Narrowband elliptic filters on microstrip

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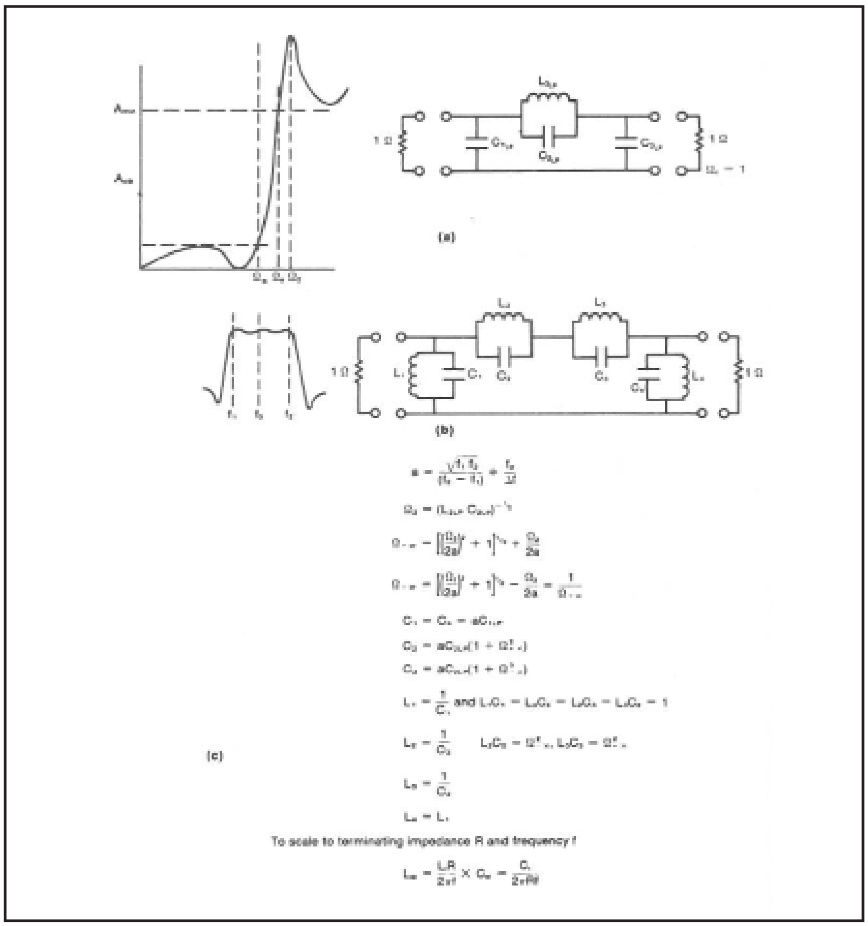


***1. Selectivity, minimum stopband attenuation, and maximum ripple or return loss** determine the lowpass prototype (a). The bandpass prototype (b and c) is calculated from the lowpass prototype.*

ELLIPTIC bandpass filters generally show lower loss and better selectivity than Chebyshev filters that have an equal number of resonators. Thus, they would seem well suited for microstrip applications where the loss inherent is low-Q microwave resonators makers Chebyshev filters a poorer alternative. However, because of the difficulty in realizing practical impedance levels, elliptic bandpass filters are not widely used in microstrip.

Adapting a design method previously given for narrow-bandwidth n(<10 percent) TEM elliptic filters,[1] we found that with appropriate modifications

> The difficulty of realizing practical impedance levels once kept elliptic bandpass filters off of microstrip. But a new design has been adapted from the literature.

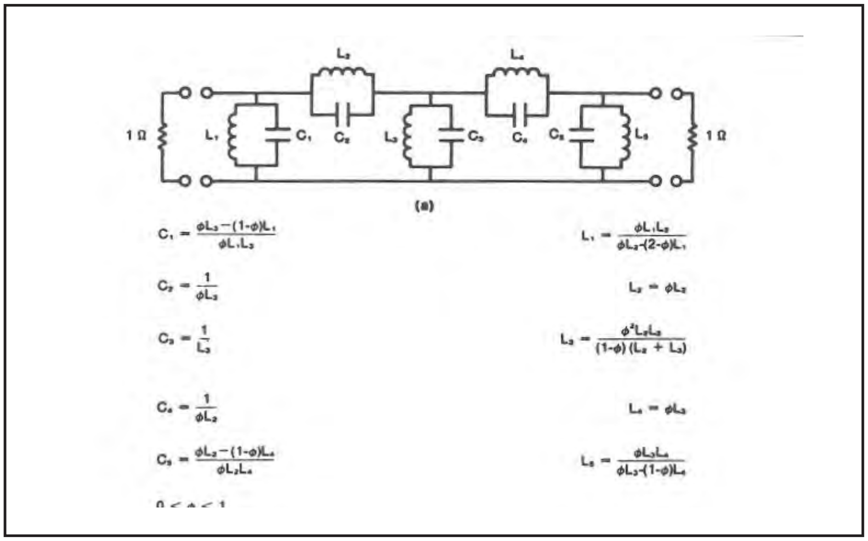
and precautions we could design the desired filters from the lowpass prototype values available in the literature. Furthermore, we found that the resultant filter structures were easy to analyze. In addition, the choice of an appropriate substrate allows the actual filter behavior to follow the theoretical response closely over wide bandwidths.

