance, whatever value it might be, should match the antenna impedance. This is more difficult, on both the input and output of the non-Foster circuit. Since the transmitter circuitry is active, finding a passive “load impedance” which represents a perfect inversion of the active transmitter is unlikely. Also, transmitter circuitry is optimized during the design stage for having zero reactance which diminishes the potential improvement from non-Foster circuits. On the output side of the non-Foster network, the transistor parameters which determine output impedance would not likely conjugately-match an electrically small antenna impedance. Therefore, transmit applications result in poor termination on the transmitter and on the antenna, in addition to the transmitter power being reduced because of the buffering nature.

Lastly, it should be noted that this deficiency of non-Foster matching networks, being unsuitable for transmit applications, has been well understood and is an area of active research [16, 60]. The research includes test scenarios where antenna and non-Foster network combinations that are excited with higher power to simulate transmit applications. Things such as the input match (return loss), linearity (IP3/third-order intercept point), and gain measurements are provided. Reference [16] shows some laboratory experiments with exciting non-Foster networks in the transmit direction. Comparisons were carried out between a normal antenna in transmit configuration and an antenna terminated with a non-Foster matching network, shown in Figure 6.1 and Figure 6.2. This work simply introduces the problem to show that using non-Foster networks for transmit applications comes with other challenges. There are no solutions to successfully achieving improvement with non-Foster network transmitters in references [16,60].



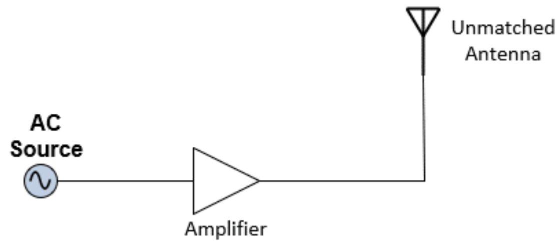


Figure 6.1: Diagram detailing measurements taken of non-Foster antenna in transmit configuration, reproduced originally from [16]

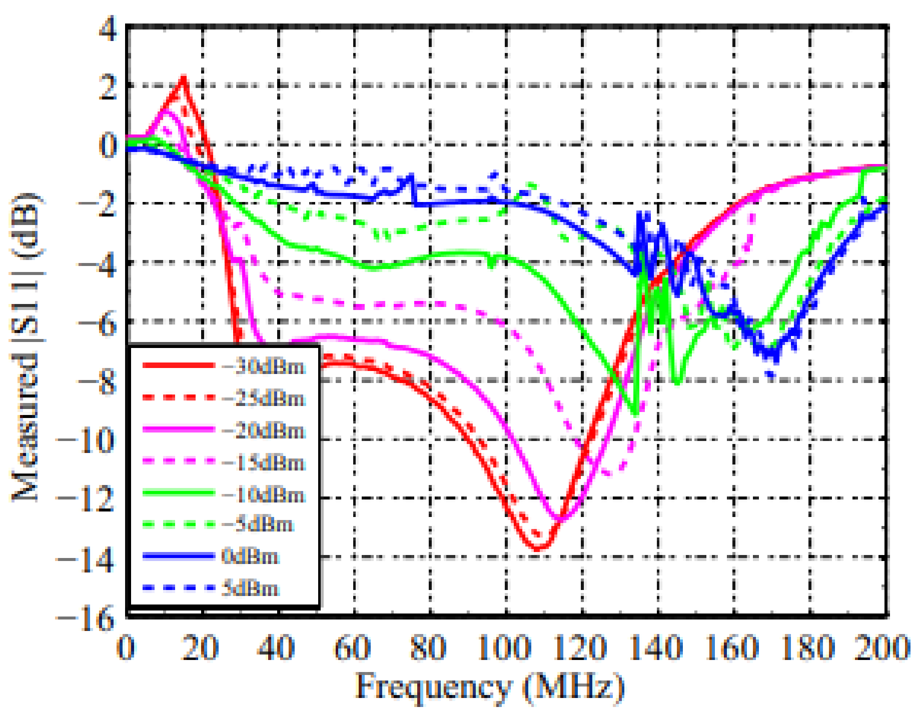


Figure 6.2: Measured return loss (S11) of a non-Foster antenna matched combination in transmit with varying input power, reprinted from [16]

