Design of an Evanescent Mode Circular Waveguide 10 GHz Filter

NI AWR Design Environment, specifically Microwave Office circuit design software, was used to design the filters for a range of bandwidths up to five percent. The ability provided by the software to optimize the design while simultaneously accounting for resonator spacing, insertion loss, and return loss, was powerful.

Ham radio operation at 10 GHz is far removed from global short-wave communication typically operating below 30 MHz. The motivation for designing the evanescent mode circular wave-guide 10 GHz filter sprang from a desire to extend ham radio operation to 10 GHz. This article describes the design of two 10 GHz filters and is a follow on to a previous application note (Design of a 10 GHz Low-Noise Amplifier for Amateur Radio Operation). Both are key components needed for ham radio station operation at 10 GHz.

Test Bench Set Up

Before any hardware could be tested, measurement capability needed to be extended to 10 GHz. Available equipment included a vector network ana-lyzer (VNA), spectrum analy-zer, and service monitor, which provided metrology through L-band, but did not support 10 GHz measurement capability. The designer decided to build a synchronous up/down conver-
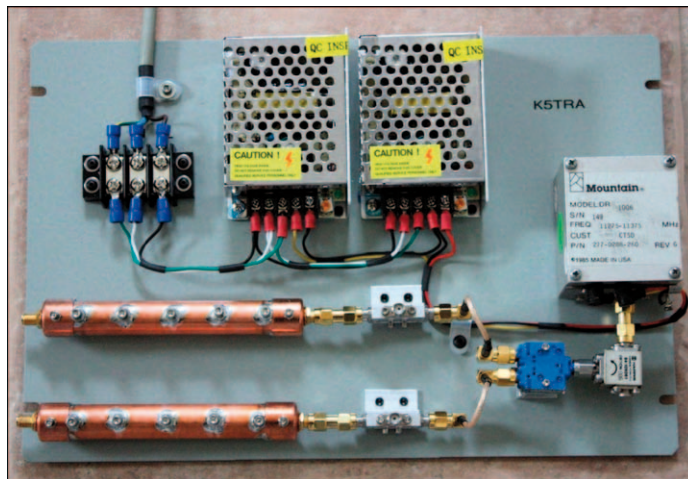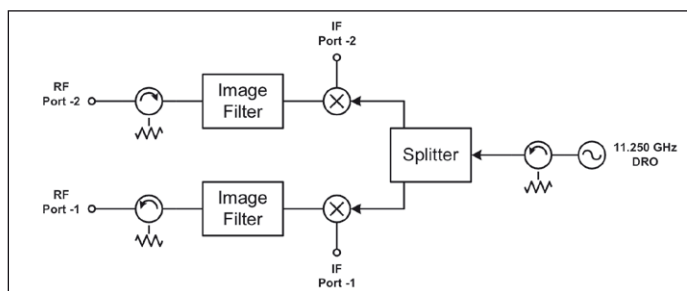
Figure 1: 10 GHz test set photo