To demonstrate the overall design process, it is helpful to examine the methodology employed for a filter in which n = 3. The process naturally applies to higher-order designs as well. The five steps involved are

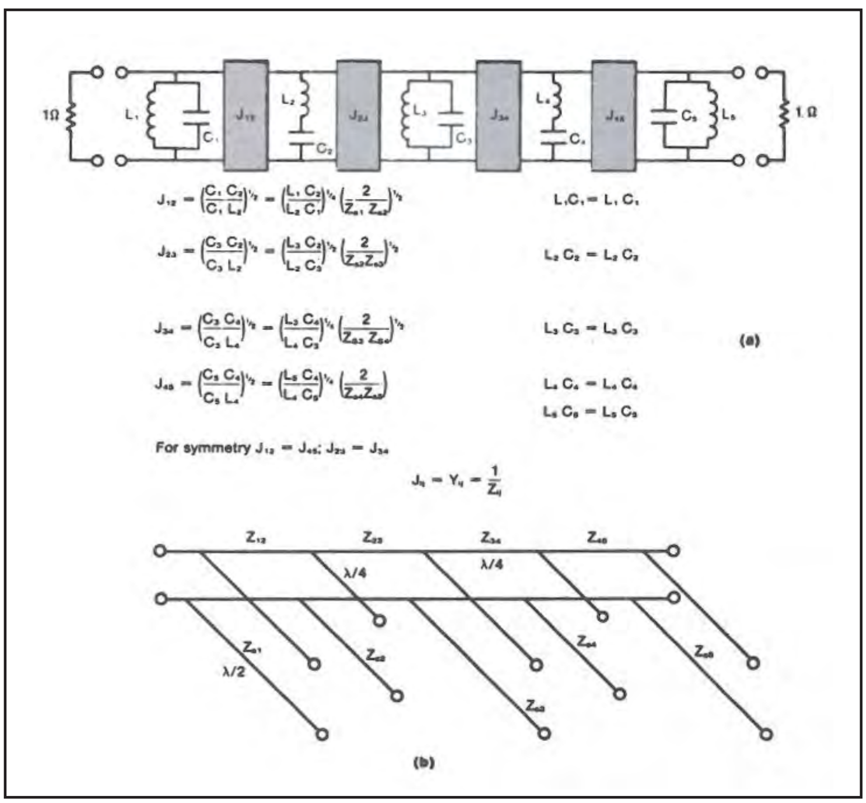
1) selecting the appropriate lowpass prototype,
2) converting the lowpass prototype to a bandpass response ($\omega c = 1$; RL =1),
3) transforming the bandpass circuit to the appropriate form,
4) converting all series resonators to shunt resonators using unit elements, and
5) transforming the resonators to equivalent transmission lines, scaling the response to the required frequency and impedance values (typically 50 ohms), and calculating the input and output couplings. Iteration may be re-

Table 1: Bandpass elements

Element Number	Unscaled L	Unscaled C	Scaled L (nH)	Scaled C (pF)	Freq. (GHz)
1	0.11808	8.4688	0.07517	2.15656	12.5
2	0.28932	4.2879	0.18418	1.09190	11.2227
3	0.23215	3.45639	0.14846	0.88016	13.9226
4	0.11808	8.4688	0.07517	2.15656	12.5



2. A more appropriate form for microstrip fabrication *is created by transforming the bandpass prototype into a five-resonator network.*



3. The model's series resonators are converted *to shunt resonators by admittance inverters approximated as one-quarter wavelength transmission lines. Inductances and capacitances need not be calculated in this step.*

