Very Compact, High Quality LTCC Filter Bank Design

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ABSTRACT

Ever-increasing interest in RF and microwave components that meet exacting miniaturization and performance needs for commercial and military applications has pushed the development of cutting-edge design and manufacturing techniques. In this project, we will investigate the design of a compact 16-channel switched filter bank operating in the 2-7 GHz range (S-band to X-band). The challenging performance, size, and packaging requirements can be met by realizing the design in low-temperature co-fired ceramic (LTCC) substrate technology. A new double-layered resonator structure allows the resonator length to be reduced to about $\lambda/8$. Initial design of the individual filters will be based on the circuit model of a combline configuration. After that, a commercial finite element method solver (HFSS) will be used to lay out the 3-D electromagnetic model of the filters, and optimize their electrical performance.

Keywords: bandpass filter, combline, compact, low-temperature co-fired ceramic (LTCC), resonator, stripline, switched filter bank (SFB)

1. BACKGROUND

Numerous system applications require the frequency band of operation to be divided into smaller bands, which can be processed separately and then recombined. The typical configuration of a 16-channel system is shown in Fig. 1. The input signal can occupy the passband of one of the filters, which is selected by the appropriate switch at the input and output.

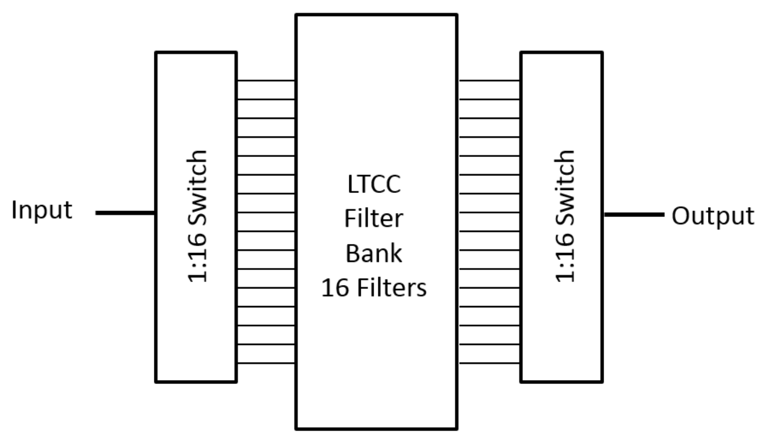

Figure 1. 16-channel Switched Filter Bank

The physical size of the proposed 16-channel SFB is 0.5″ x 1.0″ x 0.125″. The center frequencies of the filters range from about 2 GHz to 7.0 GHz. The relative bandwidths of the 16 filters are about 15% each. Passband insertion loss for each filter is required to be less than 2.0 dB, and the out-of-band rejection at the adjacent filters' center frequency is 20 dB minimum. The electrical and size requirements for this switched filter bank are extremely challenging. These specifications cannot be met by any conventional technology. In order to realize these filters, we have investigated the possibility of using a planar double-layer coupled stripline resonator structure, similar to Ref. [1].[1] A typical filter using this type of resonator is shown in Fig. 2.

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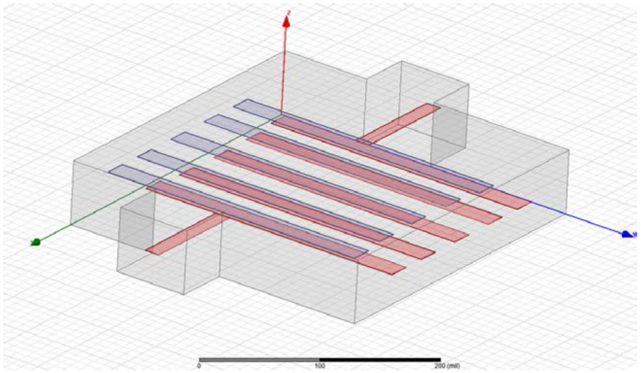


