MICROWAVE FILTER DESIGN: COUPLED LINE FILTER

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MICROWAVE FILTER DESIGN: COUPLED LINE FILTER

A Project

by

Michael S. Flanner

Spring 2011

APPROVED BY THE DEAN OF GRADUATE STUDIES
AND VICE PROVOST FOR RESEARCH:

_________________________________

Katie Milo, Ed.D.

APPROVED BY THE GRADUATE ADVISORY COMMITTEE:

_________________________________

Adel A. Ghandakly, Ph.D.
Graduate Coordinator

Ben-Dau Tseng, Ph.D., Chair

_________________________________

Adel A. Ghandakly, Ph.D.
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Radio Frequency (RF) filters operating in the microwave frequency range are needed for applications including wireless and satellite communications as well as military applications. These applications demand high performance filters that will contribute as little as possible to a system’s size and cost. Advances in materials used to construct these filters have played a significant part in meeting these demands. Planar and dielectric resonator filters are among the filter types which benefit from higher quality dielectric materials. Planar, or printed circuit board (PCB) based filters are popular and relatively practical to design. This paper presents the design and test of a planar coupled line filter constructed from relatively high quality dielectric material.

The bandpass coupled line filter presented here is specified to have a midband at 1.69GHz and bandwidth of 0.169GHz. Passband insertion and return loss is
specified to be $<5\text{dB}$ and $>10\text{dB}$ respectively. The design was derived from standard filter design theory and formula available in the literature. An optimized computer aided (CAD) design was also generated for comparison. The ‘Microwave Office’ design software was provided by Applied Wave Research Inc., operating with an educational license. Both formula and simulation based designs had nearly identical physical structure and performance under simulation. A prototype of the design was manufactured and tested on the bench using an Agilent 8714ES RF Network Analyzer (3.0GHz). The measured passband insertion loss was $<3.6\text{dB}$, meeting the specified goal and consistent with expected response based on simulation.

Placement of the filter’s midband was offset from the expected value. This was most likely a result of wide tolerance in the dielectric permittivity specified for the PCB substrate. Also the filter’s bandwidth was wider than expected. Possible causes might have been the test equipment calibration, impedance mismatch amongst the measurement system’s cables and adapters, or trace impedance errors resulting from structural defects in the etched microstrip lines.

A coupled line RF filter was designed and working prototypes of the design performed well. Passband insertion loss measured well within the target specification. The filter owes its low signal loss to refined formula published in the literature as well as to the availability of the high quality PCB material sampled by Rogers Corporation. AWR’s simulation software was also helpful in the design process. The simulation software saved design time and effort by allowing pre-production verification of the
design. The software also allowed various design iterations to be explored much more quickly than would have otherwise been possible.

Differences between simulated and prototype performance likely resulted from material specification tolerance, possible errors in transmission line trace structure, as well as other possibly measurement related factors. Measured passband insertion loss was less than 3.6dB, exceeding the target spec of 5dB maximum. Measured midband was 1.61GHz, showing an offset from the target 1.69GHz. Measured bandwidth was 206kHz, which was wider than the target 169kHz.

Performance could be further improved with refinements to the design and the construction material. The design equations could be further optimized by taking into account field fringing effects and by obtaining precise material specs from the PCB manufacturer. Redoing the filter on substrate with lower specified dissipation factor (i.e. higher quality factor) could further reduce signal loss. The material used here had a quality factor approximately double that of standard FR-4 board material. Even higher quality material is available.

The result should be an excellent and practical filter which can be designed without the need for expensive CAD tools.
CHAPTER I

INTRODUCTION

This paper describes the design of a planar bandpass microwave filter. The filter specifications are bandpass with 1.69GHz midband and 0.169GHz bandwidth. A passband insertion loss of <5dB and return loss of >10dB is desired. These specifications are intended to make the design of the filter practical. A filter used in a wireless communication system may demand a narrower passband with <1dB passband insertion loss.

The design is intended to demonstrate some issues facing modern RF filter designers. RF filters are widely used in applications such as wireless handset and base stations, satellite receivers and military applications. These applications are increasingly demanding because of crowded spectrum, higher performance specifications and tougher operating environments. In addition to higher performance, the industry continues to demand reductions in size and production cost.

It is possible to design a high performance filter with a sophisticated planar or waveguide architecture. However, these designs can involve significant complexity. The challenge here will be to improve performance of a relatively simple planar printed circuit board (PCB) based design. Dielectric substrate materials used in the construction of PCBs can significantly contribute to loss of desired signal spectrum. Higher quality
substrate materials contribute less signal loss, creating an opportunity to achieve a high performance filter without additional design complexity.

The architecture demonstrated here is a planar ‘coupled line’ type filter, since this is among the most practical and common filter types which can meet the stated specifications. The filter response will be based on the Chebychev transfer function. Chebychev type filters are popular for their high selectivity, i.e., they have a relatively fast signal cut off between pass and stop band. Filters operating in gigaherz frequency ranges rely on distributed transmission line structures to obtain the desired frequency response. Dimensions of the coupled transmission lines can be derived with published formula or minimal simulation software capability.

