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Absorptive Reflectionless Filters

by

Guy Barry Lemire

A thesis submitted in partial fulfillment of the
requirements for the degree of

Master of Science
in
Electrical and Computer Engineering

Thesis Committee:
Branimir Pejcinovic, Chair
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Portland State University
2020

Abstract

For many applications requiring some sort of signal filtering or signal conditioning, the filter requirements are usually approached with a single purpose in mind, which is to maximize both passband signal amplitude and stop band signal attenuation to the load with little to no thought given to what happens to the stop band signal energy. Many conventional filters have very poor impedance matching in the stopband resulting in reflected energy or large return loss (S_{11}). This reflected energy can then cause interactions with adjoining system components which do not in general respond well to spurious reflected energy and can result in degradation of system performance or other unintended consequences [1].

This thesis implements and verifies the design procedure and examines the frequency scaling of a novel passive filter design methodology proposed by Morgan and Boyd [2] which the authors claim results in an easy to realize reflectionless filter which has stopband reflection response superior to conventional passive filters. Following the proposed design methodology, a reflectionless filter was simulated and then realized in a hardware prototype and good agreement observed between simulation and measurement results. To determine the quality of stopband response, the reflectionless filter response was compared to a Butterworth admittance complimentary diplexer designed using accepted techniques [3], [4], [5]. The reflectionless filter showed similar measured stopband response to the diplexer but this was gained with a much simpler design process than was required for the diplexer.

To verify the ability of the procedure to scale the design requirements, a filter with a decade wider bandwidth was also designed and the measured response compared to a similarly scaled Butterworth diplexer. For this frequency range of interest, the reflectionless filter exhibited superior stopband rejection when compared to the diplexer. Simulation and measured results were in good agreement for both filters.

In conclusion, this work was able to realize a reflectionless filter using Morgan and Boyd's design procedure and the measured results were in good agreement with simulation results. The stopband of the reflectionless filter was comparable to a similar diplexer but with much less design effort required. The scaling of design parameters for reflectionless filters was also demonstrated.

Dedication

To Nancy De Martino

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Glossary

Cauer Ladder Network filter topology consisting of cascaded series and shunt active components

Passband portion of the frequency spectrum that can pass through a filter

Stopband portion of the frequency spectrum that is attenuated by a filter

Transition Region range of frequencies where the transition from the passband to the stopband occurs

Scattering Parameters provide a complete description of a network as seen at its ports

Even Mode Analysis a circuit analysis technique where signals of equal amplitude and phase are applied to each port of a symmetrical circuit.

Odd Mode Analysis a circuit analysis technique where signals of equal amplitude and opposite phase are applied to each port of a symmetrical circuit.

Duality a circuit analysis technique that allows the transformation of circuit components to their dual or equivalent components to help simplify a circuit while maintaining its equivalency

Microstrip a transmission line formed on the outer layer of a printed circuit board. It consists of a conducting strip separated from a ground plane by a dielectric layer.

Return Loss (Γ) the ratio of reflected power to incident power on a given network port.

Also known by the scattering parameter matrix element S_{nn} where n is the port

Insertion Loss the ratio of output power measured at port i to input power at port j .

Also known by the scattering parameter matrix element S_{ij}

De-embedding the mathematical act of removing the effects of a test fixture so that the device response can be directly measured.

Chapter 1: Introduction

Maximizing useful signal power from the source to the load has been a fundamental design goal in electrical networks since the dawn of modern electronics in the early 20th century [6], [7]. The mechanism for this matching source to load impedance was usually through some intermediary circuit located between the source and the load. The impedance matching network or more generally “filter” can be as simple as a single termination resistor placed near the load, to an active multitap adaptive broadband filter to quarter wavelength microstrip stubs. Regardless of complexity or target application in which the filter is required, these structures can have one or several design parameters to be optimized which are (but not limited to)

- maximum power transfer
- selective filtering (passband) for a frequency range of interest and
- controlling reflection of the signal from the load in an unwanted (stopband) frequency range of interest
- waveform shaping maximizing signal to noise ratio

Depending on the application and filter topology chosen, in general the design becomes more difficult and complex the more design parameters that need to be optimized.

Additionally, many of the current solutions have problems or shortcomings which include inherently limited bandwidth (e.g. quarter wavelength stubs), excessive passband loss, reflective peaking in the transition region and can have a large physical topology which makes implementation difficult for space limited applications. Often the shortcomings of the filter design necessitate additional elements that need to be inserted into the system design to mitigate the unwanted effects.

A subclass of filters called absorptive filters allow for optimization of both passband and stopband characteristics as compared to conventional reflective-type filters. As the name suggests, absorptive filters attenuate the unwanted out of band incident signals by absorption through resistive elements rather than reflection. The concept of absorptive filtering has been around for many decades, with one of the earliest and simplest form of this type of filter is a leaky wall filter which couples unwanted energy to an auxiliary signal path with an absorptive load [6]. No comprehensive list of the different types of absorptive filters was found in the literature, but for the purposes of this work Bulja [8] suggested that absorptive filters can be categorized into transmissive-type absorptive filters and reflective-type absorption filters. Transmission-type absorptive filters can come in several different topologies. The first provides two or more signal paths, of which the sideband path(s) are used to attenuate the stop band signals of the transmission path. The second topology contains the absorptive elements in the body of the filter.

A classical approach to a designing a filter that is matched in both the passband and the stopband is to use a diplexer (or multiplexer for more than two frequency bands) and then terminating all but the main path with matched loads [9]. A diplexer is a three-port network that splits the incoming signal from the common port into two frequency dependent signal paths (sometimes called channels). Diplexers can be realized in admittance complimentary or impedance complimentary architectures [3].

To address the out of band reflective problem a novel methodology for absorptive filters has been recently proposed by Morgan and Boyd [2]. Morgan terms these filters reflectionless rather than absorptive because theoretically they have reflection coefficients of zero at all ports for all frequencies for ideal elements [10]. To follow Morgan's terminology, these filters will be designated henceforth as reflectionless. This filter has the similar out of band properties as a diplexer but with a simpler design process and more importantly, all ports are matched due to symmetry. This allows for cascading of reflectionless filter sections without the debilitating inter-stage interaction found with many filters.

This thesis will demonstrate the process of design, simulation and realization of a reflectionless filter based on the design criteria outlined in Morgan and Boyd. The realized prototype filter will be measured with a Vector Network Analyzer (VNA), and the results compared to the Morgan paper results in a Scattering Parameter (S-parameter) format. Additionally, a Butterworth diplexer will be designed, a prototype measured, and the results will be compared to the reflectionless filter. To further illustrate the design properties, a filter will be scaled by a decade above the Morgan and Boyd paper and the results measured and presented.