Figure 2. Combline filter using coupled stripline resonators

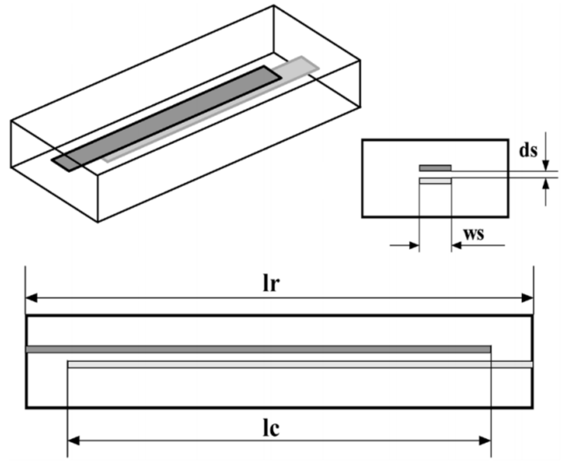


Figure 3. Double-layer coupled stripline resonator structure

2. RESONATOR STRUCTURE

The forerunner of the double-layer stripline resonator structure has its own roots in the quarter-wave transmission line, an archetypal compact resonating 1-D structure that is open-circuited at one end and short-circuited at the other. It is well known that resonant frequencies occur at frequencies where the effective length of such a structure is an odd multiple of $\lambda/4$. If compactness is an important design constraint, it is possible to "fold" the guided-wave structure on the same plane, making it a 2-D structure with about half the original dimensions. At this point, strong couplings between the two "legs" need to be realized precisely if one wanted to obtain comparable performance–no easy matter in itself. Another approach would simply be to have discrete line elements side-by-side (instead of connected to each other) which are capacitively loaded. Such "coupled-line resonators" have proven to be a highly promising line of development for many filter designers.

In this vein, a novel 3-D double-layer structure was conceived and reduced to practice, where strong capacitive loading effects between two stripline structures, one underneath the other, could be used to reduce to reduce physical lengths.[2] A single double-layer coupled stripline resonator is shown in figure 3. The opposite ends of the strips are shorted to ground, introducing a capacitive coupling effect that lowers the resonant frequency. Thus, in order to increase the resonant frequency, the physical length lr has to be reduced even further than half of the $\lambda/4$ length. In fact, it is possible to realize resonators that are as little as $\lambda/12$ in length. The overall length lr of the resonator (and thereby, the resonant frequency), is determined by the width ws of the two strips, the vertical distance ds between the two strips, and the coupled overlap length lc of the two strips (figure 3).

The width $ws = 20$ mils was previously[1] found to give optimum results for L-band filters (1-2 GHz) of the resonator lengths that meet the size specifications we are interested in. A manufacturing constraint will fix the vertical distance ds between the strips to an integer multiple of one dielectric (ceramic) layer thickness. The overall height of the resonator is similarly constrained. The available layer thicknesses for LTCC ceramic are 2.95 mil, 3.94 mil, 7.87 mil, and 12 mil. The metalization thickness is 0.4 mil. A suitable stacking configuration has been found by practice to be $(12+12+3.94+2.95+3.94+12+12)$ mils $= 58.83$ mils. The dielectric constant of LTCC is $\epsilon_r = 7.8$, and it has a loss tangent of $\tan\delta = 0.001$. For a fixed lr, the overlap lc will need to be empirically determined for the target center frequency. Since the overall resonator length lr can range between $\lambda/12$ to $\lambda/8$, it might be desirable to establish a single lr for a sub-bank of multiple resonators next to each other in frequency, and sweep the overlap lc to achieve the correct resonant frequency.

3. FILTER CONFIGURATION

For the very compact 16-channel switched filter bank, we want to have relatively narrow bandpass filters. To this end, the double-layer combline configuration (figure 2) is a proper choice for realization of the filter. The maximum achievable coupling value with the minimum realizable space in this configuration is generally larger than the required inter-coupling k between resonators for a relatively narrow bandpass filter with the electrical specifications we desire (relative bandwidth $\omega_r \approx 0.15$, minimum attenuation $L'_T \geq 20$ dB at rejection frequency ω'_T, passband insertion loss $L_{Tp} \leq 2$ dB); so it meets our needs for compactness while not compromising our ability to realize the required couplings.

After determining the electrical specifications, one may also consider the environmental specifications (temperature, radiation hardening, etc), that may give rise to a set of modified specifications. With the specifications now determined, we next need to solve a polynomial approximation problem that gives rise to the filter's transfer function. In our case, to achieve a steep rolloff in the transition band and allowable ripple in the passband, the Chebyshev characteristic was chosen as our filter's transfer function. The transfer function of this filter has no zeros of transmission, and poles positioned to give a a desirable return loss across the passband centered around f_o. As for our coupling structure, the coupled-line resonator lends itself to a cascaded inline topology for all-pole filters (such as Chebyshev).

4. FILTER SYNTHESIS

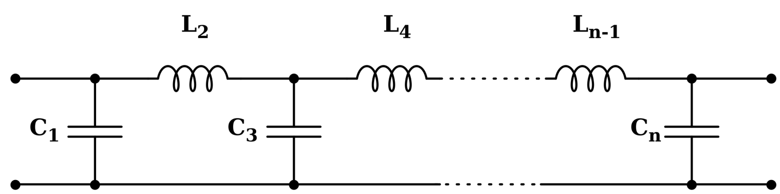


Figure 4. Low pass filter, Cauer topology

In order to design a filter to the given specification, we first need to determine the order n of the filter (figure 4), which depends on the cutoff characteristics of the filter. The specifications for a prototype Low Pass Filter are cutoff frequency ω_C, the required minimum attenuation L'_T (dB) at frequency ω'_T. If the filter is a Chebyshev, we need the maximum ripple L_{ar} (dB) in the passband to compute maximum ripple $\kappa = 10^{L_{ar}/10}$.