Other PCB based architectures such as interdigital and hairpin were investigated. However, the effectiveness of these architectures was uncertain or the design seemed to carry much greater complexity. Even higher performance filter types such as non-planar waveguide filter designs have been well established. Cost and manufacturing complexity would likely be significantly higher for these filters.

A basic coupled line filter, using high quality material, can achieve the desired specification. The low cost, ease of design and good performance will provide a helpful example of modern RF filter design.
RF filters operating in the microwave spectrum have a range of applications, including wireless handset and base stations, as well as satellite receivers and military applications.

Recently published papers investigating RF filter design reveal a need for ongoing development. These filters operate in an increasingly crowded signal spectrum. They may operate in harsh environments subject to shock, vibration, and extreme temperatures. The industry experiences continuous pressure to improve performance while reducing the size and cost of the filters and their associated systems. Improvements in the electrical properties of available materials are helping meet these demands. PCB materials with higher dielectric constants yield smaller filter structures. More importantly, filters using materials with high quality factor will have proportionally less midband signal loss. As a result, smaller, higher performance filters are becoming easier to design.

A review of some filter design basics will be helpful. Ludwig and Bretchko [1] provides a textbook overview of filter basics. Filter types, theory and design are reviewed. Filter types include Butterworth, Chebychev and Linear filters. Chebychev filters are seen to have an advantage of fast transition between midband and out of band signal spectrum. This transition is relatively slow for Linear filters, which are favored for
their relatively linear phase across the filter's spectrum. Higher phase linearity implies less variation in signal (group) delay with frequency and therefore less potential distortion across a filter's passband [1], [2].

Lumped element filters, i.e., filters constructed with discrete components such as capacitors and inductors are noted to be unsuitable for filter construction above certain frequencies, with 500MHz given as a guideline. For higher frequency filters, lumped element filter models can be converted to distributed structures. A PCB transmission line segment is an example of a distributed structure element. The length of such a segment will be proportional to filter midband wavelength.

Ludwig and Bretchko provide an overview of the design process with a coupled line filter example. The author advises a software simulation program will be needed to tune the design [1]. Other papers provide needed formula to produce a highly refined coupled line filter without simulation software [3], [4], [6].

Recent journal articles review important issues facing RF filter designers. Filters with minimal signal loss in their passband are noted to be critical in modern systems [5], [12]. Several factors compel this requirement. A receiver may need to detect signals which are weak relative to environmental noise and undesired signals. Transmission channels require minimum attenuation at the pre-amp filter stage in order to work efficiently. For these reasons, filters used in cellular base stations are noted to typically require less than 1dB of midband loss.

A filter's midband signal loss is inversely proportional to its unloaded quality factor (Qu). This parameter represents the ratio of reactance to resistance of an unloaded circuit or circuit element. Physically larger filters therefore generally have higher Qu due
to lower cross section resistance [5]. Overall quality factor of a filter will be determined by the quality factors of its dielectric material and it’s conductors, as well as the quality factor associated with the circuit’s radiated emissions [13].

A distributed filter constructed on PCB material using common FR-4 substrate may have a typical quality factor on the order of 50. This may imply an unacceptably poor midband signal loss of greater than 5dB. However higher performance PCB materials with dielectric quality factors well over 100 are available. PCB dielectric quality factor is commonly listed on the datasheet as it's inverse which may be referred to as 'dissipation factor', 'loss tangent' or 'tan-delta'.

Formula for midband signal loss of a Chebychev filter will be useful when designing such a filter to specifications. The approximate formula shown here is based on the quality factor of individual stages, number of filter poles, as well as the filter’s frequency response specifications [5]. Each stages quality factor will result from the combined losses in the dielectric material and the conductors as well as losses associated with radiated emissions.

\[ LA \approx 8.686[N-1.5]f_0/(dF*Qu) \]

\( LA \) = midband insertion loss in dB

\( N \) = number of filter poles

\( f_0 \) = midband frequency

\( dF \) = filter bandwidth

\( Qu \) = unloaded quality factor of the filter stages
The typical quality factor of common filter architectures is compared in Table 1, and some representative insertion loss resulting from various quality factors are given in Table 2 [5].

### TABLE 1
**COMMON FILTER TYPES AND THEIR ASSOCIATED QUALITY FACTOR**

<table>
<thead>
<tr>
<th>RF resonator type</th>
<th>Quality Factor @ 5 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lumped Element</td>
<td>10-50</td>
</tr>
<tr>
<td>Microstrip</td>
<td>50-200</td>
</tr>
<tr>
<td>Coaxial (TEM) resonators</td>
<td>200-5,000</td>
</tr>
<tr>
<td>Dielectric Resonator</td>
<td>1,000-10,000</td>
</tr>
<tr>
<td>WaveGuide</td>
<td>1,000-50,000</td>
</tr>
</tbody>
</table>