A description of the theoretical background and assumptions that are made in the reflectionless design methodology will be demonstrated in Chapter 2. Specifically, it will be discussed how Morgan and Boyd use even and odd mode analysis, duality and symmetry to obtain important mathematical relationships that will guide the topology

of the normalized filters. The optimum filter order is determined as well as component values. In Chapter 3 simulations are performed for a reflectionless filter with ideal design parameters and a reflectionless filter with non-ideal elements and microstrip transmission lines. Both results are compared to Morgan's results to check on the validity of the design process in this work. Chapter 4 outlines the PCB layout decisions and fabrication materials. Chapter 5 presents the laboratory results for both the reflectionless filter and the diplexer and compares them to the Morgan results. Chapter 6 is a discussion of the results followed by conclusions and some ideas for future work.

Chapter 2: Background and Theory

Filter design methodologies has evolved to where a generally accepted or canonical approach can be used to realize a design. The designer need not have intimate knowledge of the underlying mathematical concepts but simply follow a set of steps to realize a filter which meets the design requirements. Filter responses can mostly be categorized by the frequencies that the filters pass or allow through, such as low pass, band pass or high pass. Several types of filters families such as the Butterworth, Chebyshev, Elliptical and Bessel can be designed to meet the frequency responses desired but each family will have slightly different response signatures.

A basic overview of the canonical approach to filter design will be presented in this chapter and without loss of generality will involve the design of a low pass Butterworth filter. The general topology of filters are of a Cauer or ladder topology as shown in Figure 1, which for a low pass filter consists of alternating series inductors and shunt capacitors.

Figure 1: Cauer ladder low pass filter topology

As can be seen from the Cauer topology, the ladder network consists of a total of n elements where n represents the order of the filter. When designing a low pass filter the first step is to determine the number of component required to meet the design requirement. Usually a filter design has both a 3dB bandwidth (ω_c) and a desired stopband attenuation at a specified frequency (ω_s), and from these two parameters it is possible to determine the number of components or the order of the filter n .

Once the order of the filter has been determined, the next step requires a simple table lookup of the normalized component values as shown in Figure 2, [11] and can be directly read from the row corresponding to the filter order n .

TABLE 3-1. Butterworth Equal Termination Low-Pass Prototype Element Values ($R_s = R_L$)

Figure 2: Butterworth element table

The values in the table represent the component values required for a low pass filter having a 1 radian bandwidth and 1-ohm characteristic impedance. The final step is to scale the filter from the normalized format to one which will meet the design requirements for bandwidth and stopband attenuation. To do this the components are scaled from the normalized values to the appropriate values which will result in a low pass filter with the desired frequency response (ω_c) and system characteristic

impedance (Z_0). Equations (1), (2) show the relationship between the scaled component values L'_n , C'_n and the normalized component values L_n , C_n .

$$L'_n = \frac{L_n * Z_0}{\omega_c} \quad (1)$$

$$C'_n = \frac{C_n * Y_0}{\omega_c} \quad (2)$$

One of the more useful measurement metrics used to determine a filter's frequency response are S-parameters which are generally easy to measure and to use. The S-parameters for a two-port network in Figure 3 are defined to be the ratio of voltage leaving port i, ($b_i = V_i^-$) to the voltage incident on port i ($a_i = V_i^+$).

Figure 3: 2-port s-parameters

For the given two-port example in Figure 3 the s-parameters relationships are given by equations (3) through (6) and these fully characterize the two-port network.

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} = \left. \frac{V_1^-}{V_1^+} \right|_{V_2^+=0} \quad (3)$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} = \left. \frac{V_2^-}{V_1^+} \right|_{V_2^+=0} \quad (4)$$

$$S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0} = \left. \frac{V_1^-}{V_2^+} \right|_{V_1^+=0} \quad (5)$$

$$S_{22} = \left. \frac{b_2}{a_2} \right]_{a_1=0} = \left. \frac{V_2^-}{V_2^+} \right]_{V_1^+=0} \quad (6)$$

S-parameters are used extensively in filter design and each one gives unique information about the electrical characteristics for each port. For example, S_{21} given by equation (4), which is also known as the transfer function or insertion loss is strictly defined as the ratio of the voltage leaving port 2 to the voltage incident on port 1 while port 2 is match terminated. In keeping with the Butterworth low pass filter design example, Figure 4 graphically illustrates S_{21} that for signals impressed upon port 1 with frequencies in the passband, these frequencies are passed through from port 1 to port 2 generally unattenuated. For signals of higher frequency, it can be observed that S_{21} begins to decrease which indicates these signals are generally attenuated.

Figure 4: Butterworth low-pass filter $|S_{21}|$

For this work s-parameters will be used extensively to describe both simulated and measured results. They will also be used to compare and to evaluate the quality of filter results.

The procedure to realize reflectionless filters is by no means canonical. The theory and procedure for a reflectionless filter comes directly from Morgan and Boyd's paper in which the authors outline the steps required to design and realize this type of filter. The proposed methodology makes use of even-odd mode analysis, symmetry, duality and couples these basic concepts with Cauer ladder network topologies for filters to realize a unique topology for reflection-less filters. Symmetry follows from even-odd mode analysis and this allows the derived filter to have impedance matching at all ports, which implies that the cascading of these structures is at least theoretically possible without interactions between stages.

2.1 Even-Odd Mode Analysis

Even-odd mode analysis is based on Bartlett's bisection theorem [12] and uses symmetry and superposition to help analyze symmetrical circuits. The general idea is that for a symmetrical two-port network two independent types of port excitation (defined as even-mode and odd-mode) when applied to the circuit, certain circuit behaviors can be realized. Even-mode excitation is defined as impressing two signals which are equal in both magnitude and phase on each port simultaneously. Odd-mode excitation is defined as impressing two signals equal in magnitude but opposite in phase

on each port simultaneously. Even and Odd mode equivalent half circuits are sufficient to completely describe the behavior of a symmetric two-port network, but when coupled with the duality principle the dual of a network may not be unique. The topology derived for these filters is not unique as the general principles of the derivation may be applied in different ways to arrive at alternative topologies that have the exact same S-parameters.

In Figure 5a a symmetric 2-port network is shown. In Figure 5b, the network is divided on its symmetry plane and pulled apart exposing wires representing the central nodes of the circuit. First, even-mode excitation is applied to the two ports and because of symmetry no current will flow in any of the exposed wires, thus creating an open circuit which can be represented in Figure 5c. Similarly, when odd mode excitation is applied to the two ports the wires will be at virtual ground and the circuit can be represented in Figure 5d.

Figure 5: Two port representation of even and odd modes

For the even and odd mode circuits Morgan defines Γ_{even} and Γ_{odd} which represent the reflection coefficients for the respective equivalent circuits.