With this information, we determine $n = \cosh^{-1}\sqrt{(10^{L'_T/10}-1)/(\kappa-1)}/\cosh^{-1}(\omega'_T/\omega_c)$ for Chebyshev filters, where n is an integer. For our cutoff and attenuation requirements, a filter of order $n = 5$ was found to suffice. Next, we need to calculate the element values g_k; $(k = 0 \ldots n)$; of the LPF prototype. First, we calculate coefficients $\beta = \ln(\coth(L_{ar}/17.37))$, $\gamma = \sinh(\beta/2n)$. Then $a_k = \sin((2k-1)\pi/2n)$ and $b_k = \gamma^2 + \sin^2(k\pi/n)$ for $k = 1 \ldots n$. We may then find the element values $g_k = \frac{4a_{k-1}a_k}{b_{k-1}g_{k-1}}$ for $k = 2 \ldots n$, $g_0 = 1$, $g_1 = 2a_1/\gamma$, $g_{n+1} = 1$ for n even, $g_{n+1} = \coth^2(\beta/4)$ for n odd.

If this was simply a lowpass filter, we directly find the actual lumped element values C_1, L_2, ... C_n as in figure 4 by scaling the frequency and impedance of its LPF prototype: $C_k = g_k/\omega_C Z_0$ and $C_k = g_k Z_0/\omega_C$. In order to construct a band pass filter, however, we perform a frequency mapping of an LPF prototype for the band edge frequencies at which attenuation equal ripple (for Chebyshev) for normalized frequency ω'_T/ω_C.

The g_k values found in the calculation of the low pass filter can now be used to find and replace the series and shunt elements of the LPF with a pair of LC elements (a resonator) to synthesize the band pass filter (figure 5). For series elements, $C_k = \frac{\omega_r}{\omega_o g_k}\left(\frac{1}{Z_o}\right)$, $L_k = \frac{g_k}{\omega_o \omega_r}(Z_o)$, and for shunt elements, $C_k = \frac{g_k}{\omega_o \omega_r}\left(\frac{1}{Z_o}\right)$ and $L_k = \frac{\omega_r}{\omega_o g_k}(Z_o)$.

In order to synthesize the BPF as an coupled resonator filter, it is convenient to convert all the resonators to the same form (either series or shunts) by using impedance inverters. To do that, we need to find the couplings

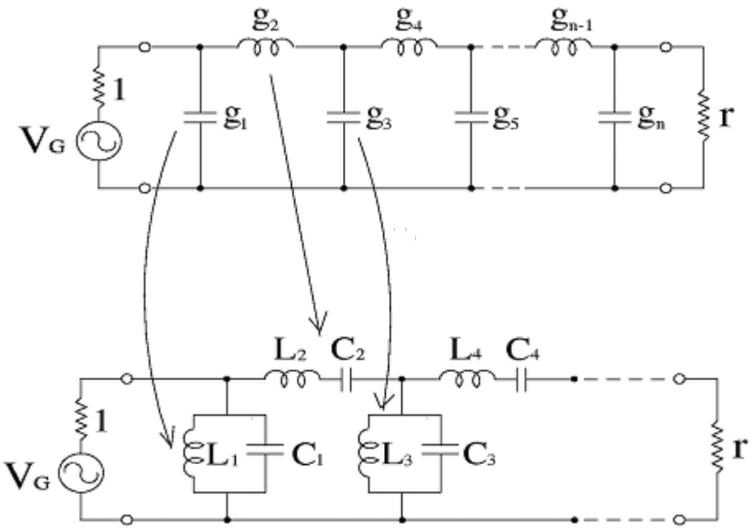


Figure 5. LPF to BPF element mapping

K between the elements. For the $k = 1, 2, ..., n-1$ elements in between the first and last resonators, we have normalized couplings $\frac{K_{k,k+1}}{Z_o} = \frac{1}{\sqrt{g_k g_{k+1}}}$, while $\frac{K_{0,1}}{Z_o} = \frac{K_{n,n+1}}{Z_o} = \frac{1}{\sqrt{g_1}}$. Since these calculations are so tedious, a spreadsheet is helpful calculate the coefficients, couplings, and element values. The chief independent variables were filter order $n = 5$, passband ripple $L_{ar} = 0.05$ dB, center frequency f_0 (see table 1), and edge frequencies $f_{1,2}$; which were actually computed from the bandwidth $f_r = \frac{\Delta f}{f_0} \Rightarrow f_{1,2} = \frac{f_0}{2}\left(\sqrt{f_r^2 + 4} \pm f_r\right)$. From these inputs, we were easily able to obtain the normalized couplings $m_{1,2}$, $m_{2,3}$, $m_{3,4}$, and $m_{4,5}$ between the normalized resonators and the normalized external couplings $\bar{R}_{in} = \bar{R}_{out} = 1/\sqrt{g_1}$.