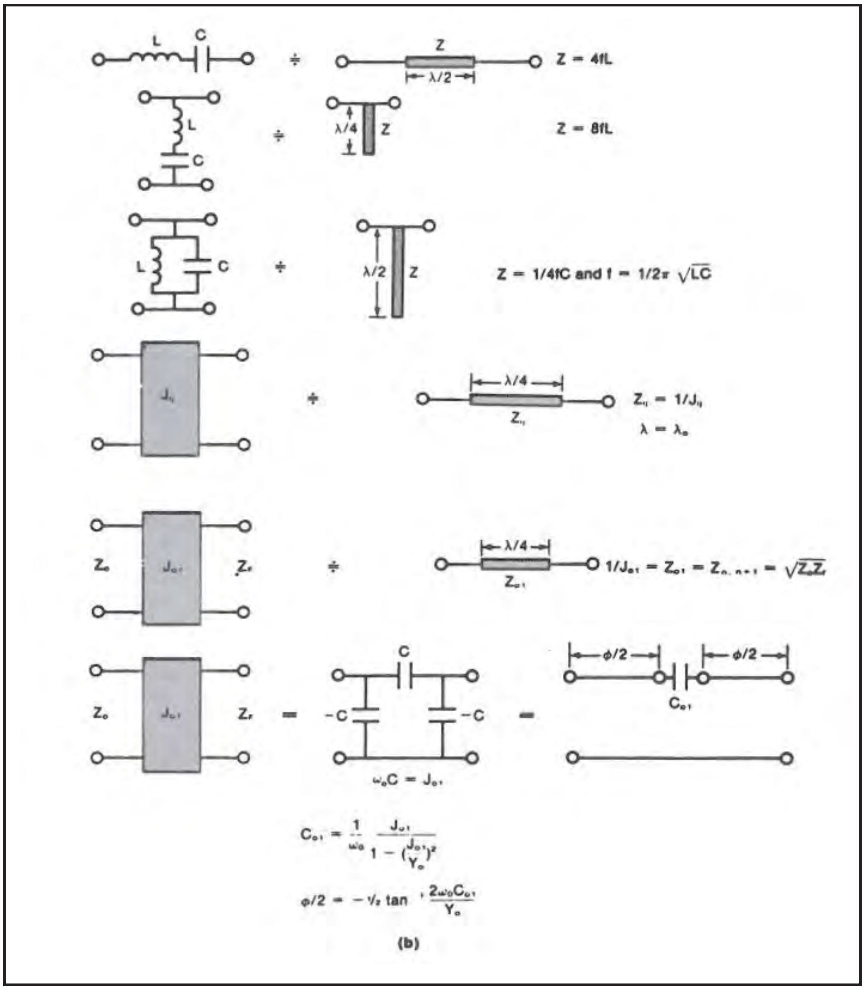
quired between Steps 4 and 5 to obtain practical impedances.

This five-step process may at first seem needlessly cumbersome. Additionally, Step 4 narrows the filter bandwidth somewhat, and some passband response degradation may similarly be noted. However, unlike the process on which it is based, this method yields a filter symmetrical with regard to impedances for the n = 3 case. Another advantage is that the required calculations are straightforward and readily programmed for a small computer. The problems created by Step 4 can be dealt with by checking the filter performance (including loss) against an analysis program prior to etching the filter.

The lowpass prototype is chosen on the basis of the required selectivity, minimum stopband attenuation, and maximum permissible ripple or return loss. With the lowpass prototype selected (Fig. 1a), the bandpass prototype is calculated (Figs. 1b and 1c). However, the basic form, with two series stopband resonators, is unsuitable for microstrip. The next step, therefore, is to transform the network in Fig. 1b to a five-resonator form as in the Geffe transform (Fig. 2). Following this transform the series resonators are converted to shunt resonators using admittance inverters approximated as unit elements, i.e.,$\lambda o/4$ line lengths (Fig. 3). The circuit is then converted to microstrip form with the given relationships (Fig. 4). Finally, the input/output inverters are selected to match the filter impedance to the terminating lines.

Using the lowpass-to-bandpass mapping discussed (Fig. 1a), unit elements reduce the bandwidth and increase the passband ripple. The stopband performance remains largely unaffected. The ideal response can be attained only if ideal inverters are employed instead of the quarter-wave lines. Since microstrip line loss also affects passband performance, required design revisions should be made only after actual microstrip circuit analysis. Thus, the final microstrip performance can be optimized to more closely approximate the prototype response.

