Tunable Combline Filter and Balun: Design, Simulation, and Test

Daniel Alex Ramirez
University of South Florida, ramirez13@mail.usf.edu

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Tunable Combline Filter and Balun:
Design, Simulation, and Test

by

Daniel Alex Ramirez

A thesis submitted in partial fulfillment
of the requirements for the degree of
Master of Science in Electrical Engineering
Department of Electrical Engineering
College of Engineering
University of South Florida

Major Professor: Gokhan Mumcu, Ph.D.
Thomas Weller, Ph.D.
Jing Wang, Ph.D.

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DEDICATION

I would like to first dedicate this work to my loving wife who has supported me through this process. Without her continued understanding this would not have been possible. Also to my parents who taught me the values of hard work, dedication, and perseverance. To my brother who encourages and motivates me to push through when times get difficult. And to my sisters who always believe in me.

I would also like to dedicate this work to Dale Leppert; the first engineer who believed in me and inspired me to head down this path.
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ABSTRACT

Reconfigurable filters are an attractive solution for many military and commercial applications due to their ability to alter the partitioned frequency band in an RF system without requiring a bank of fixed filters. The onset of this technology has the potential to revolutionize the RF industry by allowing for agile devices which consume less size, weight, and power while providing greater performance. However, at the present state, reconfigurable filters present a reduction in performance when compared to filter banks. This has led to exciting research in the field of RF tunable filters.

For many applications, planar reconfigurable filters have been utilized due to their low cost of manufacturing and ease of implementation in a system. One topology that has proven to be versatile in design is the combline filter which employs line resonators loaded with a capacitor to obtain a predetermined response. To implement a varying center frequency reconfigurable combline filter, the resonator is loaded with a capacitance that can be tuned either digitally or continuously by presenting a DC voltage. Due to their ease of use and availability, varactors are a common choice as they provide a continuously tunable capacitance by presenting a reverse bias voltage to the device.

To continue the trend of lowering size, weight and power while maintaining high vitality in performance, consolidation of RF components may prove to be a good next step. Tunable balun filters have already been presented as a viable option for consolidation of components and show good performance. However, those designs which have been presented do not demonstrate a
topology that can implement higher than a second order filter. This project, for the first time, investigates the consolidation of the Marchand balun and the combline filter into a single topology which allows for quick adaption of higher order filters while maintaining vigor in performance. A design is presented which achieves 25% tuning bandwidth centered at 1.5 GHz with less than 5 dB insertion loss, a phase balance of 180 ± 1° and an amplitude difference of ± 0.6 dB.
CHAPTER 1
INTRODUCTION

In today’s digital age, more data is being transmitted than ever before. Requirements for information and functions on a single device continue to expand while size, weight, and power consumption (SWaP) become more restrictive. This has led to extensive research in the field of reconfigurable electronics in RF systems over the past few decades.

There are many different tunable components within a reconfigurable front end of a wireless system that need to provide performance to accommodate the growing spectral densities. One of these components, which helps to partition frequencies in an RF front end, is the filter. Tunable filters are of utmost importance for future devices and require high Q, agility, and reconfigurable responses from minimal SWaP.

Before the advent of electrically tunable capacitance, all reconfigurable filters were accomplished through mechanical components such as tuning screws ever since the development of radar. Although this type of tuning provides high power handling with accuracy and works well for verification and test, this manual approach is slow taking seconds to tune and usually physically large making them a less desirable option for wireless handsets [1]. In as early as 1979 [2], the diode known as the varactor was implemented in a tunable filter and reconfiguration became more electric rather than mechanical. However, due to the resistance presented by the depletion region of a varactor, the Q of these devices is usually around 30-40 and yields an undesirable amount of loss when loading a resonator type bandpass filter, especially at frequencies greater than 1 GHz.
Although there are drawbacks in the amount of loss a varactor presents as a series resistance, varactors are fast, can be tuned in a matter of nanoseconds, and are simple to implement on planar filter types.

A more recent tuning technology that has presented itself is that of Microelectromechanical Systems (MEMS). Within the field of MEMS, there are also subcategories of tuning technologies that range from actuator type capacitance variance to switches which then digitally tune to different capacitance values. When MEMS first presented themselves as a viable tuning technology, there was an excitement in the RF as well as other electronics communities. MEMS devices support areas spanning from inkjet printers to barometric sensors. Though the field of MEMS has excelled, the use of MEMS in reconfigurable RF systems has not yet presented itself completely as a viable option to tunable filters. The loss presented by the series resistance of a MEMS device is in the 5 \( \Omega \) range [4] and still dominates the total attenuation presented by the tunable filter. The high voltage needed to control the MEMS devices, which can be as high as 100 V, also provide a level of complexity to the design of a MEMS based tunable filter as it would require a voltage level translator to be placed into the circuit. In addition, the cost to fabricate MEMS devices is high and their Mean Time to Failure (MTTF) can be as low as 44.4 million cycles when hot switching [5].

Within the last five years, there have been a vast amount of research in tunable filters utilizing MEMS [6], varactors [7], and even more elaborate tuning elements such as microfluidics [8]. In addition to these components receiving attention, many filter topologies have been attempted including combline type structures [6], hairpin resonators [7], open loop resonators [4][9], folded resonators [10], mixed resonator types [11], substrate integrated filters [12] and many more. There has also been a focus on synthesis techniques to achieve optimum filter responses [13] and [14]. These tunable filter’s show great performance between 1-2 GHz with
insertion loss from 1 dB to 5 dB at fractional bandwidths of 5% to 7%. In [4], the author demonstrates that controlling the magnetic and electric coupling of a two-pole open loop resonator type structure can yield increasing fractional bandwidth, constant fractional bandwidth, and decreasing fractional bandwidth without having to vary the structures physical size. However, this design is inherently narrowband and does not present a straightforward way to introduce higher order filters. Combline filter structures have been shown to be simple to manipulate and scale to achieve different tunable filter responses and orders [15]. This flexibility has allowed for the tuning of bandwidth in addition to the center frequency of the filter response. However, even with current state of the art tuning elements, adding more poles introduces more tuning elements to the structure which degrades insertion loss and overall filter shape.

In addition to the RF filter, a balun is also a critical component in the development of wireless systems and radars [16]. Baluns are devices which inherently convert a conventional unbalanced RF signal (e.g. 50 Ohm microstrip line on a PWB) to a balanced line (e.g. 100 Ohm dipole antenna feed). These type of devices are used in such applications as differential amplifiers, balanced mixers, frequency multipliers, and antenna feed networks [17] as well as push pull amplifiers. There are many different balun implementations with one of the more popular ways being that of the planar balun [16], [17-22] due to the ease and low cost of fabrication. Planar baluns are also passive and so inherently do not require any power. Within the category of planar baluns, the Marchand balun has proven to be a common balun of choice due to its wideband characteristics and ease of implementation.

In order to continue the trend of compacting RF and microwave circuitry, an effort must be made to consolidate components to achieve similar performance in a fraction of the physical space. This undertaking has already been taking place in the combination of antennas and baluns.
into a single design [21]. However, there are other balun applications mentioned besides that of an antenna feed and many of those applications require filtering of the signal. Integrating a tunable filter and balun structure will occupy less space compared to a reconfigurable filter and balun in cascade and provides similar performance. Such a structure can be used to minimize board real-estate in different wireless applications.

Tunable balun filters have been presented previously and have shown promising results [17], [23-24]. To the author’s knowledge, all published tunable balun filters to this point utilize a ring and open loop resonator structure type. Although good performance was achieved, future work of scaling to higher order filters beyond the 1 or 2-pole design examples may prove to be difficult. Here we present the first tunable balun circuit utilizing a combline structure. In addition to achieving superior phase balance compared to the previous listed work, this topology also allows for scaling to higher order filters.