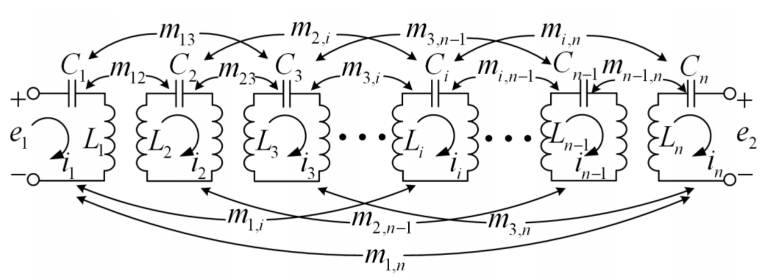


Figure 6. Generalized 2-port circuit model of all-coupled resonator filter

The coupling matrix is a filter synthesis tool used to generate an arbitrary circuit model of n lumped element series resonators with inter-couplings between all of the other resonators in the filter; adjacent as well as non-adjacent. This tool was introduced by Atia et al.[3] in the early 1970s at COMSAT laboratories in Clarksburg, MD. Instead of going through the steps of building an LPF prototype one by one, the fiter synthesis of a bandpass filter is done directly, through the machinery of linear algebra. The method recognizes that the fundamental building block of a filter is the resonator. And it is the couplings between resonators, which could be expressed in the form of a coupling matrix, that determine the some of most important behaviors of the filter. One builds a coupling matrix to be as general as possible; to not put a limit on the possible couplings between the resonators. Matrix operations are applied to bring the coupling matrix into a desired form.

Although it is a powerful tool, it is intended to be used for narrowband filters (relative bandwidth 15% or less). One makes the approximation that couplings are frequency independent in order to facilitate computation. By treating the bandwidth as narrow, one can assume that over the bandwidth of interest, frequency ω is constant, centered around ω_0 without very much variation. It is possible, but unnecessary, to introduce frequency independence, and it turns out that using the approximation gives quite accurate results.

The procedure is as follows: Kirchhoffs law is applied to the loop currents in the series resonators in the circuit, which leads to a set of equations that can be represented by a matrix $[V] = [Z][I]$, where voltage matrix $[V] = v_s[1, 0, 0, ..., 0]^T$, current matrix $[I] = [i_1, i_2, i_3, ..., i_n]$, and impedance matrix is $[Z] = [jM + \lambda\mathbf{1} + R]$. Here, v_s is the voltage source, $i_1, i_2, i_3, ..., i_n$ are the currents in each of the resonator loops. R is an $N \times N$ all-zero matrix except the first and last elements, which are equal to source impedance R_s and load impedance R_L, respectively. $\mathbf{1}$ is the identity matrix. Coupling matrix M contains all possible mutual couplings m between the resonators.

$$M = \begin{bmatrix} m_{11} & m_{12} & m_{13} & \dots & m_{1n} \\ m_{21} & m_{22} & m_{23} & \dots & m_{2n} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ m_{n1} & m_{n2} & m_{n3} & \dots & m_{nn} \end{bmatrix}$$

From here, the admittance matrix $[Y]$ is obtained from a partial fraction expansion of Y-elements, and the entries of the M matrix are solved for analytically. This will give an M matrix with all nonzero elements. All couplings except the ones on the diagonal (self-couplings) are called cross-couplings. The next step is to apply a sequence of similarity transformations to the M matrix in order to annihilate couplings that cannot be realized in practical form, until the matrix coincides with a topology matrix T, made up of 1's and 0's, which corresponds to the filter topology that we are trying to realize. For instance, if you want a a matrix with one non-zero, non-diagonal element $\hat{M}_{ik}$ (where $k \neq j$) reduced to zero, define Q; an identity matrix with a pivot, with ones on the diagonal, and $\begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix}$ on the zero replacements, choosing θ so that $\tan\theta = -M_{ki}/M_{ji}$, where $M_{ji} \neq 0$. Then $M_1 = QM_0Q^T$. After the successive annihilations are performed, and you have a coupling matrix with a convenient topology, you can denormalize the couplings and apply them to the filter structure you have at hand.