6.2 Uni-Directional Nature

Conventional matching networks are passive, bi-directional devices. One of the important properties of an antenna is reciprocity. This allows the same antenna to be used simultaneously for transmit and receive applications. Additionally, being a reciprocal device, an antenna will have identical efficiency, gain, and beam pattern characteristics for both transmit and receive scenarios. Several different devices can be used to control the directionality of the signal, such as a circulator or a transmit-receive switch. In either of these cases, a conventional matching network would be placed on the common side of this network and attached to the antenna; to be used in both the transmit and receive directions. This is illustrated in Figure 6.3.

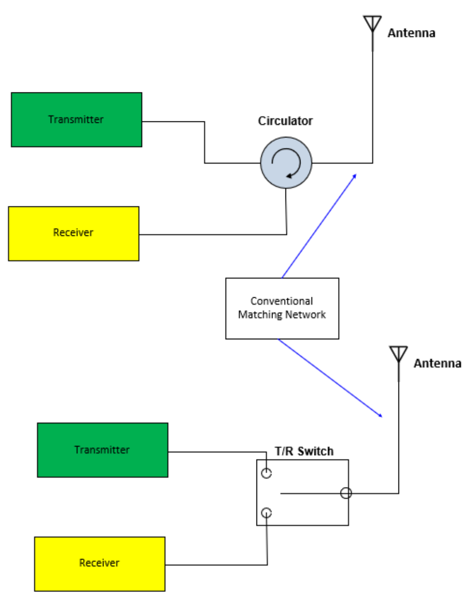


Figure 6.3: Antenna utilized for both transmit and receive using a) T/R switch and b) circulator, reproduced and modified originally from [57]

The combination of an antenna and conventional matching network can be grouped as a package. Both components are bi-directional and both can receive and transmit signals. A non-Foster circuit, however, can only be used in one direction. Despite the active research mentioned in Section 6.1, if a suitable transmit non-Foster network were to be implemented as well, two would be required for each transmit-receive system. The single directionality of non-Foster networks then integrates the non-Foster networks into the transmit-receive circuitry rather than the antenna. It is easy to see the reason for the single directionality. Once again, active circuits bring many limitations, and one of the fundamental differences between active circuits and passive copper-based circuits is the lack of reciprocity. For passive copper-based circuits, they are of symmetric nature, meaning that the transmission coefficients, S_{21} and S_{12}, are equal. In fact, this holds for all passive circuits of isotropic materials. There are some passive circuits with non-reciprocal materials that do not hold reciprocity, such as the circulator shown in Figure 6.3. Just like transistors, amplifiers, and many other active devices, there are specified input and output ports which cannot be reversed.

6.3 Power Limitations

In the receive only application of a non-Foster circuit, the typical received power on an antenna is very low. Free space path loss dominates signal power resulting in low power incident signals on an antenna. There are two cases where this assumption breaks down. The first case is for transmitters which have very high power and very high antenna gain. Despite being placed far away, the received power at an antenna could still become an issue. The second case is for transmitters that are very close to an antenna, almost in coexistence. This could be unintended leakage or interference between two systems. Also, an example could be for a transmit-receive system that transmits a signal which is reflected by a very close object, such as a large reflector close in range.

Regardless, all these scenarios will result in the received antenna signal being higher in power. There is a limitation on the maximum power a non-Foster circuit can accommodate. A conventional matching network will typically dissipate heat due to the losses when presented with a high power. The transistor-based non-Foster network will instead be pushed into saturation. This is a function of the maximum voltage incident on the transistor compared to the transistor DC bias voltage. The output of the transistor cannot exceed the DC bias voltage, and if the input drives the transistor in such a way, this circuit becomes saturated. The resulting output of the non-Foster circuit will be distorted. The frequency characteristics of the original signal will not be preserved as waveform distortion has higher frequency effects. Figure 6.4 shows the frequency components of a sine wave as it gets clipped; an identical effect to the output of a non-Foster circuit as the input power increases.