For the even-mode circuit

$$\Gamma_{even} = S_{11} + S_{12} \text{ and by symmetry } \Gamma_{even} = S_{22} + S_{21} \quad (7)$$

Similarly, for the odd-mode circuit

$$\Gamma_{odd} = S_{11} - S_{12} \text{ and by symmetry } \Gamma_{odd} = S_{22} - S_{21} \quad (8)$$

By rearranging, it can be shown that

$$S_{11} = S_{22} = \frac{1}{2} [\Gamma_{even} + \Gamma_{odd}] \quad (9)$$

$$S_{21} = S_{12} = \frac{1}{2} [\Gamma_{even} - \Gamma_{odd}] \quad (10)$$

The first and most important premise of reflectionless filters is that the reflection coefficients are ideally equal to zero which necessitates

$$S_{11} = S_{22} = 0 \quad (11)$$

$$\Gamma_{even} = -\Gamma_{odd} \quad (12)$$

Given that for the even-mode circuit,

$$S_{11} = \frac{z_{even} - 1}{z_{even} + 1} = -\frac{z_{odd} - 1}{z_{odd} + 1} = -\frac{1/y_{odd} - 1}{1/y_{odd} + 1} = \frac{y_{odd} - 1}{y_{odd} + 1} \quad (13)$$

where z_{even} is the input impedance of the even mode circuit and z_{odd} is the input impedance of the odd mode circuit. By direct inspection of equation (14)

$$Z_{even} = Y_{odd} \quad (14)$$

Equation (14) is the duality condition between even and odd mode circuits and will be important when realizing reflection-less filters. Another equally important result is obtained by substituting (12) into (10) giving

$$S_{21} = \Gamma_{even} \quad (15)$$

Examining equation (15) closely indicates that frequencies that are reflective in the even-mode circuit are transmissive in the two-port circuit. For example, if you have an even-mode circuit that is reflective at low frequencies such that $\Gamma_{even} \approx 1$ then for a two-port network the transfer function will be transmissive, i.e., $S_{21} \approx 1$ at low frequencies.

2.2 Filter Design Procedure

Given the important results from the previous section, a general procedure for realizing a reflectionless filter is presented next. To illustrate the procedure, a reflectionless low-pass filter (LPF) will be designed using results in equations (14), (15).

2.2.1 Step 1 – Even Mode Equivalent Half Circuit

For a reflectionless LPF, equation (15) implies the transfer function $S_{21} \approx 1$ and that the even mode equivalent circuit is reflective at these frequencies or $\Gamma_{even} \approx 1$. An even mode circuit which is reflective at low frequencies is a high pass filter shown in a ladder topology in Figure 6. Note that the symmetry line for an even-mode circuit represents an open circuit indicated by the dashed line.

Figure 6: Reflectionless design procedure: step 1

2.2.2 Step 2 – Odd Mode Equivalent Half Circuits

Next, using equation (14) which is the duality condition, draw the odd-mode equivalent circuit. Notice that series capacitors are replaced with parallel inductors and parallel inductors are replaced with series capacitors. Also, note that the symmetry line for an odd-mode equivalent circuit represents a virtual ground.

Figure 7: Reflectionless design procedure: step 2

2.2.3 Step 3 – Symmetrical Topology

The reflectionless filter should have symmetrical topology but upon inspection, the even and odd circuits do not appear symmetrical and the circuits start and end with different elements. Using virtual open for even, and virtual ground for odd mode equivalent half circuits we need to manipulate the circuits to achieve symmetry between the even and odd mode circuits. Component locations can be changed, and additional components added in order to create a symmetrical filter. Of course, any changes that support a symmetry must also not change the impedance or frequency response of the filter. In Figure 8, the positions of the trailing capacitor and resistor of the even mode circuit are changed. Additionally, the leading inductor on the odd-mode circuit is connected to virtual ground.

Figure 8: Reflectionless design procedure: step 3a

In Figure 9 a leading inductor is added to the even-mode circuit and connected to the virtual open. On the odd-mode circuit, a capacitor is added, but is shorted to GND. Both element additions have no effect on their respective circuits but will allow the construction of a symmetrical filter.

Figure 9: Reflectionless design procedure: step 3b

As can now be seen in Figure 9, the even and odd mode circuits are now symmetrical and the two halves of the circuits can now be combined to form a low pass reflectionless filter of arbitrary order. The final reflection-less low-pass filter topology of an arbitrary order with element values is shown in Figure 10. Note that Morgan and Boyd's procedure will only work with odd order Cauer networks [10].

Figure 10: Reflection-less low-pass filter of arbitrary order

2.2.4 Step 4 - Determination of Component Values

With the topology of the filter established, normalized component values now need to be determined, with values being guided by both symmetry and duality constraints.

Values are determined using the duality condition $z_{\text{even}} = y_{\text{odd}}$ (step 2 of the design process) and can be generalized for any characteristic impedance. Duality implies that inductor g_0 must be the dual of the capacitor g_1 . Determining component values to satisfy the duality condition we have for the first two components

$$L_0 = Z_o^2 C_1 \quad (16)$$

$$\frac{g_0 Z_o}{\omega_c} = Z_o^2 \frac{g_1 Y_o}{\omega_c} \quad (17)$$

Equation (17) is obtained by substituting the impedance of an inductor and capacitor element into equation (16). By inspection, equation (17) implies $g_0 = g_1$. Similarly, using duality, it can be seen that g_1 must be the dual of g_2 . Using this same reasoning

we can infer $g_0 = g_1 \dots = \dots g_k$ and $r=1$ [10]. Thus, we can assign all the components a value of unity without loss of generality.

2.3 Single Stage Nth Order Filters

Now that we have determined the topology of the filters and the element scaling, the topologies of a single stage $N=1, 3, 5$ order filters are shown respectively in Figure 11, Figure 12 and Figure 13. As can be seen an increase in order is given by an additional LC element pair on each port (four elements) of the filter.

Figure 11: Single stage $N=1$ topology

Figure 12: Single stage $N=3$ topology

Figure 13: Single stage $N=5$ topology

The response of single stage reflectionless filters of order $N=1, 3, 5$ can be seen in Figure 14. As the order of the filter is increased, the filter has a steeper cutoff region, but also results in higher out of band peaks. Because of symmetry and duality constraints in designing the filter it is not possible to choose different element values to try tuning or reducing the out of band peaks. As Figure 14 shows, only the third order filter can be deemed as optimal in as much as the out of band suppression is reasonable and can be shown to be 14.47 dB per third-order section [1]. Additionally [10], it can be shown that transfer function S_{21} of the third order reflectionless filter is equivalent to a Chebyshev Type II filter of order 3 with ripple factor $\varepsilon = 0.1925$ (see

Appendix F – Inverse Chebyshev).