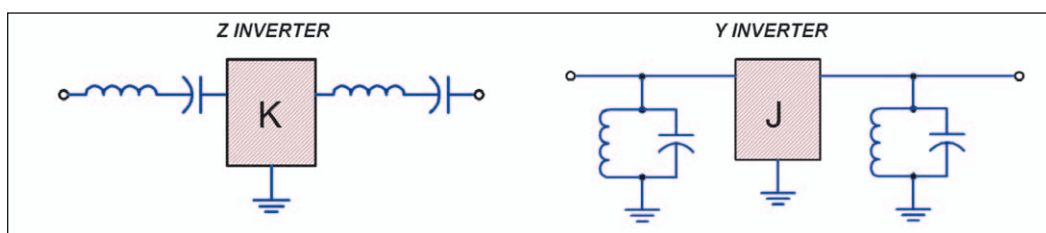**Figure 2: Test set block diagram**

Figure 3: K and J inverters in BP filter segments
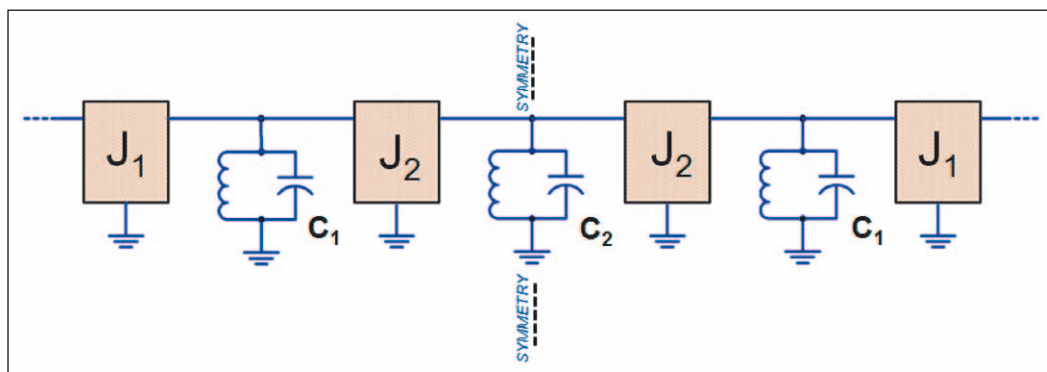

Figure 4: Three resonator BP filters with J inverters and shunt resonators

sion test set to extend the bench equipment to 10 GHz. To build this test set, image reject filters (Figure 1) were needed. Figure 2 illustrates the block diagram of the test set, which enabled measurement of insertion gain or loss, as well as return loss with a directional coupler.

Bandpass Filters and Immittance Inverters

Half-inch copper tubing seg-ments with end caps were used to realize tunable bandpass filters at 10 GHz. To better understand how this type of filter works, some basic filter theory is useful.

Many filters can be represented as cascades of series and shunt branches. In the passband fre-quency range, series branches are ideally short circuit and shunt branches are open, which yields 0 dB insertion loss. Alter-natively, in the stopband, series branches are ideally open and shunt branches are short circuit, which yields no transmission and 100 percent reflection. For

example, low-pass (LP) filters are formed from series inductor and shunt capacitor cascades and bandpass (BP) ladder canonic structures are formed from ladder cascades of series resonators and shunt resonators.

Bandpass filters obtained from LP to BP transformation produce these structures with all resonators synchronously tuned to the center of the pass band. LC ladders composed of alternating series and shunt resonators are quite far removed from waveguide filters operating at 10 GHz, requiring two conceptual steps to bridge the gap.

The first step is the introduction of impedance (K) or admittance (J) inverters. Both filter structures in Figure 3 can be viewed as equivalent to the series-shunt LC cascade.

An impedance inverter connected to a series resonator behaves as a shunt resonator when viewed from the opposite side of the inverter. Similarly, an admittance inverter connected to a shunt resonator behaves as a series resonator when viewed from the opposite side of the inverter. Impedance or admittance inverters (sometimes called immittance inverters) are symmetric networks where shorts map to opens and opens to shorts, series capacitors map to shunt inductors, and series inductors map to shunt capacitors.

The use of immittance inverters enables reuse of the same type of resonator in an overall bandpass filter design, as illustrated in Figure 4. For example, combline filters can be viewed as cascades of shunt resonators and admittance inverters, where the inter-resonator coupling is set by the particular admittance of the inverter between a pair of resonators. Filters with the same input and output port impedances are usually symmetric. This is shown in Figure 4. This property can be used to simplify the design parameters.