### TABLE 2
**FILTER MIDBAND INSERTION LOSS VS. THE FILTERS UNLOADED QUALITY FACTOR (Qu)**

<table>
<thead>
<tr>
<th>Unloaded quality factor</th>
<th>~Insertion loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10,000</td>
<td>0.3</td>
</tr>
<tr>
<td>1000</td>
<td>0.5 to 1.0</td>
</tr>
<tr>
<td>300</td>
<td>1.4 to 2.3</td>
</tr>
<tr>
<td>100</td>
<td>4.0 to 6.0</td>
</tr>
</tbody>
</table>

Source: Adapted from I. Hunter, R. Ranson, A. Guyette, and A. Abunjaileh, “Microwave filter design from a systems perspective,” *IEEE Microwave Magazine*, vol. 8, no. 5, pp. 71, Oct. 2007.
A cellular base station may require a filter's midband loss to be <1dB. Referring to the tables this may require a filter quality factor of greater than 1000. Dielectric resonators are among the filter types which meet or exceed this requirement. Dielectric resonator technology also has the advantage of low loss, small size and good temperature stability. These advantages make this type of filter popular for satellite, base station and mobile communication systems [9].

The dielectric permittivity constant (Er) of a dielectric resonator or planar filter’s core material will inversely affect its size. Filter or resonator elements for these filters are proportional in size to midband signal wavelength. The filter elements size will therefore be inversely proportional to the square root of the material’s effective Er [9].

Modern filter designers will typically have to accommodate a number of specifications and constraints. Midband insertion loss will likely be important as well as other electrical performance specs, including linearity and power handling capability. There may be mechanical specifications and constraints for size and mass, as well as operating environment requirements including temperature, vibration and moisture tolerance. Design may be constrained by the availability of manufacturing technologies such as machined or etched capability [8], [9], [10], [11], [12].

Tuning of filters in production is noted to be a common requirement to meet specifications. This process is noted to be potentially expensive, time consuming and to require specialized skills [8]. Filters with a reconfigurable frequency response may simplify the production tuning process. Reconfigurable filters also provide dynamic filter response adjustment, a valuable feature to cope with a crowded UWB spectrum for example [10].
Some recent papers investigate reconfigurable filters [10], [11], [12]. These filters allow dynamic selection of filter midband, preferably with minimal affect on other response characteristics such as bandwidth. Planar filters are noted to be convenient for designing reconfigurable filters because they are easy to produce and physically small. However the addition of tuning elements such as varactors and RF-MEMS switching technology further degrades their linearity and their midband signal loss [10]. High-Q tunable dielectric resonators are noted to meet the requirements of modern systems and offer adequate performance margin to accommodate the degradation in quality factor caused by the addition of tuning elements [9].

Dielectric resonators commonly consist of a cylindrical element of high Q dielectric material. This element is configured to operate in a selected waveguide mode. For a desired fundamental resonant frequency, the size of the element is noted to be approximately one signal wavelength as measured within the dielectric material [9]. Higher Er specifications seen for these materials result in proportionally shorter signal wavelength and therefore physically smaller filters.

Dielectric resonator filters have advantages in size and performance over planar filters, however they appear to be significantly more difficult to design and produce. Other types of filters such as machined waveguide filter may have advantages in performance, power handling and harsh environments. It can be inferred at this point that size and production issues, such as a requirement for machined parts, make this approach generally less popular for many applications.

A planar implementation of a Chebyshev type passband filter is among the more common, basic microwave filters in use. Such filters are among the more practical
to produce. Producing and testing such a filter will demonstrate the performance benefits of currently available materials.

The intent of this project is to produce a microwave filter which takes advantages of these advancements to meet a tight performance specification. Published papers and available simulation software make design of a highly optimized coupled line filter possible on a student budget. Samples of high quality PCB material should be available. These resources make it practical to establish a baseline for high quality, small size, low cost and practical design for a modern microwave filter.
CHAPTER III

DESIGN THEORY AND EXAMPLE

The coupled line filter design is derived from published coefficients [1]. The selected coefficient set represents the normalized, low pass Chebychev filter with the desired attenuation roll-off characteristic. Each coefficient represents the inductance or capacitance of a discrete (lumped) component element of this low pass filter. These coefficients can be transformed to produce a passband filter with the desired midband and bandwidth. The resulting passband filter consists of sequential LC resonator pairs alternating between series and shunt configuration. The resonant frequency of each LC pair equals the filter midband frequency (e.g., $f_0 = 1.69\text{GHz}$).

As discrete components become comparable in physical size to signal wavelength their performance deteriorates. At frequencies above a few hundred megahertz discrete components will increasingly need to be replaced with distributed structures. Published equations [1], [3], [4], [6] can be used to convert filter coefficients to distributed structure dimensions. The distributed structure for this coupled line filter design will be composed of differential pairs, i.e., two copper (microstrip) traces running parallel to each other across the surface of a two layer PCB. This structure should yield excellent performance in the desired gigahertz frequency range.
Choice of Filter Type and Order

A good passband filter has minimal signal loss in its passband, as well as a narrow passband with as much out of band attenuation as possible. Chebychev filters have narrower passband response in trade for more ripple in the passband. For a very good bandpass filter the ratio of rejection bandwidth (at -60dB) to passband width (at -3dB), called the shape factor, is 2 or less [5]. Higher order filters can have a narrower shape factor but will be physically larger.