For our cascaded 5-pole Chebyshev filter, the coupling matrix would look like this:

$$M = \begin{bmatrix} 0 & m_{12} & 0 & 0 & 0 \\ m_{12} & 0 & m_{23} & 0 & 0 \\ 0 & m_{23} & 0 & m_{34} & 0 \\ 0 & 0 & m_{34} & 0 & m_{45} \\ 0 & 0 & 0 & m_{45} & 0 \end{bmatrix}$$

The couplings, along with the frequency and bandwidth, gives us the all the parameters we need to build an ideal circuit model in lumped-element form. The frequency response of such a circuit can be easily modeled in a circuit simulation package such as ADS or Microwave Office.

5. FILTER MODELING AND OPTIMIZATION

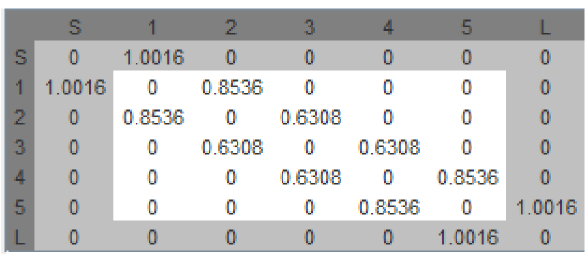
After the circuit model prototype and response has been found, the next step is to physically realize the filter with the ideal model as a guide. First, we determine the geometry of the double-layer coupled stripline resonator for a given center frequency. Material losses will be ignored, and we will assume we are using a perfect conductor. Second, we find the separations between the resonators that give the required inter-couplings k. Third, we find the position to tap-in a stripline that gives us the correct external coupling R. Fourth, the entire filter is assembled according to these found parameters. The filter response is calculated by a full field electromagnetic solver such as HFSS. Because of the loading effects of the couplings, the initial response may need further tuning to get a good starting point for optimization. Fifth, optimization goals are set that will, with any luck, converge on a good solution. HFSS simulation can take several hours, might never converge on a good solution, or even reach a minimum acceptable cost function that is still unacceptable for various reasons. The parameter extraction technique[4] could be used as a way to guide fine-tuning in HFSS, but it is entirely possible to rely solely on HFSS. Material losses are then added, and finally the filter is re-optimized.

6. FILTER DESIGN EXAMPLE

Table 1. Switched Filter Bank filter specifications

Filter index	f_0 (GHz)	Δf (MHz)	Filter index	f_0 (GHz)	Δf (MHz)
0	1.5	188	8	3.279	410
1	1.654	207	9	3.616	452
2	1.824	228	10	3.987	498
3	2.011	251	11	4.396	550
4	2.218	277	12	4.848	606
5	2.445	307	13	5.346	668
6	2.697	337	14	5.895	737
7	2.974	372	15	6.5	812

The design of a filter begins selecting a filter index from the switched filter bank (table 1). Filter index 0 is chosen. This filter is an $n = 5$ pole Chebyshev filter with center frequency $f_0 = 1.5$ GHz, bandwidth $\Delta f = 188$ MHz, passband ripple $L_{ar} = 0.05$ dB, and passband return loss $L_R = 20$ dB. These specifications give us the coupling matrix and ideal filter response shown in figures 7 and 8.



	S	1	2	3	4	5	L
S	0	1.0016	0	0	0	0	0
1	1.0016	0	0.8536	0	0	0	0
2	0	0.8536	0	0.6308	0	0	0
3	0	0	0.6308	0	0.6308	0	0
4	0	0	0	0.6308	0	0.8536	0
5	0	0	0	0	0.8536	0	1.0016
L	0	0	0	0	0	1.0016	0

Figure 7. Coupling Matrix for filter $f0$

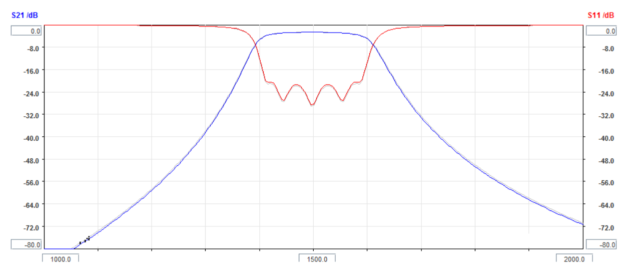


Figure 8. Insertion loss S_{21} and Return loss S_{11}

We have normalized external couplings and inter-couplings $\bar{R}_{in} = \bar{R}_{out} = 1.001561$, $m_{12} = 0.85361$, $m_{23} = 0.6308$, $m_{34} = 0.6308$, $m_{45} = 0.85361$. With this information, we proceed in building parameterized models in HFSS. We begin by finding the resonator dimensions that will have a fundamental resonating mode at 1.5 GHz.