In Chapter 2, we will have an overview of RF tunable filter and balun design principles and will focus on the combline filter and Marchand balun which are utilized in this Master’s Thesis. We will conclude this chapter by discussing the varactor diode equivalent circuit model and performance impacts which was used to tune the structures presented herein. Chapter 3 will show the design, fabrication, and testing of a tunable combline bandpass filter with a 10% fractional bandwidth that is centered at 1.5 GHz. This frequency is was chosen at L-band which is commonly used for communications purposes and fulfilled a joint effort with Harris Corporation who funded the research. Finally, utilizing the combline structure shown in Chapter 4, we will discuss a novel design which combines the well-known combline filter and Marchand balun to provide a further step in the effort to provide more functionalities in a smaller form.
CHAPTER 2
BACKGROUND AND THEORY

Research in the field of RF filters has spanned a vast amount of topics and has grown increasingly popular with time. It is important to note some of the key background information so that the reader has the appropriate understanding and appreciation for the topics which will be discussed herein. Therefore, in this background review, we will focus on the building blocks which make up the tunable filter balun whose design will be shown later on. Specifically, we will focus on planar filter and balun design principles, combline filters, Marchand baluns, and varactors.

2.1 Planar Filter and Balun Design Principles

Planar structures are attractive to designers and researchers as they are readily fabricated utilizing simple etching processes and can be integrated easily into a printed circuit board (PCB). There are many types of planar filter implementations which can be utilized to achieve many different responses. The four different responses which a designer may try to achieve are the low-pass, band-pass, high-pass, and band-stop (or notch) filter response. For our discussion, we will be focusing on the band-pass type of response.

Within planar band-pass filter topologies, there are many implementations which can result in the required response. Some of the most common microstrip examples are the open loop resonators, the interdigitated, and the combline type bandpass filters which can be seen in Figure 1 at a top view. Each of these planar structures rely on resonators to accomplish the band-pass type
response. The desired resonance occurs at the frequency at which a resonator’s electrical length is \( \lambda/2 \) for the open ended loop resonator type filter and \( \lambda/4 \) for the interdigitated and combline resonators, where \( \lambda \) is the guided wavelength in the medium of propagation [15]. For the design specified herein, we will be utilizing the combline structure which will be explained in more details in section 2.2.

![Common Types of Bandpass Artwork](image)

**Figure 1: Common Types of Bandpass Artwork**

Likewise, there are many different balun approaches [16], [18-19] and even approaches which integrate the balun design into the design of the antenna [20-21]. But unlike filters, there are much less conventional planar balun topologies to choose from. However, one balun that has remained a popular choice in research has been the Marchand balun (see Figure 2). This balun, which was first presented by Nathan Marchand in 1944 [22], continues to receive attention and is relevant even today as we progress to millimeter wave and terahertz applications [25-27]. The Marchand balun is a reference of choice due to its overall ease of design, small size, and wideband performance.
2.1.1 Low-pass Prototypes and Responses

The design method used for this filter design is known as the so called insertion loss method [28]. This method allows a designer to select the transfer function response and have control over the final shape of the passband and stopband at the beginning of the design.

There are many different mathematical transfer functions one can choose for their response and can even uniquely define a function based on their specified needs. However, there are three functions which have been utilized and are quite popular among RF filter engineers. These functions or responses are known as the Butterworth, the Chebyshev, and the Elliptic responses. Each function has its own benefits and drawbacks and it is therefore up to the designer to understand these to move forward intelligently. For example, the Butterworth filter has the least amount of ripple, but also has the smallest slope in the rejection band. The Chebyshev filter has a fast roll off (i.e. a large slope in the rejection band), however, in order to achieve this fast roll off, the topology introduces ripple in the passband. The last response is the Elliptic response. This response has a superior roll off rate compared to the Chebyshev response, but has less rejection in the rejection band. A graphical representation of their responses is shown below in Figure 3.
Figure 3: Theoretical Responses of the Butterworth, Chebyshev, and Elliptic Low-Pass Filter

Once a transfer function response is selected, a designer can then employ the low-pass prototype, see Figure 4. The low-pass prototype can be defined as a filter whose source impedance has been normalized to equal unity and has a cut-off angular frequency which is equal to unity [15]. Each transfer function provides a set of so called “g-values” which define the normalized element value of the low-pass prototype. These g-values have been tabulated but and are derived from the aforementioned transfer functions. They can then be utilized in planar resonator type filters to find the coupling coefficients as well as the external quality factors which will be utilized in the design of the combline bandpass filter later.
After a transfer function is chosen, the $g$-values of the lowpass prototype can then be transformed to realizable element values to achieve lowpass, highpass, bandpass, or bandstop responses. For the presentation herein, we will be discussing a bandpass type response and so the transformation of elements from the lowpass prototype are shown in Figure 5. Where $\Omega_C$ represents the cutoff frequency, which is normalized to unity, $\omega_0$ is the center frequency of the now bandpass filter, and $\gamma_0$ is the normalized system admittance.

\[ c_p = \left( \frac{\Omega_C}{FBW \omega_0} \right) g \]
\[ l_p = \frac{1}{\omega_0^2 c_p} \]
\[ l_s = \left( \frac{\Omega_C}{FBW \omega_0} \right) \gamma_0 g \]
\[ c_s = \frac{1}{\omega_0^2 l_s} \]

Figure 5: Lowpass Prototype to Bandpass Transformation [15]
2.1.2 Quality Factor and Coupling Coefficient

The coupling coefficient and the quality factor bridge the gap between mathematical lumped element type filters and distributed element filters. The external quality factor ($Q_e$) represents the strength of coupling which is loading the first and last resonators. The stronger the coupling, the lower the external quality factor value is, resulting in a wider fractional bandwidth. The external quality factor relates to the lowpass prototype through the formula provided in [15].

$$Q_{e1} = \frac{g_0g_1}{FBW}, Q_{en} = \frac{g_ng_{n+1}}{FBW}$$

where $n$ is the number of elements in the lowpass prototype, FBW is the fractional bandwidth desired, and $g$ are the lowpass prototype value.

To adequately define the external quality factor, we must first look at the circuit model equivalent of a singly loaded resonator shown below in Figure 6 where $C_p$, $L_p$, and $G$ are the parallel capacitance, the parallel Inductance, and the source admittance, respectively. It should be noted here that the equivalent circuit shown in Figure 6 is identical to the first ($g_0$) and second ($g_1$) elements of the lowpass prototype shown in Figure 4 when $g_1$ is replaced with its bandpass transformation as shown in Figure 5. This shows the relationship between the external coupling coefficient and the lowpass prototype. From this we can find that

$$Q_e = \frac{\omega_0C_p}{G}$$

![Figure 6: Equivalent Circuit of Input/Output Resonator with Singly Loading [15]](image-url)
Now that the external coupling coefficient has been defined, a convenient way to extract the value from a physical model will be discussed. For this, the doubly loaded resonator whose equivalent circuit model is shown below in Figure 7 [15] is employed. This circuit is identical to that shown in Figure 6 only here the LC resonator has been split into two and there is an additional admittance $G_2$ loading port 2. This is where the term “doubly loaded” comes from and now presents a symmetrical network when $G = G_1 = G_2$. Here, a revised quality factor is defined as

$$Q'_e = \frac{\omega_0 C_p}{2G}$$

since the two admittances are in parallel and therefore add to each other. Because the external quality factor is defined as a singly loaded resonator, the loading effects of the second admittance needs to be removed out of the equation. This is easily done by multiplying the revised quality factor by two, such that the external quality factor is simply twice that of the revised quality factor.