There is some latitude in selecting unit element impedance. While all



4. The process is completed *by converting the circuit to microstrip form and then selecting input/output inverters to match the filter impedance to the terminating lines.*

Table 2: Transformed elements					
Element Number	Unscaled		Scaled		Freq.
	L′	C′	L (nH)	C (pF)	
1	0.16221	5.61024	0.10327	1.42863	13.10297
2	0.17359	7.14648	0.11051	1.81983	11.2227
3	0.11621	8.60474	0.07398	2.19117	12.50
4	0.13992	5.76063	0.08908	1.46634	13.9226
5	0.17824	6.16454	0.11347	1.56978	11.9248

Table 3: Shunt resonators and unit elements (f_o = 12.5 GHz, R = 50 ohms)

	L″	C″	Z_J (ohms)	L (mH)	C (pF)	Z (ohms)	Freq. (GHz)
1	0.16221	5.61024		0.10327	1.42863	13.355	13.10297
12			0.225			11.25	12.5
2	0.36179	3.42897		0.23032	0.87182	20.679	11.2227
23			0.275			13.25	12.5
3	0.17360	5.76019		0.11052	1.46682	13.634	12.5
34			0.275			13.25	12.5
4	0.29163	2.76409		0.18565	0.70385	20.679	13.9226
45			0.225			11.25	12.5
5	0.17824	6.16454		0.11347	1.56978	13.355	11.9248

unit elements may have equal impedances, practical considerations suggest that impedances of the inner unit elements should be different from the values of the outer unit elements. Both the input and output inverters can be quarter-wave transformers of capacitive gap inverters. For filters with narrower bandwidths (<5 percent), filter impedance is generally too high to consider quarter-wave transformers. Thus, capacitive gap inverters are desirable in the narrow-bandwidth filter.

A detailed design example will illustrate the complete design process.

First, the design objectives are established: a center frequency of 12.5 GHz, a nominal 1-GHz bandwidth, minimum stopband attenuation of 25 dB with at least 30-dB attenuation 1.2 GHz from the center frequency. For a filter with n=3 meeting the minimum return loss (or maximum ripple) requirements specified, a suitable prototype can be obtained from standard tables.[3] The particular prototype selected has the following element values:

- Ripple ≈ 0.04 dB,
- Return loss ≈ 20dB,
- C1LP = C3LP =0.6775,
- C2LP = 0.1531,
- L2LP= 0.8962,
- Ω2 = 2.699;

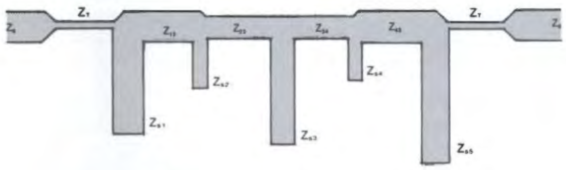
with a ≈fo /Δ f= 12.5

and, then, Ω+∞ =1.11381, Ω-10 = 0.89782.

Following the procedures detailed above (Figs. 1, 2, and 3), the element values are calculated (Table 1). Although in practice it is unnecessary to scale the elements until the final design stage, the scaled values are listed in Table 2 (RL =50 ohms, f0 =12.5 GHz) for their instructive benefit only.

The value of Φ =0.6 was selected for the example above, although other values might have been chosen. Note the middle resonator is the only one which is resonant at the center frequency (12.5 GHz), and the frequencies of the attenuation poles are unaltered by this transform. This transformed circuit has the ideal characteristics of the resonate lowpass prototype: BW = 8 percent, ripple = 0.04 dB, and return loss = 20 dB.

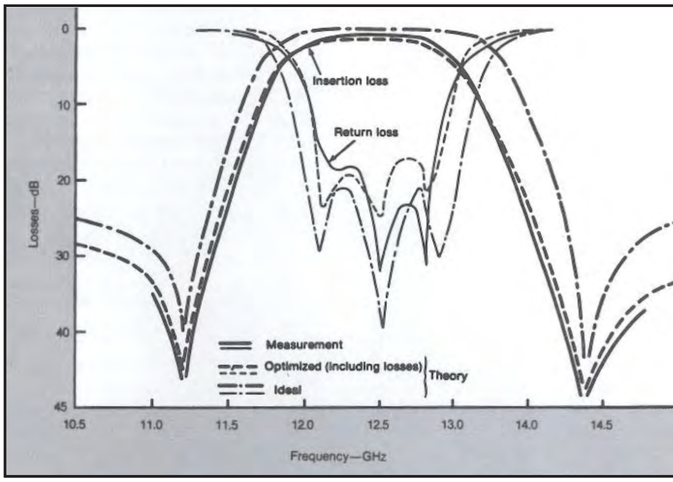
The next step converts the series resonators to shunt resonators using unit