The filter order will be chosen to achieve the desired bandwidth (0.169GHz) while minimizing the physical size of the filter. The filter specification goals for return loss (scatter parameter S11) are >10dB and for insertion loss (scatter parameter S21) <5dB. Simulations showed a filter order of N=3 will achieve this goal.

The required order for a filter meeting the given specifications, with a shape factor of 5, is calculated below; it indicates that the expected shape factor for a third order filter will be somewhat wider than 5:

\[ N \geq \frac{LA + LR + 6}{20 \log\left[ S + \sqrt{S^2 - 1}\right]} \geq 3.81 \]  

(1)

N = Filter order

LA = stopband attenuation = 60dB

LR = passband return loss = 10dB

S = Shape Factor = 5
Selection of PCB Material

Park Electromechanical Corporation provided samples of a PCB material (Nelco N4000-12) with desirable dimensions and electrical characteristics. This board material is 31 mils thick with ½ oz copper (about 0.7mils thick) laminated on the top and bottom of the PCB.

The PCB board material was selected to have a low loss tangent (i.e. high quality factor). A filter constructed with lower loss tangent material will have the advantage of lower insertion loss ($S_{21}$) in the passband. However, the higher quality PCB materials with these specifications tend to be more expensive. The Nelco N4000-12 material has a loss tangent of 0.008. An ideal loss tangent would be zero, meaning no loss of signal energy into the dielectric substrate.

The 31-mil board is thinner than the more common 62 mil but remains rigid for the small dimensions. The thinner board reduces stray inductance, it is hoped this will reduce losses due to radiated emission. Each stage of the filter will have a trace pair with a run length approximately a quarter wavelength or ~ 1 inch. The CAD design software indicated trace pair separation for this board thickness will range from more than 5mils to no more than ~100mils. This range insures the coupled pairs are physically close relative to their length and will act as differential pairs. Traces separated by no less than 5 mils will be within minimum etching limits for any PCB fabrication company.

Calculation of Stripline Dimensions

Calculation of the trace dimensions will be broken into steps for better exposition.
First, the length of each of the four stages of the third order coupled line filter will be determined. Each sequential stage of a coupled line filter contributes approximately a quarter wavelength at midband ($\lambda/4$ at $f_0$) to the physical structure of the filter.

A couple of potential trace lengths will be derived below for some representative trace widths. With:

- $E_r = \text{Dielectric constant supplied by the manufacturer as } 3.70$
- $E_{reff} = \text{Effective dielectric constant for a given PCB trace}$
- $w = \text{Trace width}$
- $h = \text{Dielectric substrate thickness } = 0.031''$
- $\lambda_0 = \text{Wavelength at midband}$
- $f_0 = \text{Midband frequency of filter } = 1.69\text{GHz}$

The effective dielectric constant for trace widths of 54 and 66 mils is computed as [1]:

$$E_{reff} (w = 54) = \frac{(E_r + 1)}{2} + \frac{(E_r - 1)}{2 \sqrt{1 + \frac{12h}{w}}} = \frac{(3.7 + 1)}{2} + \frac{(3.7 - 1)}{2 \sqrt{1 + \frac{12 \text{mil}}{54 \text{mil}}}} = 2.83$$

$$E_{reff} (w = 66) = \frac{(E_r + 1)}{2} + \frac{(E_r - 1)}{2 \sqrt{1 + \frac{12h}{w}}} = \frac{(3.7 + 1)}{2} + \frac{(3.7 - 1)}{2 \sqrt{1 + \frac{12 \text{mil}}{66 \text{mil}}}} = 2.87$$

$$\frac{\lambda_0}{4} \bigg|_{f_0 = 1.69\text{GHz}} = \frac{c}{\sqrt{E_{reff}}} \bigg|_{f_0 = 1.69\text{GHz}} = \frac{3.0 \times 10^8 \text{m/s}}{\sqrt{2.83}} = \frac{4(1.69 \times 10^9 \text{Hz})}{1039 \text{mils}} \quad (2)$$

$$\frac{\lambda_0}{4} \bigg|_{f_0 = 1.69\text{GHz}} = \frac{c}{\sqrt{E_{reff}}} \bigg|_{f_0 = 1.69\text{GHz}} = \frac{3.0 \times 10^8 \text{m/s}}{\sqrt{2.87}} = \frac{4(1.69 \times 10^9 \text{Hz})}{1031 \text{mils}}$$
A third order coupled line filter will have four, quarter wavelength segments:

\[
FilterLength = (N + 1)\left(\frac{\lambda_s}{4}\right) = (3 + 1)(1.03") = 4.12"
\]  (3)

The next step is to determine even and odd trace impedance for each filter stage. This impedance will be used in turn to determine the physical dimensions of each filter stage’s traces.