![Figure 7: Equivalent Circuit of Input/Output Resonator Which is Doubly Loaded](image)

When a network is symmetrical, it can be divided through the plane of symmetry and evaluated using even and odd modes [15]. An even excitation causes the symmetrical interface to be open circuited and an odd mode excitation causes the symmetrical interface to be short circuited.
We can begin by evaluating the odd mode condition and assume there is a short at the plane of symmetry T-T’ in Figure 7. This provides an admittance and reflection at port 1 of

\[ Y_{ino} = \infty \]

\[ S_{11o} = \frac{G - Y_{ino}}{G + Y_{ino}} = -1 \]

Then we look at the condition where the plane of symmetry T-T’ presents an open such that

\[ Y_{ine} = \frac{j \omega_0 C_p \Delta \omega}{\omega_0} \]

\[ S_{11e} = \frac{G - Y_{ine}}{G + Y_{ine}} = \frac{1 - j Q_e \Delta \omega / \omega_0}{1 + j Q_e \Delta \omega / \omega_0} \]

where \( \omega_0 = 1/\sqrt{L_p C_p} \) and the approximation \( (\omega^2 - \omega_0^2) / \omega \approx 2 \Delta \omega \) with \( \omega = \omega_0 + \Delta \omega \) has been made [7]. We can then find \( S_{21} \) such that

\[ S_{21} = \frac{1}{2} (S_{11e} - S_{11o}) = \frac{1}{1 + j Q_e \Delta \omega / \omega_0} \]

\[ |S_{21}| = \frac{1}{\sqrt{1 + \left(\frac{Q_e \Delta \omega}{\omega_0}\right)^2}} \]

Now, at resonant frequency, \( \Delta \omega = 0 \) which means the magnitude of \( S_{21} \) is equal to unity. As the \( \Delta \omega \) increases or decreases, \( |S_{21}| \) decreases. This resonating response is represented in Figure 8. It is now possible to pick some magnitude of \( S_{21} \) which is less than unity and corresponds to some bandwidth \( \Delta \omega \). For convenience, a 3dB bandwidth is chosen such that \( |S_{21}| \) is equal to 0.707. We can now solve for \( Q_e' \)

\[ 0.707 = \frac{1}{\sqrt{1 + \left(\frac{Q_e \Delta \omega_{3dB}}{\omega_0}\right)^2}} \]
\[ Q'_e = \frac{\omega_0}{\Delta\omega_{3dB}} \]

However, we must multiply the above modified quality factor of the doubly loaded resonator by two to get the singly loaded external quality factor.

\[ Q_e = 2 \times Q'_e = \frac{2 \times \omega_0}{\Delta\omega_{3dB}} \]

Figure 8: Response of Resonant Amplitude for Figure 7

Alternatively, [15] has presented a method of extracting the external quality factor similarly to the doubly loaded resonator without having a plane of symmetry. This is achieved by having \( G_2 \) of the doubly loaded resonator shown in Figure 7 much less than \( G_1 \). It can be seen that as \( G_2 \) approaches zero, the equivalent circuit model of Figure 7 approaches a transformation into the singly loaded resonator of Figure 6. Also, as \( G_2 \) approaches zero, the total admittance \( G \approx G_1 \) and so the modified external quality factor approaches the unloaded quality factor. The response of such a circuit, is similar to that shown in Figure 8 only the maximum magnitude is far below
0 dB. However, since the loading of $G_2$ is almost negligible, this provides an easy way to extract an approximate value for the external quality factor.

To set up in simulation, a feed network presents the load to the resonator and is either tapped or coupled to the structure. Then, rather than having a second load presented to the resonator, a lightly coupled probe point is introduced to the structure, Figure 9. This probe point needs to be very weakly coupled to the resonator such that any loading affects presented by port 2 are negligible. As discussed, this method creates a similar response to that of Figure 8 but since the resonator is only singly loaded, the overall quality factor is simply

$$Q_e = \frac{\omega_0}{\Delta \omega_{3dB}}$$

However, if the probe point is too tightly coupled to the resonating line, then the resonating line is loaded by the probe point. This will of course effect the external quality extraction as the line is no longer singly loaded. However, this offset error should be close enough to the desired value such that optimization of the filter later will yield the desired result.

Given this definition of the external quality factor, one can set up a simulation in an Electromagnetic (EM) software tool where a probe point is lightly coupled to the resonator which is singly loaded. Then, by varying the tapping location “d” or the spacing between the feed line and the first resonator “s” (see Figure 9), one can extract the external quality factor.

The coupling coefficient $M$ represents the strength of coupling between resonators. The stronger the coupling, the larger the coupling coefficient value is, resulting in a wider fractional bandwidth. The coupling coefficient relates to the lowpass prototype through the formula provided in [15].

$$M_{i,i+1} = \frac{FBW}{\sqrt{g_i g_{i+1}}}$$
\[ for \ i = 1 \ to \ n - 1 \]

where \( n \) is the number elements in the lowpass prototype, \( FBW \) is the fractional bandwidth desired, and \( g \) are the lowpass prototype value. Once again, this value provides what the coupling coefficient will be in order to yield a chosen filter response based on the low-pass prototype selection.

Figure 9: External Quality Factor Extraction; Tapped Feed Line (a) and Coupled Feed Line (b)

For this discussion, we will only consider a synchronously tuned coupled-resonator circuit as we will ultimately be designing a filter whose resonators resonate at the same frequency. Therefore the assumption is that the self-inductance and self-capacitance of each resonator is equal. First, the electric coupling will be discussed and an equivalent lumped-element circuit model is shown below in Figure 10a in which \( L \) and \( C \) are the self-inductance and self-capacitance of two resonators such that \((LC)^{-1/2}\) is equal to \( \omega \) of the uncoupled resonator and \( C_m \) represents the mutual capacitance. The lumped element model is only valid at frequencies close to the resonant
frequency and therefore is narrow band. From the lumped equivalent model, the following set of equations can be defined [15].

\[ I_1 = j\omega CV_1 - j\omega C_m V_2 \]
\[ I_2 = j\omega CV_2 - j\omega C_m V_1 \]

Figure 10: (a) Synchronously Tuned Coupled Resonator Circuit with Electric Coupling and (b) Alternate form with J inverter

From there, we can find the four admittance parameters

\[ Y_{11} = Y_{22} = j\omega C \]
\[ Y_{12} = Y_{21} = -j\omega C_m \]

Alternatively, a J inverter can take the place of the mutual capacitance \( C_m \) as is shown in Figure 10b. This circuit model will yield the same admittance parameters as that of Figure 10a, however, this leads to a more convenient discussion [7]. Now the circuit has a line of symmetry at \( T-T' \) and therefore an even and odd mode analysis can be performed. If the symmetry plane of
Figure 10b has an electric wall, or is shorted as is the case for the odd mode, then the resonant frequency is

\[ f_e = \frac{1}{2\pi \sqrt{L(C + C_m)}} \]

which is lower than the resonant frequency of the uncoupled single resonator since the coupling effect enhances the capability to store charge when the electric wall is inserted in the symmetrical plane T-T’ [15].