The previous discussion on the receive-only characteristics of non-Foster networks applies as well. If future research gives the possibility of using non-Foster networks for transmit applications, the power handling of becomes a significant issue. In transmit-receive systems, transmitters are increased in power to increase SNR. Once the signal is received, the SNR cannot be improved. As a result, the increased transmit power will improve received SNR. However, increasing transmit power requires higher power handling of all transmit circuitry, including a proposed non-Foster network.

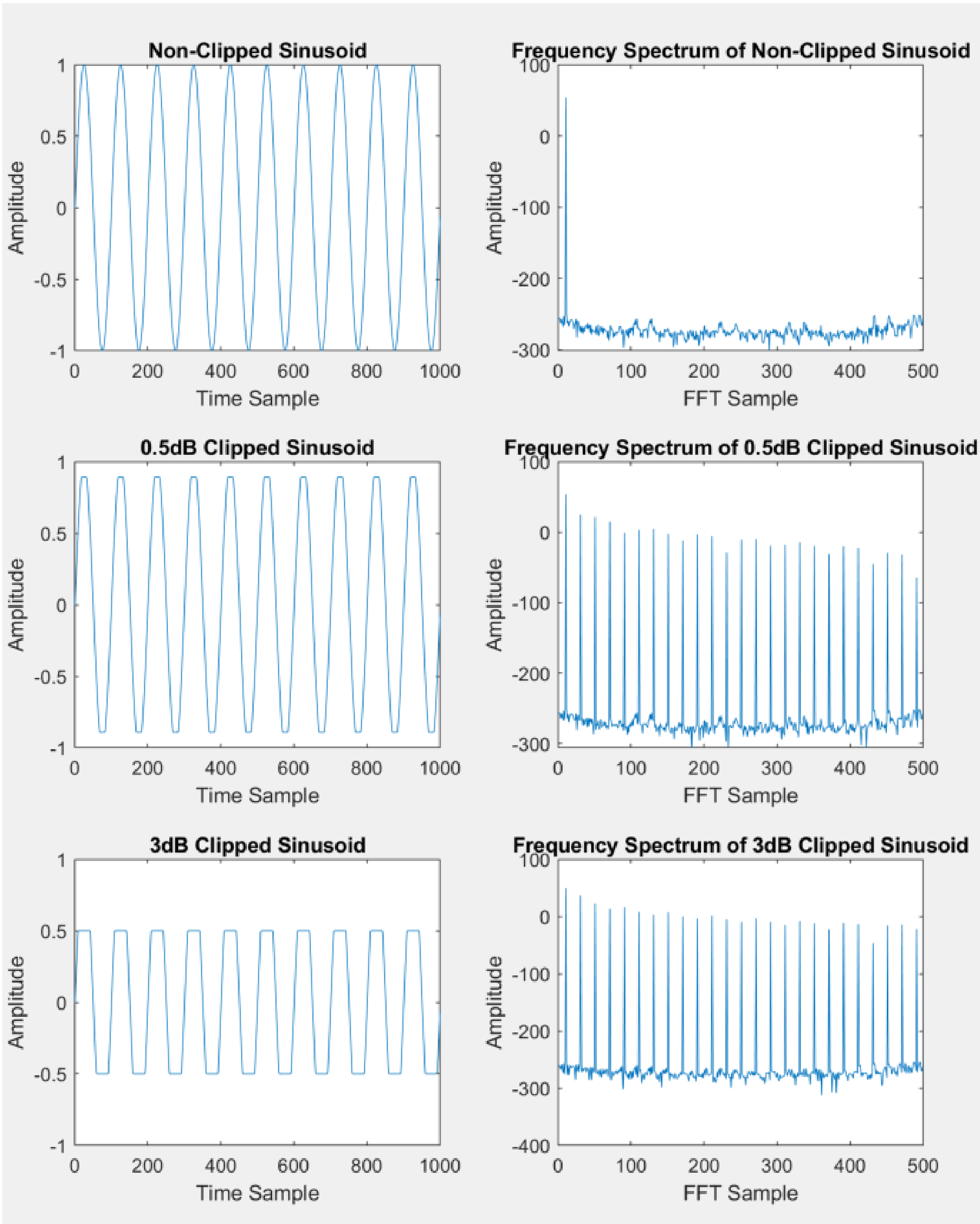


Figure 6.4: Clipping effects of a sinusoid, reproduced from [58]