	Imp.	Phase (deg.)	Freq. (GHz)
Z_T	100	90	12.5
Z_{e1}	53.42	176.9	13.103
Z_{e2}	82.72	90	11.223
Z_{e3}	54.54	180.4	12.5
Z_{e4}	82.72	90	13.923
Z_{e5}	53.42	183.8	11.925
$Z_{12}(Z_{45})$	45	90	12.5
$Z_{23}(Z_{34})$	55	90	12.5

$Z_o = 50\ \Omega$

***5. An example calculation** shows a circuit with an impedance of 200 ohms. Lower line impedances increase the uncertainty of assessing and compensating for junction effects, while higher values may make the quarter-wave transformers unrealizable.*



***6. Optimized and measured values** for the passband results of an ideal filter agreed well. A comparable three-element Chebyshev filter showed over twice the loss.*

elements (Fig. 3). The step concurrently begins the divergence from the ideal performance of the prototype to the real performance of a practical design. Since the impedances can be determined directly, it is unnecessary to calculate the inductive and capacitive values. These values are included in Table 3 for completeness only.

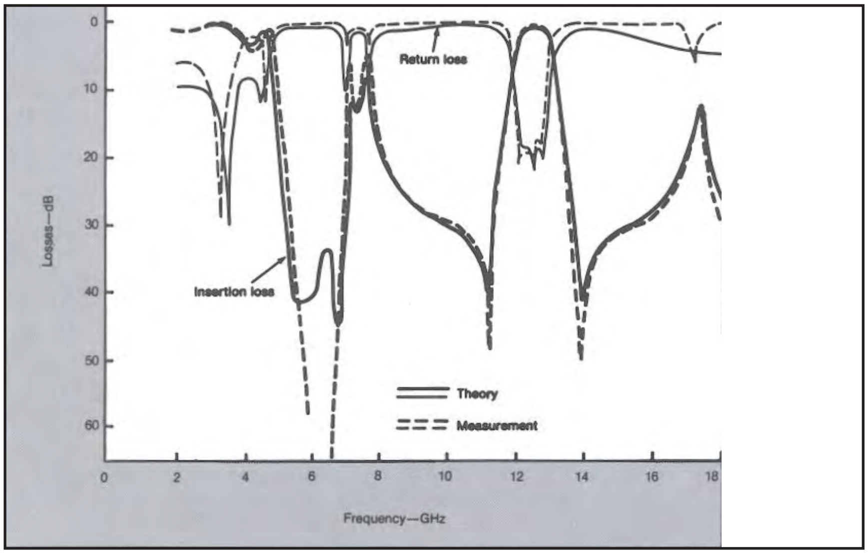
After some trial and error, the values of C"1 in Table 3 were set to C'1 and the impedances were selected as 0.225 ohms and 0.275 ohms, respectively. Note that the resonant frequency of each LC resonator is used in the calculation (based on the equations in Fig. 4) rather than on the center frequency. The impedance levels in themselves are much less significant than the ratio of the maximum to the minimum. The filter should be realizable in a practical microstrip design if the ratio is less than 3.5:1. This ratio can be as large as 7:1 if two parallel stubs are used for the low-impedance shunt elements. Subsequently, the appropriate scaling may be accomplished. These values will be combined with output inverters to match the filter to the microstrip transmission line.

In the above example, the impedance was set to 200 ohms. This value yields moderate impedance values throughout the filter, together with a convenient value (100 ohms) for the quarter-wave transformers used to approximate the input and output inverters. The final element values are shown in Fig. 5. While a lower value of filter impedance – such as 150 ohms – might have been chosen, lower line impedances increase the uncertainty in assessing and compensating for junction effects. Conversely, had a higher value of filter impedance been chose, say around 350 ohms, the quarter-wave transformers would have been unrealizable. The quarter-wave transformers have a minimal effect on filter performance up to bandwidths of about 10 percent.