Next, we place a magnetic wall at the plane of symmetry which acts as an open as is the case for the even mode. The resonant frequency for this type of structure is

\[ f_m = \frac{1}{2\pi \sqrt{L(C - C_m)}} \]

Dually compared to the resonant frequency of the shorted symmetry line, the open symmetry line causes the coupling effect to reduce the capability of stored charge. This causes the resonant frequency to be increased. We can now find the electric coupling coefficient as shown below

\[ M_E = \frac{f_m^2 - f_e^2}{f_m^2 + f_e^2} = \frac{C_m}{C} \]

Next, the magnetic coupling will be discussed and an equivalent lumped-element circuit model is shown below in Figure 11a in which L and C are the self-inductance and self-capacitance of two resonators such that \((LC)^{-1/2}\) is equal to \(\omega\) of the uncoupled resonator and \(L_m\) represents the mutual inductance. The lumped element model is only valid at frequencies close to the resonant frequency and therefore is narrow band. From the lumped equivalent model, the following set of equations can be defined [15].

\[ V_1 = j\omega L_1 - j\omega L_m l_2 \]
\[ V_2 = j\omega L_2 - j\omega L_m l_1 \]
Figure 11: (a) Synchronously TunedCoupled Resonator Circuit with Magnetic Coupling and (b) Alternate form with K inverter

From there, we can find the four impedance parameters

\[ Z_{11} = Z_{22} = j\omega L \]
\[ Z_{12} = Z_{21} = -j\omega L_m \]

Alternatively, a K inverter can take the place of the mutual inductance \( L_m \) as is shown in Figure 11b. This circuit model will yield the same admittance parameters as that of Figure 11a, however, this leads to a more convenient discussion [15]. Now the circuit has a line of symmetry at T-T' and therefore an even and odd mode analysis can be performed. If the symmetry plane of Figure 11b has an electric wall, or is shorted as is the case for the odd mode, then the resonant frequency is
\[ f_e = \frac{1}{2\pi \sqrt{C(L - L_m)}} \]

which is higher than the resonant frequency of the uncoupled single resonator since the coupling effect reduces the capability to store flux when the electric wall is inserted in the symmetrical plane T-T’ [15].

Next, we place a magnetic wall at the plane of symmetry which acts as an open as is the case for the even mode. The resonant frequency for this type of structure is

\[ f_m = \frac{1}{2\pi \sqrt{C(L + L_m)}} \]

Dually compared to the resonant frequency of the shorted symmetry line, the open symmetry line causes the coupling effect to increase the capability of stored flux and causes the resonant frequency to be increased. We can now find the electric coupling coefficient as shown below

\[ M_M = \frac{f_m^2 - f_e^2}{f_m^2 + f_e^2} = \frac{L_m}{L} \]

It can be shown [15] that having mixed coupling results in the superposition of the magnetic and electric coupling such that the total coupling between two is

\[ M = M_E + M_M \]

Therefore, one may extract the coupling coefficient M from a physical resonator by creating an EM model with the two resonators and run one simulation with an electric wall, one simulation with a magnetic wall, find the two resonant peaks \( f_m \) and \( f_e \) in both simulations, and calculate the coupling coefficient M. However, it is often not doable to place an electric wall or magnetic wall between two resonators in EM simulation tools and even less doable in experimentation. Therefore, a second approach has been presented and proven [15] that one can obtain the \( f_m \) and \( f_e \)
values from a structure or model with only two elements which are lightly probed, as defined for the \( Qe \) extraction above.

In this method, a structure which contains two resonators is excited via two probe points which are lightly coupled to the resonators (see Figure 12). The frequency response of this structure then present two peaks representing \( f_m \) and \( f_e \). Depending on the distance “s” between the two resonators, the two peaks will be either closer or further in frequency.

![Diagram showing two resonators and probe points](image)

Figure 12: Extracting the Coupling Coefficients between Two Resonators

The general frequency response of the structure in Figure 12 is shown in Figure 13 where two peaks are clearly identified. These two peaks can then be placed into the equation

\[
M = \pm \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2}
\]

where \( f_{p1} \) represents the frequency location of the first peak resulting for the simulated structure, and \( f_{p2} \) represents the frequency location of the second peak (see Figure 13). Since this is a general definition for a synchronously tuned filter, we cannot say which frequency point, \( f_{p1} \) or \( f_{p2} \), are either \( f_e \) or \( f_m \) since it will depend on whether the structure has more magnetic coupling or electric coupling.
Once a spacing is found to achieve the desired coupling coefficients and quality factors of the filter, the RF filter designer can then implement his filter in practice.

2.2 Bandpass Combline Filters

The combline filter has had historical relevance since G. L. Matthaei first published his paper “Comb-Line Band-Pass filters of narrow or moderate bandwidth” in the early 1960’s [29]. The general structure of a combline filter is shown in Figure 14 where Z is the source and load impedance, $C_{Ln}$ are the loading capacitances, and $\theta_0$ is the electrical length of the resonators. The loading capacitance can be calculated as

$$C_{Li} = \frac{1}{Z} \left( \frac{Z}{Z_i} \right) \cot(\theta_0) \frac{\omega_0}{\omega_0} \text{ for } i = 1 \text{ to } n$$

where $Z_i$ is the impedance of resonator $i$, $\theta_0$ is the electrical length of the resonator, $\omega_0$ is the center angular frequency and $n$ is the number of loading capacitors. It should be noted that the electrical length of the resonator must be less than 90 degrees, otherwise the capacitance is zero. Since the resonator lengths are less than 90 degrees, a combline filter is compact when compared to unloaded
resonator filter types such as interdigital filters [30]. This also makes for a wider stopband since the second entry point is pushed out to higher frequencies due to the natural resonance of the unloaded resonator.

![Diagram of a General Structure of a Combline Filter](image)

**Figure 14: General Structure of a Combline Filter**

Since G. L. Matthaei’s original publication, there have been many who have used this filter as the basis for a tunable filter design. In 1965, W. L. Jones first presents the idea of a tunable filter without a high impact on the bandwidth in a paper which describes how one can utilize Matthaei’s original combline topology by implementing a tunable loading capacitance at the end of the resonating lines [31].

At the time of their inception, the combline filter was implemented in coaxial and waveguide form [30]. When communications technology started getting smaller and the desire for portable electronics became prevalent, it was clear that a more compact implementation of RF filters was needed. Therefore, when RF circuitry on planar surfaces became available, filters were amongst those components that were implemented as such. Planar combline filters have been
researched since as early as 1990 [32] and have received a tremendous amount of attention since then, especially for the use of tunable filters [33-36]. These tunable filter’s show great performance between 0.5-2.4 GHz with insertion loss from 1.34 dB to 6 dB at fractional bandwidths of 11.62% to 12.5%. In [33], the author demonstrates a suspended planar structure which is tuned using discrete MEMS switching to load the combline resonators. [34] Demonstrates a tunable filter with reconfigurable transmission zeros and [35-36] show synthesis methods to present additional transmission zeros. Planar type filters continue to be the main source of filter research today as board manufactures continue to improve tolerances and lower dielectric losses.

2.3 Varactor Diodes

A varactor diode is a microelectronic component that provides a varying junction capacitance as the bias voltage shifts. This provides the designer with a component that presents a continuously tunable capacitance over some bias voltage range. The voltage limits and corresponding capacitance values are unique to each varactor and therefore a tradeoff is required by the tunable filter designer to decide which varactor will work best for the specific design. There is also packaging resistance and a junction resistance in the depletion region which degrades performance and should be included in the tradeoff. The characteristics of the varactor will be explained more in the following sections. A diagram of the varactor is shown in Figure 15 and shows how the depletion region is altered depending upon low or high reverse voltage being applied (V_L and V_H) and packaging effects are being ignored.
2.3.1 Equivalent Circuit Model

When factoring in the effects of a varactor component, it is important to have an understanding of an equivalent circuit model. This circuit is shown below in Figure 16 and contains the series resistance of the package (Rs), the resistance of the junction (Rj), and the capacitance of the junction (Cj). The capacitance and the resistance of the junction are varied by adjusting the reverse bias voltage across the diode where the package resistance remains constant.
It is typical that vendors will provide tabulated resistance and capacitance values over different bias voltages. The resistance values provided usually combine the package series resistance and junction resistance into one value. It will be seen in Chapter 3 that this makes for a good approximation and is all the designer needs to get simulation results that compare well to measurements. Figure 16 shows the tradeoff in selecting a varactor where “R” is taking into account both the packaging resistance and the junction resistance.