6.4 Integration Concerns with utilizing NFCs for Antenna Matching

The simulations discussed in Chapter 4 and Chapter 5 have all been focused on a single device. The antenna was simulated alone, assuming a perfectly matched generator. Similarly, the non-Foster network was simulated with 50-Ohm characteristic impedance terminations on each port. The physical implementation of the three components, the antenna, the non-Foster matching network, and the receiver circuitry (network analyzer in this case) has brought to light unforeseen integration concerns. The impedance differences were a known concern, and the previous discussion addresses how the impedance was optimized for the best results. One unforeseen issue is the nature of the antenna feed and the non-Foster network termination. By design, a dipole antenna has a balanced feed. A balanced feed mandates equal and opposite current along the two conductors on the dipole antenna. To achieve a balanced feed, like the structure of a dipole antenna, the two conductors should be of the same geometry, size, length, etc. There are several types of well-known balanced feeds, such as a ladder line, among others. A cylindrical coax cable, for example, is not a balanced feed since the outer conductor has more surface area than the inner conductor. Figure 6.5 shows an example of a do-it-yourself (DIY) ladder-line feed from a HAM radio website. From a visual inspection, the geometry of the ladder-line feed forces symmetry between both conductors and controls the forward and return current paths very well.

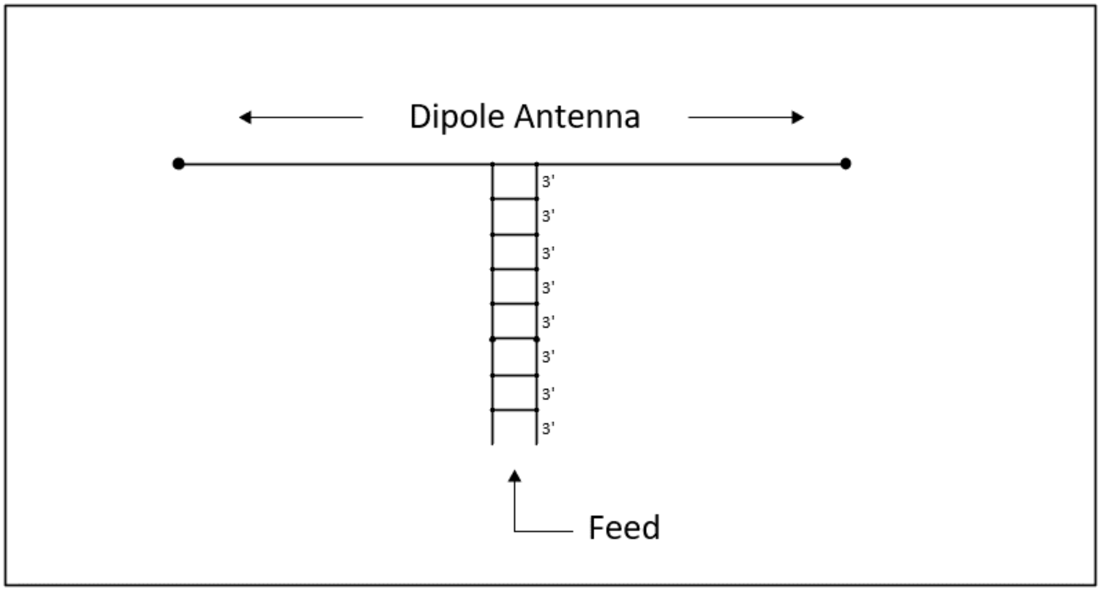


Figure 6.5: Example of a DIY ladder line from a ham radio website, originally reproduced from [59]

An additional consideration of the feed structure is its characteristic impedance. In general, commercially available coax cables come in two forms, 75-Ohm characteristic impedance for TV feeds and 50-Ohm characteristic impedance variants for any RF circuitry. A ladder line or other balanced feed structures do not necessarily need to conform to one of these impedances. As many antennas (including this one) have a generally higher impedance than 50 or 75 ohms, this becomes another point of contention on choosing the proper feed.