To make the calculation more manageable the formula relating tabulated filter coefficients, bandwidth, midband frequency and characteristic line impedance is expressed in terms of J-parameters

Calculation of J parameters will require the tabulated coefficients [1] for a third order Chebychev filter with passband ripple of less than 0.5dB. These coefficients represent normalized element values of an equivalent LC filter:

\[
g_0 = 1.0000 \\
g_1 = 1.5963 \\
g_2 = 1.0967 \\
g_3 = 1.5963 \\
g_4 = 1.0000
\]

Next we need the filter’s normalized bandwidth (BW):

\[
f_0 = \sqrt{\frac{f_u}{f_l}} = 1.69GHz \\
BW_{normalized} = \frac{Bandwidth}{f_0} = \frac{1.69GHz}{0.169GHz} = 0.1
\]  (4)

\[
f_u = 1.78GHz \\
f_l = 1.61GHz
\]

We now have the necessary parameters to calculate $J_{0,1}$-$J_{N,N+1}$ [1]:
Next calculate even and odd impedances for each stage $Z_{0e}, Z_{0o}$ [1].

\begin{align*}
Z_0 &= 50\Omega \\
Z_0 &\cdot J_{0,1} = \sqrt{\frac{\pi (BW)}{2g_0 g_1}} = 0.3137 \\
Z_0 &\cdot J_{1,2} = \frac{\pi (BW)}{2\sqrt{g_1 g_2}} = 0.1187 \\
Z_0 &\cdot J_{2,3} = \frac{\pi (BW)}{2\sqrt{g_2 g_3}} = 0.1187 \\
Z_0 &\cdot J_{3,4} = \sqrt{\frac{\pi (BW)}{2g_3 g_4}} = 0.3137
\end{align*}

(5)

The resulting even and odd impedances are tabulated in Table 3.

**TABLE 3**

<table>
<thead>
<tr>
<th>Stage: $i, i+1$</th>
<th>$Z_{0o}$ (Ω)</th>
<th>$Z_{0e}$ (Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0,1</td>
<td>39.2</td>
<td>70.6</td>
</tr>
<tr>
<td>1,2</td>
<td>44.8</td>
<td>56.6</td>
</tr>
<tr>
<td>2,3</td>
<td>44.8</td>
<td>56.6</td>
</tr>
<tr>
<td>3,4</td>
<td>39.2</td>
<td>70.6</td>
</tr>
</tbody>
</table>

The next step is to calculate the dimensions of the differential trace pairs.

Each differential pair consists of two traces of width $w$, separated by distance $s$ and lying height $h$ above the ground plane (equal to the thickness of the dielectric substrate).
The PCB material will be Park Electromechanical Corporation’s Nelco N4000-12®. This material has a low loss tangent (listed as ‘Dissipation Factor’, Df) of 0.008 and Dielectric Constant Er = 3.70, each specified by the manufacturer at 1.0GHz. Er is further specified to be 3.60 at 10GHz.

At this point published equations [3] are used to produce the differential trace dimensions for each stage of the distributed filter structure, as tabulated in Table 4. See Appendix A for a Matlab example of these equations. These equations derive the dimensions from the following parameters:

\[
Z_{0o} = \text{Odd mode differential impedance for a differential pair (from Table 3)}
\]

\[
Z_{0e} = \text{Even mode differential impedance for a differential pair (from Table 3)}
\]

\[
Er = \text{Dielectric constant of PCB material from manufacturer} = 3.70
\]

\[
w = \text{width of individual trace (TBD)}
\]

\[
s = \text{separation between differential trace pairs (TBD)}
\]

\[
h = \text{height of traces above the ground plane} = 31 \text{ mils}
\]

<table>
<thead>
<tr>
<th>Stage</th>
<th>W(calc) (mils)</th>
<th>W(design) (mils)</th>
<th>S(calc) (mils)</th>
<th>S(design) (mils)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>53.8</td>
<td>54</td>
<td>7.37</td>
<td>8</td>
</tr>
<tr>
<td>2</td>
<td>65.6</td>
<td>66</td>
<td>32.44</td>
<td>32</td>
</tr>
<tr>
<td>3</td>
<td>65.6</td>
<td>66</td>
<td>32.44</td>
<td>32</td>
</tr>
<tr>
<td>4</td>
<td>53.8</td>
<td>54</td>
<td>7.37</td>
<td>8</td>
</tr>
</tbody>
</table>

Appendix A for a Matlab example of these equations. These equations derive the dimensions from the following parameters:
Evaluation of Design Performance

The design produced by calculation was entered into Sonnet simulation software for evaluation as seen in Figure 1. A second version of this design, optimized using AWR’s CAD design software was similarly simulated (see Figure 2). The optimized CAD design was then prototyped and tested on the bench.

Fig. 1. Filter layout, filter design resulting from calculations, as displayed in Sonnet™.
Fig. 2. Response of filter designed by calculation, simulated by Sonnet™ software. Center frequency is 1.67GHz. Bandwidth is 0.180 GHz.
CHAPTER IV

COMPUTER AIDED DESIGN

As a final step the coupled line filter is redundantly designed using AWR filter design wizard. This software accepts filter parameters and produces physical dimensions of the filter layout and a simulation of the filter response (Figure 3). Parameters entered include filter type (e.g., Chebychev), filter order, PCB specifications including Dielectric constant (Er) and loss tangent. PCB physical specifications must also be provided, including substrate and copper laminate thickness and the presence of plating and solder mask.