![Figure 16: Varactor Diode Trade-off](image)

2.3.2 Tunable Filter Performance Impacts

Varactors have been demonstrated in many tunable filter papers which prove to have great performance below 2 GHz [4], [34-36]. These papers reveal the knowledgebase of varactor loaded filters and achieve tuning ranges 60-100% and insertion loss as low as 1.04 dB at the higher tuned frequencies. However, these papers also demonstrate that tunable filters tend to have performance that does not obtain the same level of vitality as fixed filters. There is not an exception to this undesired outcome in the work presented herein. It can be seen in general that tunable filters have
slower roll-off in the rejection region, more insertion loss in the passband, and less dramatic poles in the return loss response. These undesired responses are caused by the resistances presented by the varactor: $R_s$ and $R_j$. There are considerations which can be made and will be discussed in Chapter 3 that can mitigate some of the loss effects, but ultimately the $Q$ of the tuning element lowers the performance of the filter substantially. An example of how a varactor loads a resonator and equivalent circuit model is shown in Figure 18 where $R_s$, $C_j$, $R_j$, $C_r$, and $L_r$ are the packaged resistance, junction capacitance, junction resistance, the resonator’s self-capacitance, and the resonator’s self-inductance, respectively. It can be seen that the interface is relatively straightforward and easy to implement.

Figure 18: Resonator Line Loaded with a Varactor Diode (a) and Equivalent Circuit Model (b)

If more capacitance value flexibility is needed than what is available for purchase, varactors can also be used in both series and parallel configurations. However, when placing
components in series and/or parallel, not only are the capacitance values being manipulated but the combination of junction and package resistance are affected as well. These undesired elements should be taken into consideration early on in the development of a tunable filter. In addition, it should be noted that as the electrical length of the printed resonator increases, the effect of the series resistance and junction resistance on the overall Q decreases [37]. This leads to a trade-off between lower capacitance and higher resistance.
CHAPTER 3
TUNABLE BANDPASS COMBLINE FILTER DESIGN EXAMPLE

To adequately present the filter processes outlined in Chapter 2, a design is presented of the tunable combline filter design utilizing a varactor diode as the loading element. The design parameters and goals were chosen to fulfill a joint effort with Harris Corporation, consequently an L-band filter centered at 1.5 GHz with a 10% bandwidth was chosen.

3.1 Conventional vs. Tunable Filter Design Considerations

This section will discuss some of the main design considerations when designing a tunable combline filter which are different from those details considered in a fixed combline filter. This includes the tuning element filter performance impacts and the tuning limitations of the combline filter.

3.1.1 Tuning Element Filter Performance Impacts

As was discussed in Chapter 2, the largest form of degradation in tunable filters comes from the unwanted parasitic resistances found in the tunable elements. Until a tunable component becomes available that does not present such a high resistance, tunable filters will continue to have slow roll-off in the rejection region, high insertion loss in the passband, and less dramatic poles in the return loss response when compared to fixed filters. Here we discuss how exactly the loss of the tuning element directly impacts the filter performance.
We will begin by discussing the so called quality of an RF resonator is defined by the equation

\[ Q = \frac{\omega (\text{average energy stored})}{(\text{energy loss/second})} = \frac{W_m + W_e}{P_l} \]

where \( W_m \) and \( W_e \) are the energy stored by the inductance and capacitance of a resonator and \( P_l \) is the power dissipated due to the resistance of a resonator [28]. A perfect resonator would have a “Q” equal to infinity as there would be no energy loss in the circuit. Therefore, any degradation in insertion loss from the ideal filter response is directly related to the Q of each resonator. The resistance presented by the printed resonator is dependent upon the materials used for the printed circuit board (PCB). Generally, a low loss substrate will be used with a resonator that is not made with lossy metallic materials. This causes the resistance presented by the printed resonator line to be negligible in comparison to the resistances presented by a tuned loading element. This can be expressed as

\[ R_{\text{loss}} = R_{\text{resonator}} + R_{\text{load}} \]

therefore,

\[ \text{if } R_{\text{load}} \gg R_{\text{resonator}} \]

\[ R_{\text{loss}} \cong R_{\text{load}} \]

To combat the resistance presented by the load, it is preferred to select a resonator length that is physically long. Rebeiz presented in [37] the effects of resistance of a loading element on the overall Q of the filter. He shows that the higher the electrical length of the printed resonator, the higher the Q of the overall circuit. In fact, when the printed resonator length reaches 10% of the guided wavelength the Q of the varactor dominates the overall Q of the resonator. Therefore, it is preferred to select the resonator length based on the highest center frequency point and using the highest value of capacitance at this frequency point since this provides the longest printed
resonator length. This assures maximum resonator length achievable over the desired tuned bandwidth which provides the highest Q and lowest insertion loss which is especially important for high power applications.

3.1.2 Theoretical Limitations of the Tunable Combline Filter

In addition to the loss of the resistance presented by the tunable element, there are also limitations in tuning of the combline filter. Assuming there was no loss in the tuning element, a combline filter would still not be able to tune over bandwidths that are limited by the physical structure itself. The coupling coefficients (M) and external quality factor (Qₑ) were defined in Chapter 2. For a 3 pole filter, these values are associated with the pair of resonators as shown in Figure 19. Both the M values and the Q values are directly related to the length of the resonators as well as the space between them.

![Figure 19: Resonators Relations to Qₑ and M](image-url)
Since the center frequency of the bandpass filter is tuned by varying the loading capacitance, there is no physical change on the printed resonator as the loading capacitor value is tuned. This means that the center frequency is tuned up or down the electrical length of the printed resonators is effectively changing. For example, if a resonator was designed to have an electrical length $\theta_0 = \lambda/6$ at a center frequency of 1.5 GHz, then there is a capacitance value $C_L$ which would be derived and load the resonators to achieve the desired response (see section). Then, as the capacitance value $C_L$ is lowered by raising the reverse bias of the varactors, the center frequency of the bandpass filter can be shifted to 2 GHz. Now, at 2 GHz the electrical length $\theta_0$ of the printed resonators will increase to $\lambda/4.5$. This will change the coupling coefficients and external quality factors because the mutual capacitance will yield different admittance values. Since the $M$ and $Q_e$ values are directly derived from the $g$-values, as the coupling coefficients and external quality factors are shifted further from their nominal derived values, the filter response shifts from the desired filter response (e.g. Butterworth, Chebyshev, and elliptical).

3.2 Design Example of a Tunable Combline Filter

Before getting to the tunable balun filter, it is worthwhile to show the design, simulation, and test of a tunable filter which will be utilized in the design of the tunable balun filter. Therefore, we now present the design of a tunable combline bandpass filter with a 10% fractional bandwidth that is centered at 1.5 GHz. This frequency is was chosen as it is in L-band is commonly used for communications purposes. The filter will be fabricated on a 60 mil thick Rogers 4003 material which has a dielectric constant of 3.55 as this material was readily available to the author. For this design, a 3 pole Chebyshev response was chosen with an in band ripple of 0.04321 dB since the $g$-values were tabulated in [15]. The $g$-values associated with this type of filter are $g_0 = g_4 = 1$, $g_1$
= g_3 = 0.8516, and g_2 = 1.1032. These values will be used for the derivation of the external quality factor (Q_e) and the coupling coefficient (M_{12} and M_{23}).