The non-Foster network is unbalanced on both ends. Since the ground conductor is also the return path, this results in an unequal surface between the two conductors and the surface currents may not be equal and opposite with a difference current flowing elsewhere on the circuitry. There have been alterations to Linvill's original transistor-based non-Foster circuit to create both balanced and unbalanced versions [60]. However, not all antennas are a balanced configuration. Some antennas, such as monopole antennas, are unbalanced and should be fed accordingly. The resulting connection between a balanced antenna and an unbalanced matching network causes significant issues. It should also be noted that the receiver circuitry side is typically fed by an unbalanced coaxial connector at 50-Ohm characteristic impedance. One of the non-Foster matching network's chief functions is to impedance match the antenna to the receiver.. However, typical matching networks, non-Foster or otherwise, focus on the reactive portion of the impedance. The resistance portion is rarely changed with a device other than an impedance transformer.

If there is an improper transition between a balanced and a non-balanced feed, the feed lines are likely to act as radiators themselves due to unequal currents. This, of course, disorts the antenna radiation pattern as well as compromising impedance match. Controlling line lengths, in terms of wavelength, can be an effective method for minimizing feed radiation. If the feed line is radiating but with a very small electrical length, the radiation could be negligible and that might be a satisfactory solution. Otherwise, a "balun" is typically called for. A transformer balun, for example, contains inductive coils that transfer electrical energy through a magnetic field. They separate the forward and return currents from both ends of the balun making them a great option for antenna feeds. They are band-limited, however. It is difficult to find a balun that operates across the bandwidth for the proposed antenna and non-Foster combination discussed here.

A transformer balun can act both as a balancing feed mechanism and an impedance transformer if the number of coils on each end of the balun are chosen properly. This is also helpful for feeding balanced antennas that have a different impedance. For our purposes, the antenna has a much higher impedance (resistance) than our non-Foster circuit input. However, that part was not as critical due to the nature of the non-Foster circuit. Since the non-Foster circuit input feeds transistors which sample the voltage, the impedance of the input is only important to ensure stability. However, as mentioned in Chapter 4, there are additional transistor stages added to the core non-Foster circuit to force the impedance to guarantee stability. In this circumstance, the impedance transformation on the antenna side is not needed. However, on the receiver side, the output impedance of the non-Foster circuit is much lower than that of the characteristic impedance of a typical 50-Ohm receiver. The output impedance of the non-Foster circuit is around 10 Ohms without any additional components. One thing to note: In general, when a transformer balun includes an impedance transformation, the bandwidth is further reduced proportional to the ratio of the primary and secondary windings. This is apparent by a quick examination of the datasheets of available parts and their bandwidth.

Additional issues present themselves with the use of a balun on the antenna input to the non-Foster network and an impedance transformer on the receiver output. Without the impedance transformer, the non-Foster network will reduce the reactance of the resistively loaded dipole antenna. However, the resistance is much lower than 50 Ohms. When the impedance transformer is added to the non-Foster network, the real part of the impedance is more closely matched to 50 ohms, however the reactance increases due to the transformer and naturally the bandwidth is limited. It becomes a balancing act between matching the real parts and imaginary parts of the impedance (by choosing an impedance transformation value).

The use of a balun or impedance transformer forces the separation of the forward and return current paths. Since the power coupling is done by a magnetic field, there is no (RF) electrical connection between the two ports on a balun or impedance transformer. Theoretically, this doesn't present any problems and should result in a better impedance transformation. However, in practical circuits, there will always be connections that give other paths for the currents to flow. For example, a power supply that is connected to the non-Foster circuit which powers the transistor bias voltages will share a common earth ground with that of the receiver circuitry power supplies. Ideally, these are only DC ground paths that are isolated from RF, however at the very low frequencies there is less isolation between RF and DC currents. Furthermore, this becomes an especially troublesome issue when other RF circuitry is added which shares RF grounds. For example, a low noise amplifier (LNA) is quite common to be one of the first components in a RF receiver. An LNA will likely share the RF ground and the DC ground which could cause ground loop issues, i.e., the ground path for the return signal is not explicitly confined to a single path and could cause the feed radiation issue as discussed earlier. In the case of a "bypass" implementation, which would allow the non-Foster matching network to be bypassed at high frequencies where the non-Foster network becomes less effective (although it will continue to add noise), a pair of switches are required. One switch would be inserted before the non-Foster network and the other after. This would allow the user to choose the path of the matching network or a direct antenna path. A circuit like this would also present problems since the grounds across the circuit then are not connected and the switches would not have a common RF ground reference. Much of this application could continue to be refined and offers opportunity for future work.