Figure 14: Simulated $|S_{21}|$ for ideal 1st, 3rd and 5th order normalized reflectionless low-pass filters

To obtain higher order filters in practice, third order filters are simply cascaded which is possible since the filters are matched at each port and little interaction is expected between stages.

Chapter 3: Reflectionless Filter Realization - Simulation

Using the procedure described in chapter 2 this section will describe the design and simulation of both ideal and non-ideal filters based on the design constraints outlined in Morgan and Boyd's paper.

3.1 Design of Ideal Filter

Morgan and Boyd's paper outlines an anti-aliasing lowpass filter with the design parameters shown in Table 1.

Table 1: Anti-aliasing filter design requirements

3dB Bandwidth	188MHz
Out of Band - alias Suppression	60 dB

The 3dB bandwidth requirement will determine the L, C component values required to meet this specification. Each 3rd order section of the filter has a stopband rejection of 14.7 dB, so to meet the 60dB anti-alias suppression, four cascaded 3rd order sections will be necessary to meet this requirement. Filter element values for the lowpass filter can be determined directly from equations (18), (19) and (20) given by Morgan and Boyd when designing for the pole frequency (ω_p)

$$L = Z_o / \omega_p \quad (18)$$

$$C = Y_o / \omega_p \quad (19)$$

$$R = Z_o \quad (20)$$

However, it is often not useful to design a filter for the pole frequency; instead, a 3dB, 1dB or stopband frequency is specified. Because the 3rd order reflectionless filter transfer function S_{21} is identically an inverse Chebyshev, this relationship can be used to scale the filter to the appropriate target frequency see

Appendix F – Inverse Chebyshev for details. Since the filter design for this work requires four cascaded sections to meet the overall BW_{3dB} requirement at 188 MHz, this implies that each of the four cascaded sections must have an insertion loss of 0.75 dB per section at 188 MHz for an overall cascaded loss of 3dB. Using the 3rd order filter topology outlined in Figure 15, the filter requirements in Table 1 and equations (18), (19), (20) and the Inverse Chebyshev scaling factor which is used to scale from the pole frequency to the desired design target frequency, an idealized representation using lossless elements can be realized. Only three unique element values (one each of R, L and C) are required per stage and these values are listed in Table 2

Table 2: Ideal reflectionless element values per stage

L (nH)	C (pF)	R (ohms)
22.7176	9.08706	50

Figure 15: Third order low-pass filter section

3.2 Simulation of Ideal Low-Pass Filter

Once the filter element values have been derived, the next step is to setup a simulation testbench. Keysight's Advanced Design Software (ADS) is ideally suited to simulate such a filter as it allows the user to build all possible elements graphically and sweep the system in the frequency domain. To meet the stopband requirements, four sections of the filter are cascaded as shown in Figure 16. The testbench is setup to sweep the frequency domain in 1 MHz steps from 1 MHz to 10 GHz (a total of 10000 frequency steps), and the results are saved as S-parameters in a touchstone s2p format file.

Figure 16: ADS four section cascaded low-pass filter

The initial filter illustrated in Figure 17 below represents a single stage of the reflectionless LPF and is an idealized representation using lossless ideal elements whose values were determined in the previous section.

Figure 17: ADS reflectionless ideal low-pass filter topology

3.3 Ideal Low-Pass Filter Results

Using ADS to simulate and the ideal element values previously obtained, Figure 18 illustrates the results. Several things from the graph can be immediately noticed. First, the return loss is not zero, but around 180 dB. The reason for this is that the optimal element values have a default of 6 significant digits in ADS. To get the return loss to be ideally zero, an infinite number of significant digits would be required. Increasing the number of significant digits to 9 (order of magnitude increases by 3), the return loss increases by approx. 60 dB which is expected. The 3dB frequency of the filter is 188MHz as expected and the maximum stop loss is about 57.9 dB as opposed to the theoretical

maximum of 58.8 dB which again is due to the number of significant digits used in the simulation.

Figure 18: Simulated ideal results for $|S_{11}|$ and $|S_{21}|$ for 188 MHz low-pass filter

3.4 Simulation of Non-ideal Filter

The previous section simulated an ideal filter with both ideal components and lossless transmission lines and showed good agreement with Morgan and Boyd's results. To more accurately simulate the filter to be realized, both the lossy nature of the PCB materials, and the non-ideal nature of the filter components need to be considered. Using ADS software, a more robust simulation can be realized where the lossy material properties of the PCB and the microstrip transmission lines are included.

3.4.1 Non-ideal Components

Accurate models of multiple vendor passive components can be obtained from the Modelithics EXEMPLAR Library [13]. These library parts have several parameters which allow a range of values for substrate material dielectric and thickness, component values and pad sizes. Depending on the Sim_mode and Pad_mode settings chosen at simulation time, different pad or component parasitics will be included in the simulation. For all final simulations presented in this work (sim_mode, pad_mode) = (0,0) was chosen so that a full simulation including pad, parasitics and dielectric effects was performed. The components packages chosen for this work were either 0402 or 0603 standard surface mount passives. The smaller component package sizes were chosen as in general smaller packages have smaller parasitic effects. See Appendix D – Simulation Library for more information regarding the Modelithics library and the pads.

3.4.2 PCB Stack-up

The PCB stack-up and material used for this work is well beyond what is required for both the frequency of operation as well as the complexity of the circuit. The specific materials were chosen because the stack-up is familiar, well controlled and PCB panels are regularly run at Synopsys which allows the opportunity to include the test cards in a standard panel. The stack up is 12-layer low loss (dissipation factor of 0.0021 @ 10GHz) Tachyon-100G material and the material properties are shown in Figure 19.