The software generates physical layout dimensions which can then be translated to a PCB layout. The resulting layout specifications as seen in Figure 4 showing trace width (W), length(L) and inter-trace separation (S) for the coupled line structure. These parameters are entered into a layout package such as ORCAD layout (Figure 5) which can generate Gerber files required by PCB fabrication facilities. The resulting filter design was nearly identical to that produced by published formula. With confidence established in the design layout, the prototype can be fabricated and tested on the bench.
Fig. 3. Frequency response of the Microwave Office filter design. The response was generated by Sonnet ™ software.
Fig. 4. AWR’s Microwave Office filter design ‘wizard’ feature generated the coupled line filter design. Filter order, desired frequency response and PCB specifications were among the entries. PCB specifications are for Nelco® N4000-12 High Performance two layer Material with 0.5oz un-plated copper both sides.
Fig. 5. Board layout of Microwave Office filter design as it appears in Sonnet™ software package.
CHAPTER V

RESULTS

A prototype of the CAD deigned filter was successfully fabricated and tested. Passband insertion loss ranged from 2.9 to 3.6dB. The contribution of the dielectric quality can be estimated to be about 1.0 to 1.5dB (see Eqn for ‘LA’ in Literature Review) [5]. The remaining one or two dB loss is assumed to result from radiated emission loss, losses occurring in the conductors and losses due to interface adapters. Reduction in conducting area due to skin effects at these frequencies will contribute to losses in the copper conductors. Midband frequency was 1.61 GHz, having a 0.08GHz offset from the target of 1.69GHz. Bandwidth was 0.206GHz, which is 23% wider than the goal of 0.169GHz. Stop-bandwidth was 1.06GHz closely matching expectations. The filter response curve is shown in Figures 6 and 7 and some data points from these figures are tabulated in Tables 6 and 7. Table 5 compares the simulated responses for the various approaches and measurement.

The offset in the filter midband might be explained by variations in the electrical length of each stage’s traces due to variations in substrate Er. The electrical length was calculated from the datasheet’s nominal value of Er and a tolerance for this parameter was not listed. Each stage’s electrical length is inversely proportional to the root of Er as seen in the previous calculation of $\lambda_0/4$. An approximate 10% deviation from
the specified Er of 3.70 would be enough to account for the 5% midband deviation. Refinements to the calculation of effective dielectric, as well as electrical length of each stage are implied in various papers [3], [4], [6], an approach which requires further investigation.

Measured bandwidth was wider than simulated bandwidth. Desired bandwidth figure heavily in the trace dimensions so the error is likely do to production errors. The filter and measurement interface were verified to be symmetric by comparing frequency responses S11 and S22 (return loss at each port) and by comparing S12 and S21 (insertion loss at each port). This establishes a measure of confidence in production quality and in the consistent quality of the measurement interface. The source of the bandwidth error requires further investigation.

Measurement Procedure

The Agilent 8714ES RF Network analyzer was calibrated in ‘User two-port’ configuration using ‘SHORT’, ‘LOAD’ (50ohm) and ‘OPEN’ calibration fittings. Each N port was fit with an N to BNC adapter. A twelve-inch BNC cable connected the BNC adapters to the board. Each board was fit with edge-mount SMA connectors which in turn were mounted with SMA to BNC adapters.

The setup can be illustrated as follows:

Analyzer(Port1)<>N to BNC<>BNC Cable<>BNC to SMA<>SMA edge con<>filter
Analyzer(Port2)<>N to BNC<>BNC Cable<>BNC to SMA<>SMA edge con<>filter

The measured data is summarized in Table 5. Figure 6 is a screen copy of the measured insertion loss and Table 6 is selected measurements from this response. Figure
TABLE 5
COMPARISON OF FILTER RESPONSES

<table>
<thead>
<tr>
<th>Goal</th>
<th>Design by calculation (simulated)</th>
<th>CAD generated design (simulated)</th>
<th>Measured Result</th>
</tr>
</thead>
<tbody>
<tr>
<td>Passband Insertion Loss (S21)</td>
<td>&lt;5dB</td>
<td>&lt;2.5dB</td>
<td>&lt;3.6dB</td>
</tr>
<tr>
<td>Midband (f0)</td>
<td>1.69GHz</td>
<td>1.72GHz</td>
<td>1.61GHz</td>
</tr>
<tr>
<td>$BW_{3dB}$</td>
<td>169KHz</td>
<td>130KHz</td>
<td>206KHz</td>
</tr>
<tr>
<td>Passband Return Loss (S11)</td>
<td>&gt;15dB</td>
<td>&gt;12dB</td>
<td>&gt;9dB</td>
</tr>
</tbody>
</table>

7 is a screen copy of the measured return loss and Table 7 is selected return loss measurements from this return response.