7 Conclusion

In Chapter 1, the problem of systems utilizing a large bandwidth and simultaneously operating at low frequencies was introduced. These wideband systems often require the minimum operating frequency to be as low as 20MHz. Furthermore, with the trending miniaturization of electronic devices, the use of electrically small antennas has become a necessity. The quest to further reduce size has pushed designers to combine many receivers into a single device, thus driving the need for multi-octave bandwidth. Creating a matching network for three simultaneous challenges, those being the combination of low frequency operation, with a multi-octave bandwidth and utilizing an electrically small antenna, is a very difficult design challenge. Conventional solutions have been successful at mitigating two out of the three, e.g., low frequency operation and multi-octave bandwidth but without an electrically small antenna. However, conventional solutions fall short when addressing the three constraints together. Detail was provided on bandwidth limitations of conventional matching networks and how they compare with active matching networks.

Active matching networks, such as the non-Foster type were proposed as a solution to this problem. A description of what classifies a non-Foster network was given and some of the advantages and setbacks of non-Foster networks were portrayed. A comparison between non-Foster matching networks and conventional matching networks was presented demonstrating the potential benefit of non-Foster networks.

Next, a literature search was provided, which discussed some of the major contributors to non-Foster matching network research for antennas. Different implementations of non-Foster matching networks were discussed, such as the transistor-based versions, operational-amplifier versions, and even some less-common variants such as those using metamaterials. Papers of academic research as well as industrial implementation were examined. The subject of tunability with non-Foster networks was researched and the work of several previous authors and their implementations of tunable non-Foster networks was presented.

The discussion on wideband matching was continued in Chapter 2. The proposed solution of non-Foster matching networks is just one way to solve this problem. They bring unique advantages particularly as regards satisfying the three aforementioned constraints (wideband, low frequency, ESA), but with their advantages come disadvantages that were identified. Other methods for wideband operation were discussed, with comparisons to the non-Foster method. To be clear, the non-Foster method is not a perfect solution to the antenna matching problem. Also, it should be noted that non-Foster networks have other applications outside of antenna matching. For this dissertation, however, the focus was on antenna matching although, for completeness, other uses were discussed, too. Wideband matching is a problem that has existed for many years with many proposed solutions, each with their specific strengths. A few of these solutions were identified including varying antenna geometry, applying matched LNA networks, and varying characteristic impedance. Each was discussed in detail and a comparison table was presented which compares non-Foster matching networks to these alternate solutions for wideband operation.

In Chapter 3, negative impedance inverter circuits were discussed. This category of non-Foster circuits includes circuits that are specifically tailored for antenna matching in that they mimic negative impedance of an antenna. The combination of the negative impedance with the antenna load provides a conjugate match that may enhance wideband operation. The discussion began with Linvill's classic transistor-based version of a negative impedance inverter circuit. The inner workings of the circuit were discussed, and some simulations were provided using updated circuit components. Initially, the classic transistor-based non-Foster circuit provides a reactance curve that closely matches an ideal inverted reactance. Much of the content which followed refers to the transistor-based non-Foster circuit that was modeled after this original implementation, but with modifications.

Next, we presented the non-Foster networks implemented with operational amplifiers. While the operational amplifiers are indeed just a combination of many transistors, this is still a different class of non-Foster circuits. A detailed explanation on the theory of operation was given and some previous work focused on these circuits was referenced. Data showing the performance of the operational amplifier versions was provided. Because of their close relationship with the transistor-based non-Foster circuits, some key advantages and disadvantages of each method were discussed.