Layer	Cu Thick. (mils)	Cu Foil wt (oz)	DK	Lam. Thick. (mils)	Description
1	2.10	.5 oz	3.04	3.25	Foil .5 oz
2	0.60	0.5 oz	2.98	5.00	Prepreg Tachyon 100G 1078(70.5) 18.25Gx24.25
3	0.60	0.5 oz	2.97	4.40	Core Tachyon 100G 5.00mils 2x1067 0.5 oz / 0.5 oz VLP2 18.25Gx24.25
4	0.60	0.5 oz	2.98	5.00	Prepreg Tachyon 100G 1035(75)/1035(75) 18.25Gx24.25
5	0.60	0.5 oz	2.98	5.00	Core Tachyon 100G 5.00mils 2x1067 0.5 oz / 0.5 oz VLP2 18.25Gx24.25
6	0.60	0.5 oz	2.97	4.40	Prepreg Tachyon 100G 1035(75)/1035(75) 18.25Gx24.25
7	0.60	0.5 oz	3.09	6.00	Core Tachyon 100G 6.00mils 2x1078 0.5 oz / 0.5 oz VLP2 18.25Gx24.25
8	0.60	0.5 oz	2.97	4.40	Prepreg Tachyon 100G 1035(75)/1035(75) 18.25Gx24.25
9	0.60	0.5 oz	2.98	5.00	Core Tachyon 100G 5.00mils 2x1067 0.5 oz / 0.5 oz VLP2 18.25Gx24.25
10	0.60	0.5 oz	2.97	4.40	Prepreg Tachyon 100G 1035(75)/1035(75) 18.25Gx24.25
11	0.60	0.5 oz	2.98	5.00	Core Tachyon 100G 5.00mils 2x1067 0.5 oz / 0.5 oz VLP2 18.25Gx24.25
12	2.10	.5 oz	3.04	3.25	Prepreg Tachyon 100G 1078(70.5) 18.25Gx24.25
					Foil .5 oz
			56.10		Thickness over Laminate
			60.30		Thickness over Copper
			61.30		Thickness over Soldermask

Figure 19: PCB Stackup

3.4.3 Microstrip Transmission Lines

For this design all the components are surface mount and microstrip transmission lines will be used with a 50-ohm characteristic impedance. Once the stack-up and dielectric properties of the PCB materials are known, the transmission lines can be designed. The microstrip transmission line dimensions depend on the calculator chosen and the

author's experience with the material, stackup and fabrication facilities. For this design a 50-ohm microstrip transmission line was determined to be 6 mils in width given the PCB material properties. Additional information regarding microstrip transmission lines in addition to the calculator used are in Appendix D – Simulation Library.

3.4.4 ADS Non-ideal Testbench

Pulling all the non-ideal components into ADS, the testbench now includes lossy PCB effects, lossy transmission lines, non-ideal components and component parasitics. As in the ideal testbench case, the simulation is swept from 1 MHz to 10GHz in 1 MHz steps. The component values used are no longer the ideal theoretical values but what is available for the component families chosen. For this work, the component families were from available libraries of the Modelithics software and which were available to be purchased and are shown in Table 3 with the comparative ideal values.

Table 3: Reflectionless 188MHz component values

	L (nH)	C (pF)	R (ohms)
Ideal	22.7176	9.08706	50
Non-ideal (5%)	23	9.1	49.9

[illegible]

Figure 20: ADS non-ideal filter section

3.5 Comparing Ideal and Non-Ideal Simulated Filter Results

To get an idea of the effects of non-ideal components and lossy PCB materials, the S-parameter results of both lossy and ideal reflectionless filter simulations are shown in Figure 21. Note that all measured results shown in this work have been de-embedded from the SMA launches – see Appendix C - Automatic Fixture Removal (AFR).

The Morgan and Boyd paper showed results up to 1.0 GHz, but for this work the bandwidth has been expanded to 5 GHz to see what effects occur above the immediate passband frequency. Firstly, the general shape of the results below 1.0 GHz is very similar between the two plots although as can be seen the return loss (S_{11}) is much more sensitive to the effects of non-ideal components and a lossy PCB. The Insertion Loss (S_{21}) is slightly higher below 1.0 GHz due to copper, dielectric losses and skin effect.

Figure 21: Simulated lossy versus ideal $|S_{21}|$ and $|S_{11}|$ results

For the results above about 1.0 GHz, there are some differences, the most obvious being the resonance peaks at 1.9 GHz and 2.25 GHz. These resonances are most likely from component parasitics and PCB pad capacitive effects and seems to have the effect of pushing the return loss S_{11} up around these resonances. Other than the resonant peak at 2.25 GHz the overall S_{11} shows greater than 10 dB suppression up to 5.0 GHz and 30 dB suppression in the stopband S_{21} .

Chapter 4: Filter Realization – Fabrication

In order to appraise the validity of the simulation results, a filter must be fabricated, and the frequency response measured in the lab. The design and fabrication process of the reflectionless filter designated the Device Under test (DUT) is outlined in the following sections.

4.1 PCB Schematic Design

Once the simulations have been completed and the results deemed acceptable, the first step in fabricating the filter is to create a netlist with associated component footprints and the software chosen to generate the netlist was Allegro Orcad Schematic Capture. A generalized component layout of each stage of the filter can be gleaned from the ADS testbench representation of a single stage of the ideal filter. The DUT will require four identical cascaded single stages. To evaluate and measure the filter performance over frequency a Vector Network Analyzer (VNA) will be the equipment of choice and the VNA interfaces to the DUT via SMA connectors. The DUT is a two-port device, which requires that the test card must have two SMA ports to allow for proper stimulation by the VNA. Figure 22 and Figure 23 illustrate the Orcad hierarchical schematic representation of the cascaded stages making up the complete filter.

Figure 22: Orcad schematic of Ideal Low-Pass Reflectionless filter Cascaded Sections

Figure 23: Orcad schematic of Ideal Reflectionless Low-Pass Filter Section

The Orcad schematic capture tool is used simply to generate a list of connections between components called a netlist. The components have an associated footprint which represent the physical pads which are used to attach the components to the PCB when soldered and are pulled into the layout tool when the netlist is imported. The

component values in the schematic are the same as in Table 3 but other than allowing a bill of materials (BOM) to be created, the component values have no other function.

4.2 PCB Layout and Assembly

The completed netlist is imported into a layout tool which is used to place components with respect to each other on the representative PCB and for this work Allegro PCB was used. The component layout was guided by several factors.

- 1) Surface mount parts of size 0402 and 0603
- 2) Microstrip characteristic impedance of 50 ohms
- 3) Microstrip transmission lines only to avoid vias in signal nets
- 4) Microstrip 2.92mm SMA launch

A layout resulting in all the above design guides is illustrated in Figure 24 with the overall PCB dimensions given in inches. The components are surface mount 0402 and 0603 body sizes and layout is relatively clean with a 2.92 mm SMA connector used as a launch. The specific connector used was an SVMicro SF1521-60070-1S-ND, with an optimized layout for the specific stack-up. The connector is a press-fit connector type which does not require soldering, but correct torqueing of the press-fit screws is required.

Figure 24: Reflectionless PCB layout

The fabricated assembled PCB with components is shown in Figure 25.