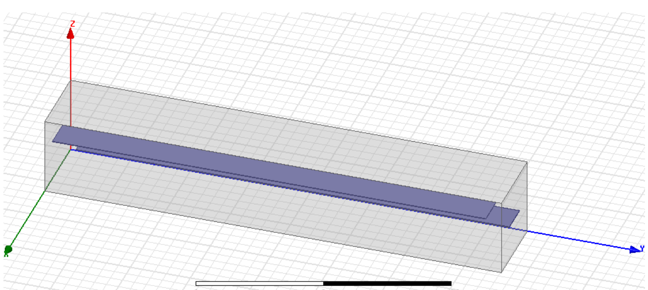


Figure 9. Resonator with length $lr = 300$ mils

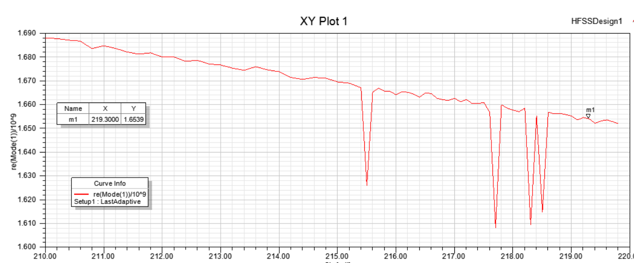


Figure 10. Eigenmode parameter sweep $lc = 210$ to 220 mils

The nominal dimensions for this structure are height $b = (12+12+3.94+2.95+3.94+12+12)$ mils $= 58.83$ mils, resonator length $lr = 300$ mils, stripline separation $ds = 2.95$ mils, stripline width $ws = 20$ mils, and metallization thickness $t = 0.4$ mils. The overlap lc is stepped in 0.1 mil increments in a parameter sweep. Eigenmode analysis is used to quickly find the first 2-4 resonating modes for each step change of lc. By interpolation, the value of lc corresponding to a resonant frequency $f_0 = 1.500$ GHz was found to be 269.1 mils. This value of lc is used as an initial starting point for subsequent analyses.

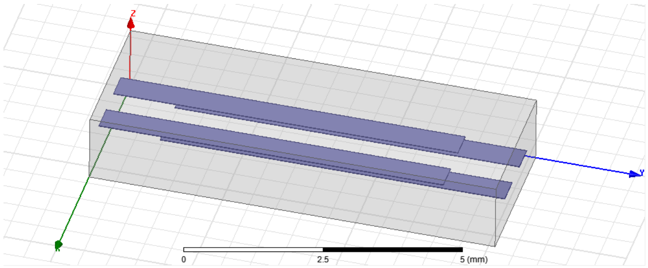


Figure 11. Finding resonator separation *sep*

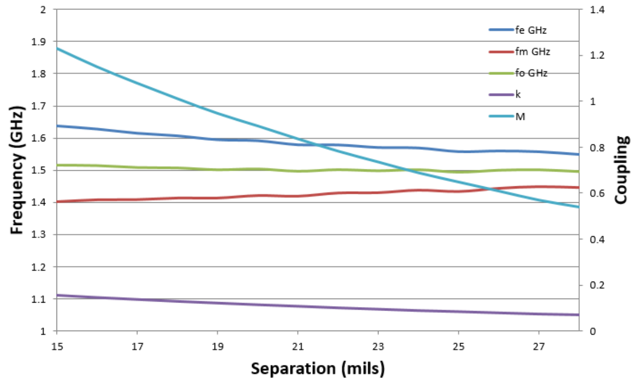


Figure 12. Separation versus Coupling

The next set of dimensions to find are separations that correspond to inter-couplings. For a five pole Chebyshev filter, only two couplings need to be found since $m_{12} = m_{45}$ and $m_{23} = m_{34}$. In order to find how the coupling varies with spacing of two identical resonators, one must perform eigenmode analysis on the two-resonator assembly to find f_m and f_e, the two lowest eigenmodes. Normally, to find f_m, one would replace the symmetry plane between the resonators with a perfect magnetic conductor (PMC) and obtain the lowest eigenmode. To find f_e, one would replace the symmetry plane with a perfect electric conductor (PEC) and obtain the lowest eigenmode. For a given separation *sep*, the resonant frequency of the two-resonator structure is $f_0 = \sqrt{f_e f_m}$. One can also determine the coupling coefficient $k = \frac{f_e^2 - f_m^2}{f_e^2 + f_m^2}$. And the normalized coupling is $m = k\frac{f_0}{\Delta f}$. As separation *sep* increases, f_0 will converge to a single frequency. Because of the capacitive loading between the elements, the resonant frequency of both resonators combined is lower than that of a single resonator. In order to bring the f_0 back up to 1.5 GHz, it is necessary to slightly decrease the overlaps lc by a few mils. The new lc, then, would be 252.7 mils. For this resonator length, the separation that corresponded to $m_{12} = m_{45} = 0.85361$ was 20.61 mils, and that of $m_{23} = m_{34} = 0.6308$ corresponded to 25.45 mils.