Impedance tunable non-Foster circuits are an extension of the standard non-tunable version. The tunability allows the observed impedance to be changed without the need for circuit modification. This opens the door for many applications and provides extensive flexibility to match different kinds of antennas and for antennas in different environments. Some of the advantages to using a tunable non-Foster circuit were given. Additionally, we looked at some of the previously introduced tunable non-Foster circuits in the literature. Over the years, several different implementations of tunability have been published due to the realized advantages of modifying the resulting reactance. Some methods are simpler than others, but each come with their pros and cons. A comparison table was given which details some of the key considerations relating to each tuning method.

Chapter 4 introduced a novel method for impedance tuning transistor-based non-Foster circuits. Starting with Linvill's classic circuit, a tuning transistor is applied in series with the load, and that transistor can be adjusted rapidly in quickly modifying the impedance. A detailed explanation on how this additional tuning transistor creates a changing observed reactance was provided. Simulations demonstrated the range of tunability and we compared the reactance curves to one of the other tuning methods found previously in the literature.

An in-depth discussion on circuit stability followed the simulations. Active circuits are known to become unstable, and a proper analysis of stability is critical. The potential for instability is one of the leading limitations of non-Foster circuits. The state of unconditional stability is desired and, although not observed directly, can at least be calculated. Some of the historical stability methods are based on assumptions commonly misunderstood by circuit designers. Observation of a stable circuit does not indicate unconditional stability. It may be sufficient that a designer's circuit is stable under the conditions of which it is subjected, but it may become unstable under other conditions. Thus, rather than continuing with Linvill's classic circuit, which has the potential to be unstable, the circuit of Harris and Myers was used as a baseline for demonstrating the value of the added transistor. That circuit was expected to be stable for the purposes of matching an electrically small antenna. While Harris and Myers did not provide a stability analysis with their original publication, they observed stable operation when loaded with an antenna.

The unique concept of the added transistor for impedance tuning was applied to a modified version of Harris and Myer's circuit of which tunability was observed. A rigorous detailed stability analysis was shown to indicate stability under all conditions. The stability method used is based on the normalized determinant function and shown to relieve common assumptions on unloaded stability. Simulated and measured results were presented.

When compared to Linvill's original circuit simulations, the performance suffered considerably. As discussed, there is a trade-off between unconditional stability and maximum performance. An optimization study would be an area for future work.

Chapter 5 focuses on the practical applications of the tunable non-Foster networks. Most of the applications given center on antenna impedance matching. Uses for tunable non-Foster networks are not limited to antenna matching, however. Different applications including tunable filters with negative group delay, and phase shifters were also discussed. The advantages of tunability come with realizability concerns as well. Those were discussed in detail and resulted from the effort of physically implementing and building a non-Foster matched antenna. A noise analysis also was given which shows an increased noise contribution from the active transistors but fortunately the SNR gains were demonstrated to be significant.

Finally, Chapter 6 discussed some of the limitations of non-Foster circuits. Not limited to tunable circuits or even transistor-based circuits, there are areas where non-Foster circuits are not suited for use. However, one of the primary examples, using non-Foster networks for transmit applications, while never seen before, is an active area of research. It's important to note that things which seem unpractical today can always be challenged with creative thinkers and innovative engineers in the future.

7.1 Future Work

Some aspects of the work given here remain open ended. There are many potential areas where future work can be performed to overcome some of the obstacles experienced. The non-Foster circuit that was implemented was based on Harris and Myer's version of the transistor-based impedance inverter. This circuit was designed around a specific antenna. Fortunately, it provided a good test bed and proved that the method of adding a transistor for impedance tunability is applicable over different circuit implementations. However, an area for improvement could be to relax some of the constraints needed for unconditional stability.

Another area for potential growth is a hybrid implementation of the non-Foster matching network. As discussed in several chapters, the main positive contribution to the non-Foster matching network is at the low frequencies, but performance could suffer elsewhere. A quick attempt at implementing a hybrid method was made by bypassing the non-Foster network with a silicon based MMIC switch. This intent was to allow the positive contributions at the lower frequencies to be capitalized and at the same time foregoing the added noise at higher frequencies. Unfortunately, the common signal ground is shared by the two switches used to bypass the non-Foster network which caused problems with the balanced termination. There are several possible solutions to mitigate these problems, and an investigation on some of the considerations of the different methods could bring further utility out of a tunable non-Foster matching network.

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