The passband insertion loss ranges from -2.9dB to -3.6dB. This passband ripple of 0.7dB is slightly over the expected 0.5dB ripple for this filter. Bandwidth and midband are derived from the cutoff points $f_i$ and $f_u$, taken at -6dB.

$$f_{i\, meas} = 1.508GHz$$

$$f_{u\, meas} = 1.714GHz$$

$$f_0 = \sqrt{f_u f_i} = 1.608GHz$$

$$BW_{3dB} = f_u - f_i = 206KHz$$

(7)

The measured shape factor for the filter prototype is 5.2. This is consistent with the earlier estimated value for an N=3 filter.

$$ShapeFactor = \left(\frac{f_{midband} - f_{60db}}{Bandwidth / 2}\right) = \left(\frac{1610KHz - 1079KHz}{(206KHz / 2)}\right) = 5.2$$

(8)

A sample of calculations based on filter attenuation measurements from Table 4 is shown in Table 6. These results show the 60dB attenuation at $\Omega = -6.42$ corresponds closely with expectation for an N=3 filter. However, the passband region remains wider.
Fig. 6. Measured frequency response (Insertion Loss S21) of prototype. Measured on Agilent 8714ES RF Network Analyzer.

than expected until about -40dB of attenuation. This is also reflected in the wider than expected bandwidth.
### TABLE 6
**MEASURED FREQUENCY RESPONSE (S21) OF PROTOTYPE. MEASURED ON AGILENT 8714ES RF NETWORK ANALYZER**

<table>
<thead>
<tr>
<th>Measurement</th>
<th>f(MHz)</th>
<th>S21(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Marker 1</td>
<td>1438</td>
<td>-20</td>
</tr>
<tr>
<td>Marker 2</td>
<td>1508</td>
<td>-6</td>
</tr>
<tr>
<td>Marker 3</td>
<td>1714</td>
<td>-6</td>
</tr>
<tr>
<td>Marker 4</td>
<td>1818</td>
<td>-20</td>
</tr>
<tr>
<td>A</td>
<td>1301</td>
<td>-40</td>
</tr>
<tr>
<td>B</td>
<td>1079</td>
<td>-60</td>
</tr>
<tr>
<td>C</td>
<td>1533</td>
<td>-3.6</td>
</tr>
<tr>
<td>D</td>
<td>1588</td>
<td>-2.9</td>
</tr>
<tr>
<td>E</td>
<td>1633</td>
<td>-3.6</td>
</tr>
<tr>
<td>F</td>
<td>1668</td>
<td>-3.25</td>
</tr>
<tr>
<td>G</td>
<td>1689</td>
<td>-3.6</td>
</tr>
</tbody>
</table>
Fig. 7. Measured frequency response (Return Loss S11) of prototype. Measured on Agilent 8714ES RF Network Analyzer.

### TABLE 7
RETURN LOSS (S11), MEASURED FREQUENCY RESPONSE OF PROTOTYPE.
MEASURED ON AGILENT 8714ES RF NETWORK ANALYZER

<table>
<thead>
<tr>
<th>Measurement</th>
<th>f(MHz)</th>
<th>S11(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Marker 1</td>
<td>1502</td>
<td>-5</td>
</tr>
<tr>
<td>Marker 2</td>
<td>1520</td>
<td>-10</td>
</tr>
<tr>
<td>Marker 3</td>
<td>1690</td>
<td>-10</td>
</tr>
<tr>
<td>Marker 4</td>
<td>1718</td>
<td>-5</td>
</tr>
</tbody>
</table>
CHAPTER VI

CONCLUSION

A coupled line bandpass filter was successfully designed by direct calculation and by using a CAD design tool. Both designs were nearly identical in physical dimensions and in their simulated frequency responses. These frequency responses were well within specified ranges.

A prototype of the CAD designed filter performed well on the bench. Differences between simulated and prototype performance likely resulted from material specification tolerance and other, possibly production or measurement related factors. These differences included an offset in the filter midband, wider than expected passband and higher than expected midband loss. The most significant error was the 80MHz offset in placement of the midband. Tuning capability would have been a helpful calibration capability to resolve this error. However, the addition of tuning circuits degrades circuit quality, resulting in higher midband loss.

Midband loss measured less than the 5dB goal, however further improvements would be needed for some applications. Losses in the dielectric substrate likely account for less than 1.5dB of the total loss. Measurement and production errors may account for another 1 dB. This was implied by the difference between measured and simulated loss. Other design and manufacturing refinements would be needed to exceed 1dB demanded by some applications.
PCB material was sampled by Roger’s Corporation. The quality factor of this material was relatively high compared to a standard FR4 substrate and much higher quality materials are available. A higher quality material would have reduced some of the substrate loss. Also a higher dielectric constant would result in a smaller and thinner board, likely further reducing the contribution of emissions and conductor losses. However the contribution of measurement and production errors remains uncertain.