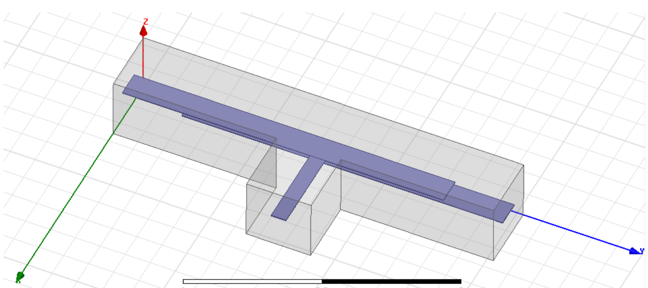


Figure 13. Finding resonator separation *sep*

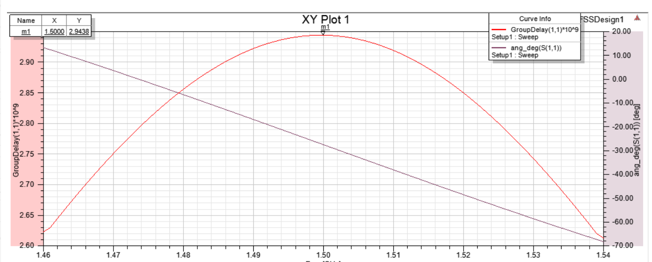


Figure 14. Finding *htapin*, group delay method

To find the correct tapped-in position that corresponded to input and output couplings $\bar{R}_{in} = \bar{R}_{out} = 1.001561$, we added a tapped-in line to position onto the single resonator model. The tapped-in line is usually $Z_0 = 50\Omega$ for impedance-matching purposes (preferably with no taper). The width of the line was computed using the ADS tool *Linecalc*. The stripline geometry "SLIN" was used. For a given ground plane spacing $B = 58.83$ mils, dielectric constant $\epsilon_r = 7.8$ and metal thickness $t = 0.4$ mils, the width of the line was found to be 14 mils.

The cross-sectional face of the tapped-in port is defined in HFSS as an excitation port. The driven modal solution is used, and a frequency sweep is added to measure S_{11}. Instead of measuring the magnitude, the phase $\phi(f_0)$ and the group delay $\tau_g = -\frac{d\phi(f_0)}{d\omega}$ are plotted. The objective is to find out where the group delay is at a minimum. At this value, the actual external coupling is $R = -\frac{4}{f_1 \left.\frac{d\phi}{df}\right|_{min}}$. Since $\frac{d\phi}{df} = 2\pi\tau_g$, we calculate $R = 2/(\pi f_0 \tau_g)$, actual R in MHz $R_{MHz} = 2/(\pi\tau_g)$, and normalized R: $\bar{R} = R_{MHz}/\Delta f$. To find the correct *htapin* position, the driven modal analysis is used to find the group delay for an interval of *htapin* position divided by steps. The group delay is used to calculate R_{MHz} or $\bar{R}$, and interpolation is used to find the exact position of the tapped-in stripline. For this 5-pole filter, this procedure yielded $htapin = 161.9$ mils.

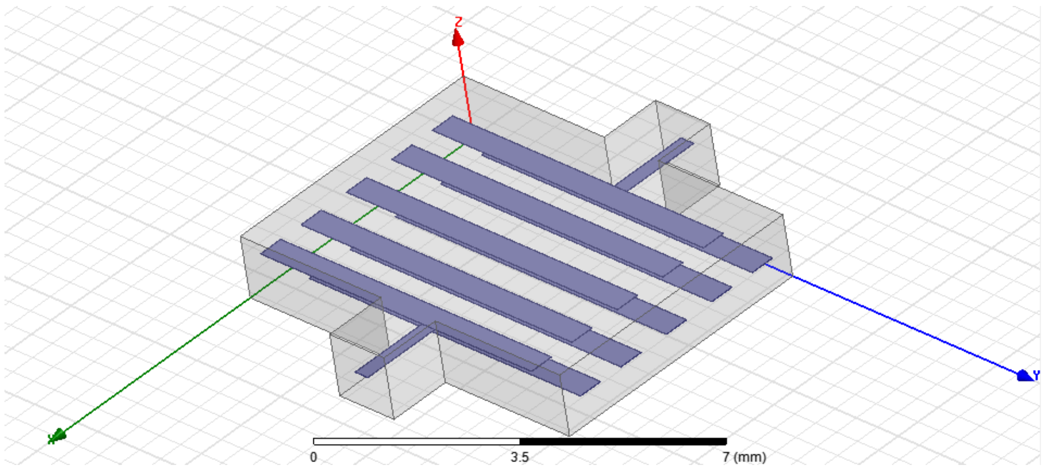


Figure 15. 5-pole filter, unoptimized

Now it was time to assemble the complete filter, find its initial response using driven modal solution, and tune it using HFSS until it was ready for optimization. This is not a trivial process. There are a total of 6 parameters that are simultaneously optimized: resonator overlaps lc_1, lc_2, and lc_3 for the the outer, middle, and inner resonators, respectively; corresponding to the inter-element separations between resonators 1-2 & 4-5 (sep_1), and 2-3 & 3-4 (sep_2), and tap-in position *htapin*.