Before we begin extracting the Q_e and M values, we must first choose a resonator electrical length and impedance. As a starting point, the resonator was chosen to have an impedance of 100 \Omega. This impedance allowed for a printed resonator width that matched well with the varactor package soldering pins. Since the width of a 100 \Omega line is approximately 33 mils wide, the trace is robust enough such that line losses where not found to be high. An electrical length of 60° was chosen at 1.5 GHz so that at 2 GHz the electrical length will be 80°. This means the center frequency can be tuned comfortably to 2 GHz assuming the loading device can achieve the needed value of capacitance (i.e. at 2 GHz the resonator length is less than 90°, see section 2.2).

The loading capacitance is then defined as

\[
C_L = \frac{1}{Z_S} \left( \cot(\theta_0) \right) \frac{\cot(\theta_0)}{2\pi f} = 0.6125 \text{ pF}
\]

where “Z_S” is the system impedance of 50 \Omega, “Z_R” is the resonator impedance of 100 \Omega, “\theta_0” is the resonator’s electrical length of 60°, and “f” is the center frequency of 1.5 GHz. This formula can then be used to calculate the loading capacitances of the different tuned frequencies.

Every calculated value will be optimized as is shown in section 3.2.3 but these derivations give a good starting point. Next we will find the external quality factor and the coupling coefficients as outlined in section 2.1.2. For the extraction of these values from the physical structure, the High Frequency Structural Simulator (HFSS) design tool by ANSYS was used.

3.2.1 Derivation and Extraction of External Quality Factor

For this design example, we will be utilizing a coupled feed line to inject the signal into the filter structure. Therefore, the external quality factor will determine the spacing between the
feed line and the first resonator of the filter. First, using the $g$-values and the formula presented in Chapter 2, the external quality factor is calculated.

\[
Q_{e1} = Q_{e2} = \frac{g_0 g_1}{FBW} = \frac{g_3 g_4}{FBW} = 8.516
\]

We can manipulate the formula found in section 2.1.2 and find the 3 dB bandwidth required to obtain the wanted external quality factors calculated above since we know the desired $Q_e$ and the center frequency.

\[
\Delta f_{3dB} = \frac{f_0}{Q_e} = \frac{1.5 \text{ GHz}}{8.516} = 0.175 \text{ GHz}
\]

Now we can set up a model in HFSS to find the spacing that provides this 3dB frequency delta. As presented in Figure 9.b, one side of the model consists of the coupled feed network and the other is comprised of a probe point line is used in close proximity to the first resonator in the filter structure, see Figure 20. The spacing “S” between the coupled line and the first resonator is then varied until the desired 3 dB delta frequency is achieved. This is an iterative process of changing the spacing and calculating the 3 dB delta frequency but after a few iterations the appropriate value was found. The box that surrounds the HFSS model is an air box and the six faces of that box have been defined with a radiation boundary. All the metal is defined as copper and the substrate is defined according to the material chosen for the design example (namely Roger’s 4003 type with a thickness of 60 mils). The ports are waveports and the capacitance which is loading the resonator line is defined as a lumped impedance. At this point in the design we are ignoring resistive affects so the loading element is purely capacitive.
Figure 20: Model for Extracting the $Q_e$ in HFSS from a 3D View (a) and a Top View (b)

The small signal S21 from the model shown in Figure 20 can be seen in Figure 21. A marker has been placed at the peak of the transmission loss curve and shows about -20 dB of transmission loss. A delta marker is placed 3 dB below the peak value at -23 dB to obtain the 3 dB delta frequency. It can be seen that a 3 dB delta frequency of approximately 0.175 GHz was achieved with a spacing of 0.46 mm or 18.1 mils. Typically for a less expensive fabrication, it is best to keep any metal to metal spacing greater than 5 mils so 18 mils is a comfortable starting point. Using the method outlined, one can plot $Q_e$ vs. Spacing “S” as shown in Figure 22.
Figure 21: Simulated S21 for Extracting Qe from the Model of Figure 20

Figure 22: External Quality Factor vs. Spacing
3.2.2 Derivation and Extraction of Coupling Coefficient

Next we will find the coupling coefficient with the g-values provided. The coupling coefficient is determined through simulation by varying the spacing between the loaded resonators. Therefore, the coupling coefficients will determine the spacing between the first and second resonator as well as the second and third resonators. First, using the g-values and the formula presented in Chapter 2, the coupling coefficients are calculated.

\[ M_{1,2} = M_{2,3} = \frac{FBW}{\sqrt{g_1 g_2}} = 0.103 \]

Now we can set up a model in HFSS to find the spacing that provides the coupling coefficients “M” calculated above. As presented in Figure 12, both sides of the two resonators are lightly coupled by a probe line feed which is 0.125 mm from the resonators which later caused errors in the derivation of the coupling coefficients due to loading of the resonators, see Figure 23. The spacing “S” between the resonator lines is then varied until the appropriate “M” values are extracted from the corresponding graph. This is an iterative process of changing the spacing and calculating the M value but after a few iterations the appropriate value was found to be 2.4 mm. The model was set up similarly as that described for the coupling coefficient with an airbox, radiation boundary, waveports, and lumped impedances loading the resonators.
Figure 23: Model for Extracting the Coupling Coefficient in HFSS from a 3D View (a) and a Top View (b)

Figure 24: Simulated S21 for Extracting M from the Model of Figure 23
Two markers were placed in the graph shown in Figure 24, both at the peaks represented by the response of the structure. The coupling coefficients can then be calculated.

\[ M = \pm \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2} = 0.1097 \]

Although this value is not identical to the coupling coefficients derived previously, this is close enough to move forward and integrate the filter together. As for the external quality factor, a plot can be generated to show the coupling coefficient vs. spacing “S”, as shown in Figure 25. Once all the pieces are together, it can be seen that the filter performance is not quite perfect as the extracted values only approximate the desired response. Therefore, a process of optimization must be made to fully achieve the wanted performance.

![Graph showing coupling coefficient vs. spacing](image)

**Figure 25: Coupling Coefficient vs. Spacing**
To support the claim stated in section 3.1.2, here we will show how the coupling coefficient and the external quality factor vary as the center frequency is tuned. These models are set up the same as those shown in Figure 20 and Figure 23 however this time instead of varying the spacing, we will vary the loading capacitance. By changing the capacitance to 1.068 pF, 0.6125 pF, and 0.3310 pF we will tune the center frequency of the filter to 1.25 GHz, 1.5 GHz, and 1.75 GHz, respectively. The variation in external quality factor and coupling coefficient are what account for the change in bandwidth when tuning as well as overall filter shape since the formula for both $Q_e$ and $M$ contain the lowpass prototype values as well as the fractional bandwidth.

![Figure 26: External Quality Factor Variance over Frequency](image-url)
3.2.3 Full Filter Model Including Tuning Element Parasitics

Once the coupled feed lines spacing and the resonator lines spacing were extracted, a full model can then be created in HFSS which can then provide a full filter response. This model, shown in Figure 28, consists of three resonator lines, three loading lumped impedance boundaries, two coupled feed lines, an air box with radiation boundaries, and two wave ports to inject the signal. The first and third resonators have added length to compensate for the effects the feed network has on their resonant frequency. [15]

The response is shown in Figure 29 and it can be seen that the desired center frequency and bandwidth are not exactly realized. This shows that the probe point used to extract the external quality factor and coupling coefficients were loading the resonators more than desired. However, we can use this as a starting point for optimizing to obtain the desired response.
Figure 28: Full Ideal Model with a Side View (a) and a Straight View (b)

Figure 29: Initial Small Signal Response for Filter Network Shown in Figure 28
After approximating the filter structure with the process shown above, the resonator length, spacings, and loading capacitance value were all varied while observing the performance until the desired response was achieved. The demonstrated performance is shown in Figure 30 which has in-band performance of $S_{21} > -1$ dB and $S_{11} < -15$ dB with rejection greater than 35 dB and a fractional bandwidth of approximately 10%. However, this response does not include the effects of the added series resistance presented by the varactors. Once series resistance is added, which demonstrates a more realistic case, we will begin to see degradation in the overall filter response. This degradation is not something that can be tuned out as it is the result of fixed parasitic resistances within the tuning elements.