Figure 25: Assembled reflectionless low-pass filter PCB

4.3 Butterworth Diplexer

To evaluate the performance of the reflectionless filter and specifically the out of band performance, a conventional filter with similar low reflection out of band qualities was chosen to be used as a comparison by Morgan and Boyd. Specifically, a Butterworth Diplexer with similar passband and stopband characteristics as per Table 1 will be designed, fabricated and prototyped and the laboratory results compared to that of the reflectionless filter. A Butterworth diplexer of order $n=16$ is required based on [3], [4] and [9]. As can be seen in Figure 26 the diplexer is a three-port device, see Appendix A - Butterworth Diplexer. Also, of note is the de-embedding structure located to the right of the diplexer which will be used to de-embed the effects of the SMA connector launches – see Appendix C - Automatic Fixture Removal (AFR).

Figure 26: Fabricated Butterworth diplexer low-pass filter prototype

Chapter 5: Experimental Results

In the Chapter 3 a reflectionless low-pass filter was designed and simulated in ADS with simulation results showing good agreement with Morgan and Boyd's published results.

In Chapter 4 a physical prototype of the same filter was realized, fabricated and assembled. To further confirm the Morgan and Boyd's reflectionless theory, the prototypes must be measured and compared to simulated data and to theory.

To do a direct comparison between the ADS simulated results and the lab measured results, the PCB SMA launches must be de-embedded from the measurements and then the de-embedded measured results compared to simulation results. The de-embedding is an important step as no SMA model was obtained from the manufacturer, so no simulation could be run in ADS which would include the SMA connector effects which would be cumulative with the filter response. As can be seen in the ADS hierarchical schematic in Figure 16, no connector model was included in the simulation and the filter was terminated with ideal 50 ohm terminations at each port. Any measured results presented in this work will only be with the SMA launches de-embedded unless specifically noted. Appendix C - Automatic Fixture Removal gives more details regarding this de-embedding procedure.

To measure the filter response a Keysight N5227B PNA Microwave Network Analyzer [14] is the preferred equipment of choice. Before any measurements were undertaken, the VNA was calibrated using a standard SOLT calibration procedure – see Appendix B -

VNA Calibration for details. Figure 27 shows the results of the comparison between ADS simulation and lab measurements of 188 MHz LPF and excellent agreement is evident. The experimental results are also in good agreement with Morgan and Boyd's results as can be seen when comparing Figure 27 and Figure 28.

Figure 27: 4-stage 188MHz measured versus simulated $|S_{21}|$ and $|S_{11}|$ results

Figure 28: Morgan and Boyd's measured versus simulated results (Fig. 17 from [2])

Morgan and Boyd's results were limited to about 1 GHz, and to further investigate the out of band results, the simulation and measured results are shown out to 5 GHz in Figure 29. The agreement between simulation and measured is still very good out to 5 GHz, but there is a slight shift in resonance between the simulated and measured results.

Figure 29: Measured versus simulated $|S_{21}|$ and $|S_{11}|$ results - 5GHz

With good agreement established between simulation, experimentation and published results, the next step was to compare the Butterworth diplexer results with the reflectionless filter results. Morgan and Boyd show no direct comparison of the out of band suppression between these two filters, but they do show the passband comparisons between the filters. The comparison of Butterworth diplexer and reflectionless LPF designed in this work is shown in Figure 30. The reflectionless LPF displays superior responses for S_{21} except the transition region is initially less steep. Similarly, the S_{11} response for reflectionless LPF is superior to Butterworth diplexer across almost the entire frequency range.

Figure 30: 188MHz measured reflectionless versus diplexer $|S_{21}|$ and $|S_{11}|$ results

Next, we visually compare this work's results shown in Figure 30 with Morgan and Boyd's results shown in Figure 31. Results for the insertion loss S_{21} in Figure 30 and the "Gain" curve in Figure 31 show a very good agreement for both the diplexer and the

reflectionless filter. This agreement also confirms that our diplexer design and implementation is very close to that reported by Morgan and Boyd.

Fig. 12. Insertion loss and complex gain slope for three potential antialiasing filters.

Figure 31: Morgan and Boyd's diplexer versus reflectionless results (Fig. 12 from [2])

Another important feature of the reflectionless filter is the symmetry property. The filter is a two-port device and because of symmetry the return loss at both ports is theoretically identical. This allows these filters to be cascaded with other filters or with other sensitive 50-ohm components with little interaction. Figure 32 demonstrates that reflectionless filter indeed has symmetrical response and that both ports S_{11} and S_{22} have very close to symmetrical return loss. Without the resonance at 2.25 GHz the reflectionless filter would have S_{11} better than 10 dB out to 5 GHz. The diplexer is

matched for S_{11} (common port) relatively well but S_{22} shows almost unity reflection in the stop band. Any noise above the passband frequency on diplexer port 2 will tend to reflect from the filter back to the load rather than being absorbed with possible resonance ramifications between the filter and any connected circuits.

Figure 32: 188MHz measured reflectionless versus diplexer return loss $|S_{11}|$ and $|S_{22}|$ results

5.1 Frequency Scaling

Morgan and Boyd's results were for a 188MHz Low Pass filter and as such the frequency range of interest was up to 1 GHz. To investigate the response of the reflectionless filter topology to higher frequencies, the original filter design 3dB bandwidth was increased by a decade to 1.88 GHz. Of specific interest was how well the return loss would behave out to 10 GHz. As with any of the classical filters, the reflectionless filter can be scaled using standard scaling techniques. The simulated and measured results for this frequency scaling are shown in Figure 33 . The measured and simulated responses show very good agreement up to 10 GHz.

Figure 33: Reflectionless 1.88GHz LPF measured versus simulated $|S_{21}|$ and $|S_{11}|$ results

There are two notable resonances in the measured S_{21} response at 5.75 GHz and 6.65 GHz. They are approximately 25 dB suppressed in the stop band, but they seem to have an effect of pushing the Return loss to above 5dB. A possible reason for this could be a component package or pad resonance effect.

To further compare the response quality of the scaled reflectionless filter, a scaled Butterworth diplexer was also designed and measured in the lab with the responses compared in Figure 34. The reflectionless filter shows overall about a 10-dB

improvement in S_{11} in the stopband region compared to the diplexer but has higher mismatch at lower (less than 500 MHz) frequencies which was confirmed in simulation.

Figure 34: 1.88 GHz measured de-embedded reflectionless vs diplexer $|S_{21}|$ and $|S_{11}|$ results

Chapter 6: Discussion, Conclusions and Future Work

6.1 Discussion

The basic undertaking of this work was conceptually relatively straight forward. Firstly, we set out to understand Morgan and Boyd's reflectionless filter design procedure. Secondly, we verified the results of this procedure by measuring realized filter prototypes in the lab. In the process of this work several unexpected issues arose while other issues that at first were expected to be difficult proved to be simpler than originally anticipated.