Figure 16 shows a filter that is ready for tuning. Figure 17 shows the same filter after it was successfully optimized.

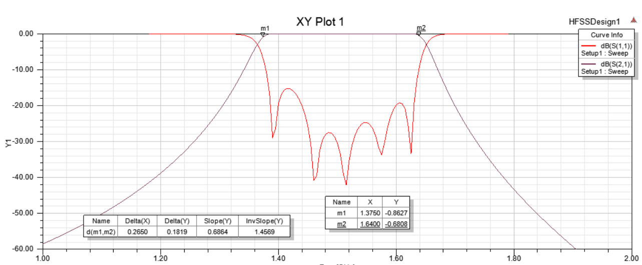


Figure 16. S_{11} after tuning

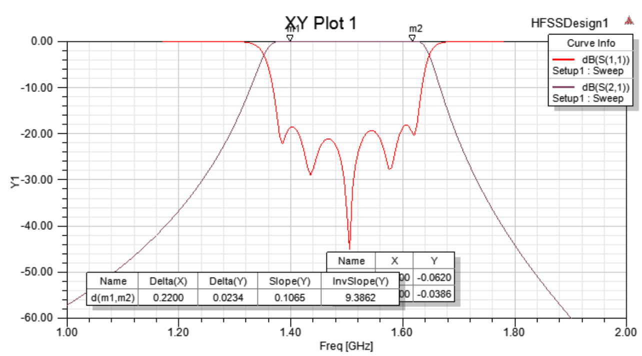


Figure 17. S_{11} after optimization

The penultimate step was to add material losses and rerun the simulation. The LTCC material properties were edited to include a loss tangent $\tan\delta = 0.001$. The metallization of the striplines was changed from PEC (perfect electric conductor) to copper.

Figure 18 shows how the performance of the filter degraded after adding the losses. Slight optimization was needed to give the response shown in figure 19. The final parameters of this filter are tapped in position $htapin = 175.9$ mil; overlaps $lc_1 = 229.7$ mil, $lc_2 = 211$ mil, and $lc_3 = 207.7$ mil; separations $sep_1 = 20.77$ mil,

$sep_2 = 27.09$ mil; resonator length $lr = 300$ mil; height $b = 58.83$, mil, stripline width $ws = 20$ mil, and tapped in line width $wtapin = 14$ mil. The edge distance from the resonators to the walls is $b/2 = 29.42$ mils.

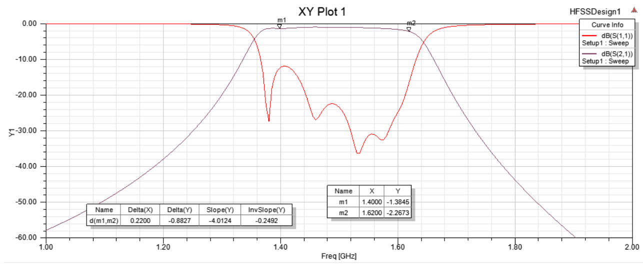


Figure 18. S_{11} after adding losses

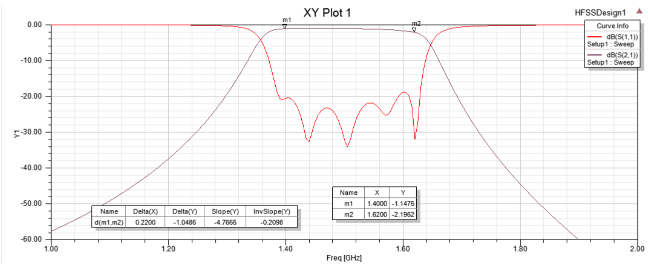


Figure 19. S_{11} of finalized filter

7. CONCLUSION

A systematic design procedure has been demonstrated to realize a miniature narrowband combline filter composed of double-layer coupled stripline resonators. Such filters need to meet very exacting specifications required for high-performance communications applications. To meet these requirements, a rigorous synthesis and modeling procedure was used to find initial specifications. Then, careful tuning and optimization was used to find minima of the cost function subject to the desired goal of -20 dB return loss in the passband. We are able to develop very compact microwave assemblies by manufacturing the filter in low-temperature co-fired ceramic. The same steps are carried out for each of 16 filters ranging from 1.5 GHz to 6.5 GHz. Switched filter banks are useful devices, which find applications in areas such as in wideband receivers, test sets, and front-end multiplexing.

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