However, for narrow-bandwidth filters, say 5 percent or less, the required filter impedance might be several thousand ohms, which would promote realizable line impedances. In these cases a more suitable solution would employ a capacitive gap inverter, as follows: At 12.5 GHz and for J01 = 0.01 (ZF = 200 ohms, Z0 =50 ohms), the value of C01 is 0.16976 pF and $\Phi/2$ = - 26.57 deg. The negative input and output line length is inconsequential, since for the filter the first resonators can be converted to series resonators (Fig. 4a).1 Thus, the negative line lengths can be absorbed into the resonator itself. The input and output resonators then become series resonators of length (prior to optimization) 152.1 deg. (13.103 GHz) and 154.6 deg. (11.925 GHz).

Note that the phase length of 26.57 deg. At 12.5 GHz has been scaled to the relevant length for each resonator frequency. The capacitive gap inverter is represented by a series capacitor embedded in two negative line lengths as shown in Fig. 4b. The input and output resonators are converted into series resonators (Fig. 4a), and these resonators absorb the negative line lengths on the filter side. The negative line lengths on the input and output are inconsequential, since these are simply part of the 50-ohm input and output lines. In the example above, the filter was scaled to 187.2 ohms. The theoretical performance of this filter with capacitive gap inverters is superior to that which it would have been with quarter-wave coupling because of the frequency-invariant nature of the capacitive gap coupling.

The value of C01 will decrease with reduced filter bandwidths, but in many cases will remain too large to be realized with microstrip line gaps. However, commercially available chip capac-



7. The measured filter response tracks *the predicted response of the filter very closely over the full 2-to-18 GHz bandwidth. Experimental results often diverge from predicted results at frequencies outside the passband.*

itors come in a wide variety of values. Alternatively, simple overlay capacitors using dielectric spacers might be tried, although reproducible results may be rare. Thus, some manner of adjustment might be required after etching. A further alternative for input and output coupling is tapped resonator commonly used for interdigital and hairpin filters.4,5 However, experimental results remain unavailable for this method of realizing the microstrip elliptic filter.

Improving return loss

The microstrip filter derived above had a bandwidth of 800 MHz and minimum return loss of only 12 dB. Computer optimization was performed on the passband resonator phase lengths to improve the return loss (and amplitude ripple) across the filter passband. Subsequent to optimization, the return loss was improved to more that 17dB. Since etching and substrate tolerances would add to overall variability, further computer optimization was deemed unnecessary.

The actual filters were etched on 0.25-mm CuFlon (Polyflon Corp., Norwalk, CT) clad with 0.02mm copper. Using a pure PTFE substrate, CuFlon has notably lower loss than the fiber-based substrates. This is especially true above 10 GHz. Allowances were made in the circuit layout for T-junction and open-circuit effects using methods prescribed in the literature.6,7 The step and T-junction analyses are not precise, so some trimming was anticipated. For the bandstop resonators the lengths are easily and independently adjusted by observing the frequencies of the transmission zeros. The bandpass stubs are adjusted interactively to achieve the maximum return loss. In practice the T-junction analysis yielded reasonably accurate results typically in the 0.25-mm range. Tuning was therefore relatively easy.

Also of interest are the passband response of the ideal filter with no inverters or loss, the optimized response including line loss, and the measured response (Fig. 6). As may be seen, the agreement between the optimized and measured values is quite good for both transmission and reflection. Considering the 0.5=dB loss in the connectors and interconnecting lines, the 1.0-dB loss of the filters is quite favorable. A three-element Chebyshev filter on Duriod9 at 11.5 GHz with comparable bandwidth but reduced selectivity exhibited over twice the loss.

In addition to the in-band response performance, the measured filter response very closely tracks the predicted results over the full 2 to 18 GHz bandwidth (Fig. 7). This result is significant, since it is more often the case that experimental results diverge from predicted results with increasing distance from the passband. Radiation appeared minimal as verified by moving an absorber around the filter and changing the filter enclosure. Neither change has major effects on the filter response.

The design can readily be extended to odd higher-order filters (n = 5, 7 etc.) by applying the method given above to each set of resonators in turn. The bandpass circuit of an n = 5 filter, for example, comprises four shunt resonators in series and three shunt resonators in parallel. Applying the Geffe transform twice converts these to four shunt resonators in series and five in parallel. The inclusion of unit element (eight of them in this case) will convert the four shunt resonators in series to four series resonators in parallel. Since the original prototype is not symmetrical, symmetry will generally be absent in the final filter structure as well.

For an even number of elements, an addition series resonator in series will be obtained from the lowpass prototype filter. This can be approximated as a half-wave section of line in series or converted to a shunt resonator in parallel using unit elements. However, the filter impedance levels will be asymmetrical. ••

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