![Nominal S-Parameters](image)

**Figure 30: Final Ideal Small Signal Performance of Filter Network**
Next, we will show an example of a non-idealized model will be that of the varactor loaded resonator. The varactor used for this demonstration is the MACOM MA46H072 which has a series resistance in our frequency of interest between 0.4 Ohms and 0.6 Ohms and a capacitance of 1 pF to 4.5 pF according to data provided by MACOM. Since it was desired to have a capacitance value below 1 pF such that the printed resonator length remains longer, two varactors were placed in RF series and DC parallel, as shown in Figure 31. This resulted in resistance between 0.8-1.2 Ohms and a capacitance of 0.5-3.2 pF.

![Figure 31: Varactor Configuration](image)

To implement the addition of the series resistance in the HFSS model is fairly simple for the varactor model. The only thing that needs to be added is a sheet with a lumped impedance boundary defined with the resistance provided by the manufacturer. This resistance was added between the capacitor and the via to ground. Bias lines were also added to the model to assure there were not any unknown parasitics associated with those as well. The model can be seen in Figure 32.

To tune the center frequency in the model, only the capacitive and resistive values need to change and the solve frequency of the simulator needs to be shifted to the center of the resonance

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when using the HFSS model setup with an interpolating sweep. Therefore, Figure 33 shows the final simulated performance at three different tuned frequencies. The loading capacitances in series required for the shown responses are 3.5 pF at 1.25 GHz, 2.4 pF at 1.5 GHz, and 1.75 pF at 1.75 GHz. In addition, Figure 34 shows the wideband filter performance at the tuned frequency of 1.5 GHz. The re-entry point can be seen around 5.5 GHz which is nearly 4 times the center frequency of 1.5 GHz.

Figure 32: Varactor Loaded Resonator Configuration in HFSS

Figure 33: Varactor Loaded Filter Simulated Response at 3 Tuned Center Frequencies
Figure 34: Simulated Wideband Response of Filter Tuned to 1.5 GHz

3.2.4 Experimental Verification

As stated earlier, the filter will be fabricated on a 60 mil thick Rogers 4003 material which has a dielectric constant of 3.55 as this material was readily available to the author. The measured responses can be seen in Figure 35 which show good correlation to the simulated data of Figure 33. The degradation in insertion loss is due to fabrication tolerances of the in-house prototype build. The voltages used to achieve the center frequencies shown are 5V, 8V, and 13V. Passbands can be achieved as low as 1 GHz and as high as 2 GHz with degraded insertion loss and return loss. The wideband performance is shown in Figure 36 and shows good rejection with the re-entry point pushed out to nearly 4 times the passband at 5 GHz. This is slightly lower than the simulated response shows, but the correlation in overall shape is very close to what was simulated otherwise. There is also additional insertion loss at the high end of the band which is due to varactor parasitics that are not accounted for in the model.

It can be seen that the achieved performance was better than 10 dB return loss, better than 5 dB Insertion loss, and the bandwidth is approximately 10%. This filter provides a good starting point for the tunable balun filter which will be presented in Chapter 4.
Figure 35: Measured Tunable Filter Response

Figure 36: Measured Wideband Response of Filter Tuned to 1.5 GHz (S11 Blue Curve, S21 Red Curve)
In order to continue the trend of compacting radio frequency (RF) and microwave circuitry, an effort must be made to consolidate components to achieve similar performance in a fraction of the physical space. Integrating a tunable filter and balun structure takes less space when compared to a reconfigurable filter and balun in cascade and provides similar performance. Such a structure can be used to minimize board real-estate in wireless applications. Two examples are connecting a 50 Ω unbalanced system to a dipole antenna in a cell phone, or connecting to a balanced type mixer where filtering the unwanted signal products is necessary.

Here we present the first tunable balun circuit utilizing a combline structure. In addition to achieving superior phase balance compared to the previous listed work, this topology also allows for scaling to higher order filters. A three-pole example tunable filter and integrated balun layout is shown in Figure 37 and the design is discussed in section 4.2.
4.1 Theory of Performance

The design of the combline filter is mature and the design steps for the tunable filter-balun shown here follows the same design steps as shown in Chapter 3. However, once the filter has been designed, the next step of adding balanced port coupled feed is unique when compared to a conventional combline filter. In a conventional combline filter, the physical structure is a mirror image about the center of the filter such that the feed lines on both the input and output side are unbalanced. The extraction of the coupling coefficient M follow the same steps as shown in Chapter 3 since there is no change in the physical structure between the first and last resonator. However, here we place ports at the center of the coupled feed such that \( L_2 \approx L_1/2 \) and ground the outer portion (see Figure 37) to achieve a balanced port as a Marchand balun which changes the external quality factor of the balanced feed. To demonstrate the affects that the balanced port introduces to the external quality factor, a simulation was set up in HFSS as is shown in Figure 38. The coupled feed port is set up to be a differential pair in the HFSS terminal mode simulation.

Figure 37: Dimensioning of an Example Three Pole Tunable Filter-Balun
Next we plot the variation in the external quality factor with respect to changes in “S”, “G”, and the tuned center frequency which are shown in Figure 39, Figure 40, and Figure 41. It can be seen that the slope of $Q_e$ vs the spacing in Figure 39 is roughly the same as shown in Figure 22 with an offset according to spacing which shows the balanced feed will need to be closer to the first resonator than the unbalanced feed. In Figure 40, we can see that the gap has a small effect on the external quality factor when compared to the spacing “S”. This means that we can place these lines close together to maintain a closely coupled differential line and not have to worry about fabrication tolerances. Compared to the unbalanced feed, the balanced external quality factor shows a higher sensitivity to the change in center frequency. This is probably due to the unbalanced nature of the coupling presented by the balanced port to the unsymmetrical resonator and suggests that this design will inherently be narrower band than the purely combline filter.
Figure 39: External Quality Factor vs. Spacing

Figure 40: External Quality Factor vs. Feed Gap

Figure 41: Change in External Quality Factor over Frequency
4.2 Design Example of the Tunable Combline Filter-Balun

The design utilized for this tunable combline balun filter is the same Chebyshev three-pole combline filter with a bandwidth of 10% and a ripple of 0.4321 at the center frequency of 1.5 GHz as shown in Chapter 3. Once the preliminary design steps outlined in Chapter 3 and section 4.1 were completed, the design was then simulated in HFSS. After approximating the filter structure with the process shown above, the resonator length, spacings, gap, and loading capacitance value were all varied while observing the performance until the desired response was achieved. A Rogers 4003 substrate with a height of 60 mils was used in the simulated structure of Figure 42. This model also has placement for the varactors, isolation resistors for bias, and bias lines.