The design equations could be further optimized by taking into account field fringing effects and by obtaining precise material specs from the PCB manufacturer. Both of these efforts would produce refinements to trace length dimension which are critical to establishing placement of filter midband. Field fringing effects produce an error factor in the electrical length of the traces. Small errors in Er will change signal wavelength within the substrate, thereby further altering trace electrical length.

Taken together refinements in board material quality and in the design layout, along with some tuning capability may resolve most of the observed errors. The result should be an excellent filter design, achievable without the need for expensive CAD tools.
REFERENCES
REFERENCES


APPENDIX A
MATLAB CODE EXAMPLE

Matlab code example, determines trace dimensions which achieve the target even and odd impedance.

The remaining equations are accurate for $0.1 \leq u \leq 10$, $0.1 \leq g \leq 10$, where $u = w/h$ (width of trace over thickness of substrate) and $g = s/h$ (separation between traces over thickness of substrate) [1].

Executed at the Matlab command line:

```
Zcet = 50.8; Target even mode impedance for the differential pair
Zcot = 49.2; Target odd mode impedance for the differential pair
x = fminsearch(@(x) optimfun(x, Zcet, Zcot), [30, 20, 25])
```

Associated M-file code:

```
function f = optimfun(x, Zcet, Zcot)

%w = 2.46
%h = 1.905
%s = 0.7

w = x(1);
h = 20;
s = x(3);

Er = 3.70;

u=w/h;
g=s/h;

%Eeff = (Er + 1)/2 + (Er - 1)*((1 + 12*h/w)^-0.5 + 0.04*(1-w/h)^2)/2 %w/h<1
%Zc = (376.8 * log(8*h/w+w/(4*h)))/(6.283 * Eeff^0.5) %w/h<1

Eeff = (Er + 1)/2 + (Er - 1)*((1 + 12*h/w)^-0.5)/2; %w/h>1
Zc = 376.8 / ((Eeff^0.5) * (1.393 + w/h + .667 * log(w/h + 1.444))); %w/h>1
```
\[ v = \frac{(g^2 + 20)u}{g^2 + 10} + g\exp(-g); \]
\[ Ae = 1 + \frac{(1/49)\log((v^4 + (v/52)^2)/(v^4+0.432))}{(1/18.7) * \log(1+(v/18.1)^3));} \]
\[ Be = 0.564 * (\frac{(Er-0.9)/(Er+3))^{0.053};} \]
\[ Ereeff = \frac{(Er + 1)/2 + (Er - 1)/2 * (1 + 10/v)^{-Ae*Be)}; \]

\[ Q1 = 0.8685*u^{0.194}; \]
\[ Q2 = 1 + 0.7519*\frac{g}{g + 0.189*2.31}; \]
\[ Q3 = 0.1975 + (16.6 + (8.4/g)^6)^{-0.387 + \log(g^10/{(1+(g/3.4)^10)})/241}; \]
\[ Q4 = 2*(Q1/Q2)/(u^{Q3 * \exp(-g)} + (2-exp(-g))*u^{(-Q3)}); \]
\[ Zce = Zc * (Eeff/Ereeff)^{0.5}/(1-Q4*(Eeff^{0.5})Zc/377) \]

\[ a0 = 0.7287*(Eeff-0.5*(Er+1))*(1-\exp(-0.179*u)); \]
\[ b0 = (0.747*Er)/(0.15+Er); \]
\[ c0 = b0 - (b0 - 0.207) * \exp(-0.414*u); \]
\[ d0 = 0.593 + 0.694*\exp(-0.526*u); \]
\[ Eroeff = Eeff + (0.5*(Er+1)-Eeff + a0)*\exp(-c0*g^{d0}); \]

\[ Q5 = 1.794 + 1.14*\log(1+0.638/(g+0.517*2.43)); \]
\[ Q6 = 0.2305 + \log(g^{10}/(1+(g/5.8)^10))/281.3 + \log(1+0.598*g^{1.154})/5.1; \]
\[ Q7 = (10+190*g^2)/(1+82.3*2^3); \]
\[ Q8 = \exp(-6.5-0.95*\log(g)-(g/0.15)^5); \]
\[ Q9 = \log(Q7)*(Q8+1/16.5); \]
\[ Q10 = Q4*(Q5/Q2)*\exp(Q6*\log(u)/u^{Q9}); \]
\[ Zco = Zc * (Eeff/Eroeff)^{0.5}/(1-Q10*Eeff^{0.5}Zc/377) \]
\[ f = \text{abs(Zcet-Zce)}+\text{abs(Zcot-Zco)}; \]

REFERENCES

FABRICATION INSTRUCTIONS TO
SUNSTONE CIRCUITS

Mike Flanner
California State University Chico
Part Number: filter70_v1

Board dimension: 2" X 4.5"
Layers: 2
Material: Nelco N4000-12
Board Thickness: .031"
Copper Weight 0.5oz top and bottom.

Layers:
Top gerber file: filter70.cmp
Outline gerber file: filter70.plc

Bottom side is 0.5oz copper, no etching.
No plating
No drill holes
No solder mask or silk screen

Figure B1. Top side of board. Minimum spacing is 6 mils.