The procedure outlined by Morgan and Boyd to design a reflectionless filter required some knowledge of duality, symmetry and even-odd mode analysis. Using the preceding theory and stepping through the procedure was relatively intuitive as presented by Morgan and Boyd. The design of the diplexer turned out to be more of a challenge conceptually. With common filter parameters defined for both filters and the component values designed initially on spreadsheets, the next step was to run simulations of the resulting filters. Using ADS, simulations were run on ideal filters of both types to confirm the desired attributes (3dB bandwidth, stopband attenuation). The next step was to add non-ideal components and lossy PCB materials. One of the advantages of the reflectionless filter that became very apparent early in the design process was the much smaller number of unique components required to construct the filter when compared to the analogous Butterworth diplexer. For the reflectionless

filters in this work, there were three unique component values (one each of L, C, R) required versus fully 32 unique component values for the similar Diplexer filter. Because so many unique component values were required for the diplexer and given that there are only a limited number of available component values, some of the component values chosen for diplexer differed by more than five percent from their theoretical values. The element values chosen were confined to be from family simulation libraries available in Modelithics and from components that were readily obtainable. Given more time and effort, components with less than 5% deviation from the ideal might have been obtained.

The parasitic effects of the components were not unexpected but the accuracy of the simulations in showing these effects using the Modelithics models resulted in the simulation and measured results matching very well for all filters. The primary parasitic effect was the component pad capacitance, which unfortunately was exacerbated by the chosen stackup. The stackup used is of a much higher quality than is required for the frequency range used in this work and was chosen because of the author's familiarity and access to the material. The number of layers in the stackup was 12, but this specific work only requires a standard two-layer stackup as the components used were surface mount and the transmission lines were microstrip located on the top layer only. Because of this, only the top two layers of the stackup were used with the second layer being used as the microstrip reference. This means that the dielectric thickness of only 3mils is present between the microstrip and the reference plane. This relative thinness of the

dielectric material exacerbated the parasitic pad caps of the components. For example, to obtain a 50-ohm microstrip impedance with this dielectric material and thickness, a 6 mil microstrip line was required. The pads for the 0402 components were approximately 20 mils square. By just increasing the thickness of the dielectric from 3 to 30 mils the parasitic capacitance would decrease by a factor of 10.

The 3-dB frequency of the reflectionless filter - both measured and simulated - was 172MHz but the initial target frequency was 188MHz using ideal components and no PCB effects. This lower frequency was not an unexpected result due to the lossy nature of the PCB and the components and the precision of the components which was 5%.

The reflectionless filter was scaled using standard frequency scaling techniques and compared to a similarly scaled Butterworth diplexer. The scaled reflectionless filter was found to have good return loss over the measured frequency range and was comparable to the Butterworth diplexer.

Because of symmetry, the reflectionless filter showed good return loss at both ports across all frequencies investigated. Also, the return loss of the reflectionless filter in the stopband was superior to the diplexer over all frequencies investigated. Additionally, the diplexer only had good return loss characteristics at the common port whereas the other ports had poor return loss which precluded the diplexer from being cascaded. The symmetry of the ports of the reflectionless filter is one of its best features which allows

the filter or multiple reflectionless filters to be inserted in circuits without interacting unpredictably with circuit elements on either port.

6.2 Conclusion

The main goals of this work to validate and to verify the novel filter design procedure outlined in Morgan and Boyd's paper were achieved and a low pass reflectionless filter prototype was realized. A classical complimentary Butterworth diplexer prototype was also realized and used to evaluate the quality of the return loss characteristics for the reflectionless filter. The filter prototype measurement results obtained for both filters show the reflectionless has improved return loss characteristics to the diplexer over a wide frequency range.

In summary, this work addressed

- reflectionless filter design procedure
- frequency scaling property of reflectionless filters
- good matching between simulation and measured response for reflectionless filters
- return loss of both ports on reflectionless filter were superior to classical design

6.3 Future Work

This work was an introduction into reflectionless filters and allowed for the replication, verification, and frequency extension of Morgan and Boyd's novel ideas. However,

several additional areas of investigation could be done which could build directly on this work.

1. Determining the sensitivity of component values in reflectionless filters would allow the designer to specify the tolerance of values that could be used in a filter without radically changing the performance of the filter.
2. Minimizing the effects of the parasitic pad capacitance could be done in several ways. The first method would be a direct continuation of this work and would use the same substrate and materials. The idea would be that the reference plane directly beneath the component and pads would be voided and as a result, the reference layer would then become one of the deeper layers. This would have the effect of increasing the dielectric thickness and decreasing the parasitic pad capacitance. However additional GND vias surrounding the component would be needed to tie the reference planes. The second method would be to use a standard two-layer PCB and simply the increased thickness of the dielectric material would mitigate some of the pad capacitance effects.
3. Investigating how high in frequency one can push these filters when using discrete components. The component values will scale with frequency but most likely a different family of components will need to be used.
4. Use a prototype in a system which has a source that is sensitive to out of band reflections and determine if the interaction is mitigated by using a reflectionless filter [15].

5. Cascade multiple reflectionless filters such as a low pass and a high pass to obtain a bandpass filter to demonstrate the ability of the reflectionless filter topology to allow concatenation of filters [16].

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Appendix A - Butterworth Diplexer

A diplexer is a passive device that implements frequency-domain multiplexing and can be used for dividing a broad frequency band into smaller bands using selective filters. It is fundamentally different from a passive combiner or splitter in that the ports are frequency selective. This separation of desired frequency bands, however, is a much more difficult task than appears at first sight. If filters are simply connected in a parallel manner, interaction between the inputs of each filter will degrade the overall performance of the diplexer unless the filters are carefully designed. For the purposes of this study, a dual diplexer formulation is considered where one constructs a low pass (passband) and high pass (stopband) filters to have complimentary input admittances and the filters are then connected in parallel. The input admittance of the diplexer is $Y_p + Y_s$ where Y_p and Y_s represent the input admittances of the passband and stopband filter respectively. A matched filter requires $Y_o = Y_p + Y_s$ which means that the sum of the input admittance of the lowpass and high pass filter is real and constant for all frequencies. Filter pairs that satisfy this condition are termed complimentary.

Figure A.1: ADS Butterworth diplexer testbench

As with the reflectionless LFP filter, the Butterworth diplexer was first simulated in ADS to verify the initial design, as shown in Figure A.1. As mentioned, of specific note for this type of filter is the input admittance Y_{11} looking into the common port must be strictly real. Figure A.2 is the ADS simulated Y_{11} results showing real and imaginary terms. As can be seen the real admittance is 0.02 siemens which corresponds to an impedance of 50 ohms.

Figure A.2: Diplexer common port Y_{11} input admittance

To help clarify the complimentary nature of the input admittance, the common port is separated into a high pass and low pass legs and the admittance measured for each leg. Figure A.3 and Figure A.4 show the complimentary nature of the admittance for each path. When added in parallel, the real components add to a constant value of 0.02 siemens and the imaginary components ideally add to zero.