![Simulated Balun Filter with Varactor Loaded Lines](image)

Figure 42: Simulated Balun Filter with Varactor Loaded Lines

The structure shown here can be either a fixed filter and balun circuit or a tunable network depending on whether the loading elements are capacitors or a tunable component, such as the varactor diodes used herein. To accurately model the varactor, a series RC network was used in HFSS. Series capacitance and resistance values over bias voltages were provided by the varactor supplier which was Skyworks. Due to component availability, these varactors are different than
those used for the tunable filter presented in Chapter 3 but present similar loading capacitance values. By the simulation plots in Figure 43, you can see that these varactors provided slightly more series resistance that the MACOM parts used before. The tuned capacitances used were 2.75 pF, 2.3 pF, and 1.87 pF in order to achieve the responses centered at 1.35 GHz, 1.5 GHz, and 1.6 GHz, respectively.

![Simulation Small Signal Response of Tunable Balun Filter](image)

Figure 43: Simulation Small Signal Response of Tunable Balun Filter

A wideband performance plot is also shown below in Figure 44. It can be seen that the re-entry point is about the same frequency location but it does have a different shape. This provides an idea of what the expected re-entry point should be for the fabricated balun filter.

For this structure to provide a balanced port, the simulated phase response generated by the structure of Figure 42 must present approximately 180 degree offset and equal amplitude between the balanced feed lines. The phase response of the simulation was found to be a 180 ± 1 degree delta phase, and an amplitude difference of 0.3 dB is achieved between tuned channels as shown in Figure 45 and Figure 46, respectively. For ease of viewing, the phase and magnitude
imbalances have been normalized to the center frequency of each tuning state which were accomplished by changing the loading capacitances to 3.34 pF, 2.75 pF, 2.3 pF, and 1.87 pF.

Figure 44: Simulated Wideband Response of Tunable Balun Filter Centered at 1.5 GHz

Figure 45: Simulated Balun Phase Imbalance
To display the radiated loss shown in Figure 47, the material parameters of the model shown in Figure 42 were set to lossless and the radiated losses were calculated as

\[
\text{Radiation Loss}_{dB} = 10 \cdot \log(1 - |S_{11}|^2 - |S_{21}|^2)
\]

This plot implies that at the center frequencies, the balun filter is losing some energy due to radiation, although those losses are not significant and therefore most losses are due to dissipation. If this was an ideal filter and balun, then all the energy would be accounted for by S21 and S11 and therefore the equation shown above would yield an answer which is negative infinity.
Figure 48 shows the current distribution of the tunable balun where plots a-c are the current distribution at the low, mid, and high frequency tuned responses. It can be seen that all the line elements are resonating as would be expected at the specified frequencies for the combline type structure. The tuned center frequencies are 1.35 GHz for (a), 1.5 GHz for (b), and 1.62 GHz for (c).

![Figure 48: Current Distribution at the (a) Low, (b) Mid, and (c) High Tuned Frequencies](image)

**4.3 Experimental Results**

To verify the accuracy of the design approach and simulation, a board was fabricated on a 60 mil Roger’s 4003 substrate (see Fig. 4). Skyworks varactors were used for the tuning elements which load the resonators and a 2 MΩ resistor provided bias isolation for the varactor diodes. A comparison of the achieved measured results to previous work is outlined in Table 1.
Table 1: Comparison of Measurements to Previous Work

<table>
<thead>
<tr>
<th>Reference</th>
<th>Percentage Bandwidth</th>
<th>Tuning Range (GHz)</th>
<th>Phase Balance (degrees)</th>
<th>Amplitude Imbalance (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>This Work</td>
<td>10%</td>
<td>1.25 – 1.60</td>
<td>180 ± 1</td>
<td>0.6</td>
</tr>
<tr>
<td>[23]</td>
<td>6.5%</td>
<td>1.49 – 1.545</td>
<td>180 ± 5</td>
<td>0.5</td>
</tr>
<tr>
<td>[24]</td>
<td>14-18%</td>
<td>0.62 – 1.04</td>
<td>180 ± 5</td>
<td>0.5</td>
</tr>
<tr>
<td>[17]</td>
<td>Not Provided</td>
<td>0.62 – 1.02</td>
<td>180 ± 7</td>
<td>0.8</td>
</tr>
</tbody>
</table>

Figure 49: Fabricated Tunable Comline Filter with a Balanced and Unbalanced Port

The measured responses are compared to the simulated responses in Figure 50 and Figure 51 and show excellent correlation. The voltages used to tune the varactors to the appropriate center frequency were 16V, 25V, and 40V. It can also be seen in Figure 52 that all of the tuned center frequencies tested achieve a phase difference of 180 ± 1 degrees across the designed 10%
bandwidth of the filter response. Likewise, Figure 53 shows the amplitude imbalance at different tuned frequencies and a difference of 0.6 dB max is achieved.

Figure 50: Measured versus Simulated S21 at 16V, 25V, and 40V.

Figure 51: Measured versus Simulated S11 at 16V, 25V, and 40V
Figure 52: Measured Balun Phase Response

Figure 53: Measured Balun Amplitude Imbalance
CHAPTER 5
CONCLUSION

The motivation for this thesis was to fulfill a joint effort between Harris Corporation and the University of South Florida to increase the understanding and development of reconfigurable filters. This thesis presented the design, simulation, and test of a tunable combline filter and balun. It demonstrated in Chapter 1 an introduction to the field of tunable filters and baluns and presented the direction of research at its present state. Chapter 2 showcased the theory and design approaches to filters including the lowpass prototype, the bandpass transformation, and the realization of the external quality factor and coupling coefficients. There was also a focus on combline structures and we touched on the history of the Marchand balun and its uses.

Next we demonstrated a practical design of a combline filter in Chapter 3 and focused on the consideration differences between a fixed and tunable device. The combline structure was chosen due its versatility and ability to be a planar structure. A design process was proven and a Chebyshev filter with 10% bandwidth at 1.5 GHz and a tuning range of 33% was achieved. The design was fabricated on a Rogers 4003 substrate that was 60 mils thick and had a dielectric constant of 3.5. The varactors used were provided by MACOM and showed low series resistance which allowed the filter’s insertion loss to be between 2 and 3 dB.

We also discussed issues that were faced during the design process. For example, in Chapter 3, we show a way to extract both the external quality factor and coupling coefficient from an HFSS model. It was later discovered that the so called lightly coupled probes were loading the
resonators which caused an overall filter response that did not meet expectations. However, we were able to optimize the filter performance later to achieve the desired response. The effect of series resistance and how it affects the insertion loss of the filter was also presented along with a way to mitigate its affects. We discussed that more loss occurs when the printed resonator is electrically small vs. electrically large and suggested to take this into account early on when selecting a resonators length if you are designing for high power.

Finally, in Chapter 4 a novel filter structure with balanced and unbalanced ports was presented. This paper provides design details as well as a comparison of simulated and measured results which show excellent correlation. A phase balance of 180 ± 1 degrees and an amplitude imbalance of 0.6 dB were achieved within the filter’s 10% bandwidth which are competitive results when compared to state of the art tunable balun’s. A third order filter was selected for demonstration purposes, however the structure can easily be scaled to higher or lower orders. The balun filter can be made a fixed or tunable network by loading the combline structure with a fixed or tunable capacitance (e.g. varactors, MEMS, etc.). The fabricated balun was done on a Rogers 4003 material and Skyworks varactors were used based on availability.

The design of this tunable balun filter can be built upon using already known methods of designing reconfigurable combline filters. Through these design methods, one can choose to implement reconfigurable baluns which can tune transmission zeros and bandwidth in addition to the center frequency. There is a wealth of knowledge on tunable combline filters and now, with a convenient method for implementing balun structures to designs, consolidation of the planar balun and filter can lead to saving real-estate and continue the trend of shrinking RF electronics.
REFERENCES