Figure A.3: ADS diplexer real admittance

Figure A.4: ADS diplexer imaginary admittance

Once the concept has been verified, the filter is then constructed using Orcad and Allegro PCB layout tools. The Butterworth schematic is shown in Figure A.5 and shows a Cauer ladder topology common to many classical filters.

Figure A.5: Butterworth diplexer schematic

The Butterworth design called out for 32 unique elements, but in some cases a common available element sufficed for several of the ideal values. There were a handful of elements that had greater than 5% difference between available and ideal values and they are highlighted in Table A.1. The part family chosen had a 5% tolerance on values.

Table A.1: Butterworth diplexer BOM

Capacitor	Ideal (pF)	Actual (pF)	Inductor	Ideal (nH)	Actual (nH)
C1	31.97	33	L1	66.38	68
C2	31.56	33	L2	80.59	82
C3	28.92	30	L3	76.04	82

C4	25.02	24	L4	67.77	68
C5	20.13	20	L5	56.73	56
C6	14.46	15	L6	43.45	43
C7	8.217	8.2	L7	28.48	27
C8	1.66	1.7	L8	12.41	12
C9	10.8	11	L9	22.41	22
C10	8.893	8.9	L10	22.71	23
C11	9.425	9.4	L11	24.79	24
C12	10.58	11	L12	28.64	30
C13	12.63	13	L13	35.59	36
C14	16.49	16	L14	49.57	51
C15	25.16	24	L15	87.21	82
C16	57.77	56	L16	431.8	390

Appendix B - VNA Calibration

For all measurements in this work a Keysight PNA 5227B Vector Network Analyzer was used. It has a sweep range of 10 MHz to 67 GHz, but for this work the sweep range was chosen to be 10 MHz to 10 GHz in 1 MHz steps for a total of 9991 frequency points. A standard SOLT calibration was done just before measurements were taken using the Keysight Electrical Calibration module N4964D. For all measurements the IF bandwidth was set to 100 Hz to maximize the SNR. All SMA connectors were torqued as per the standard using the calibrated torque wrench.

Figure B.1 Keysight VNA N4964N calibration module

Appendix C - Automatic Fixture Removal (AFR)

The prototype designs included a 2.92mm SMA launch and 500 mils of microstrip on each port of the filters which was necessary to allow a measurement interface to the filters. An ADS model of the SMA connectors was not available and so the simulations were done without the SMA launch and 500 mils of microstrip. To remove the effect of the connectors and extra microstrip from the filter measurements, a test structure was designed that would allow for the de-embedding or removal of the SMA interface on each port. Figure C.1 shows the two launches and the device under test (DUT). Figure C.2 shows the diplexer filter and the de-embed structure or AFRx2.

Figure C.1: Reflectionless filter with port mappings

Figure C.2: Diplexer filter with AFR structure

This de-embed structure is equivalent to two SMA launches connected by 1000 mils of microstrip and is located on the diplexer prototype PCB. The test structure is two back to back SMA launches. The S-parameters of the structure is measured (afrx2.s2p) and then using Keysight's Automatic Fixture removal (AFR) software the structure is bifurcated into two identical S-parameter files (afr_1.s2p and afr_2.s2p) representing one SMA connector and 500 mils of microstrip. It is these files that are then de-embedded from the measured filter S-parameter files.

Figure C.3: AFRx2 prototype

Figure C.4: AFRx2 S_{21}

Appendix D – Simulation Library

Component footprints – various vendors have slightly different recommended footprints based on board density and assembly. The Modelithics component models had several supported pad sizes that are recommended by the component manufacturer and supported by the model. To keep the layout simpler one set of pads was used for all 0402 capacitor, inductor and resistors. The 0402 and 0603 pad dimensions are illustrated in Figure D.1. All pad dimensions are in mm. When simulating in ADS, the user can choose to include the effects of the pad capacitance or exclude them for a simpler simulation.

Figure D.1: Pad Dimensions

The system impedance is targeted at 50 ohms and given the layer thickness, dielectric constant the microstrip transmission line width can be determined. There is a plethora of calculators online and in software tools, and in this user's experience many are erroneous or do not clearly state under what physical conditions they can be used. The calculator used was found at <https://www.microwaves101.com/calculators/1201-microstrip-calculator> and is based on David Campbell's calculator.

Figure D.2: Microstrip calculator

Additionally, because the stackup and material are well known to this author, the microstrip width was known a-priori to be 6mils. The calculator gave 6.4 mils to achieve 50-ohm characteristic impedance.

Table D.1: PCB material properties

Microstrip Copper Thickness t (mils)	2.1
Dielectric height h (mils)	3.25
Dielectric constant ϵ	3.04
Microstrip Width w (mils)	6
Characteristic Impedance Z_0 (ohms)	50
Loss Tangent	0.0021

Appendix E – Duality

Duality is an important concept and it is one of the base foundations of reflection-less filters. The concept of duality can be traced back to the dual nature of electric and magnetic fields proposed by Maxwell, but the first use in circuit theory can be attributed to Russel [17].

Duality defines that there exists a list of dual relationships, that can be interchanged in an expression with the result that the dual expression is equivalent to the original one.

Table E.1: Table of duals

Voltage	Current
Parallel Circuits	Serial Circuits
Resistance	Conductance
Impedance	Admittance
Capacitance	Inductance
Reactance	Susceptance
Short circuit	Open circuit

Dual conversions used in this work are undertaken with normalized impedance circuits. To scale the impedance of the networks the following relationships can be used, where L' , C' and R' are the duals of C , L and G

$$L' = (Z_0)^2 C$$

$$C' = (Y_0)^2 L$$

$$R' = (Z_0)^2 G$$

Appendix F – Inverse Chebyshev

Morgan and Boyd outline how the third order filter shares the same transfer function as a third-order Inverse Chebyshev filter with ripple factor $\varepsilon = 0.1925$. Specifically, the relationship for the reflectionless filter is

$$s_{21} = \Gamma_{even} \quad (F.1)$$

What this implies is that the transfer function of the Chebyshev type II can be used to scale the reflectionless filter element values to desired frequencies for the stopband, passband, zero frequencies where ω is the frequency of a specific IL requirement.

$$|H(j\omega)| = \frac{1}{\sqrt{1 + \varepsilon^{-2} T_n^{-2}(\omega^{-1})}} \quad (F.2)$$

This relationship is what was used in this work to determine the component values based on frequency requirements of the design. Specifically, the design in this work called out for a 3dB passband of 188 MHz, which when coupled with a 4-stage filter, required 0.75 dB of insertion loss per stage for a total design target of 3dB.