J and K inverters can be realized in many ways. Quarter-wave transmission lines are a common example of an impedance (or
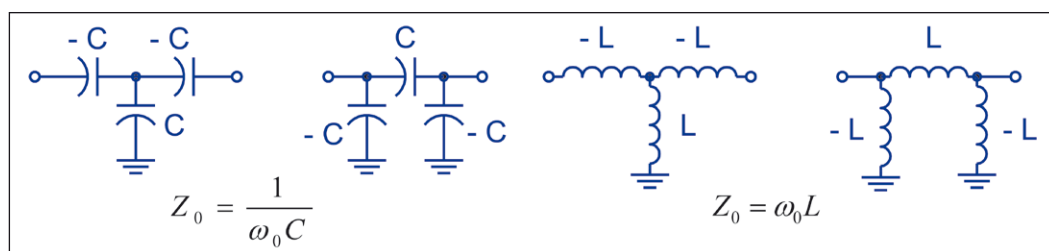

$$Z_0 = \frac{1}{\omega_0 C} \qquad Z_0 = \omega_0 L$$
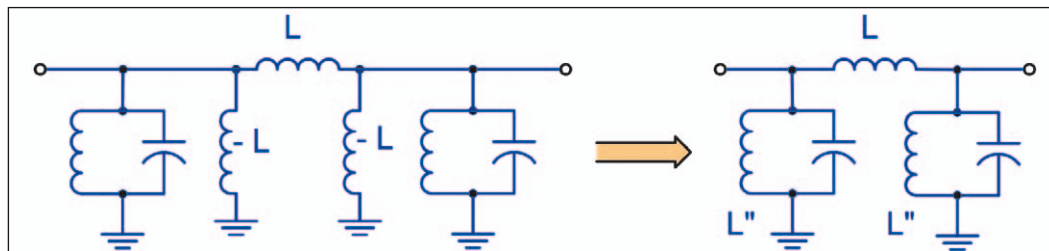
Figure 5: LC realizations of K and J inverters

Figure 6: Absorption of J inverter negative elements into resonator tuning

admittance) transformer placed between similar resonators. Inverter forms commonly used in LC filter designs are shown in Figure 5, where the T networks are the impedance inverter representations and the Pi networks are the admittance forms.

Note that a T to Pi transformation easily shows that they are equivalent. The T inverter networks forms are better suited to interface with series resonators and the Pi representation is better suited to interface with shunt resonators. The obvious question now arises: Where are negative valued inductors or capacitors found?

It is clear that inverters will be placed between shunt resonators. The left side of Figure 6 shows a pair of shunt resonators coupled with an inductive admittance inverter. The inverter negative shunt inductance can be absorbed into the resonator representation. This results in a parallel equivalent inductance larger than the original, which is represented as L" in the equivalent circuit on the right side of Figure 6.

It may seem like the resonant frequency has been lowered. It is important to note that when the adjacent resonator is short circuited, the resonant frequency is the same as the original resonator. When one side of an inver-

ter is shorted, the other side is open. This is also true for interior resonators with inverters on both sides, as in the center resonator in Figure 4. When the adjacent resonators on both sides are short circuited, the interior resonator tunes the center frequency of the passband. In this way, all resonators are seen to be tuned to the center frequency. Filters of this type are called synchronously tuned BP filters. The relationship is a direct result of the canonic BP ladder structures.

Recognition of the resonant frequency behavior of each resonator when adjacent resonators are shorted will be important later in this note, greatly simplifying the design process. It should also be mentioned that this behavior is the basis for Dishal's tuning method4.

Circular Waveguide

The second step is to address the waveguide equivalent circuit. Metal pipes of various cross-section shapes have been used as low-loss transmission media at microwave frequencies for many years. Circular cross-section copper pipe is the medium of interest here. As long as the RF frequency is above a critical frequency related to the pipe radius, RF will propagate. A half-inch copper pipe is actu-

ally 0.565" inside diameter. The corresponding TE11 cutoff frequency is 12.2516 GHz. Below the cutoff frequency, RF field amplitude falls exponentially with distance and the transmission path becomes reactive. This is called evanescent mode propagation.

The equivalent circuit for a length of waveguide below the cutoff frequency can be seen in Figure 7. The possibilities for using a below-cutoff waveguide to form admittance inverters now become apparent. Figure 8 shows the equivalent circuit redrawn on the left side.

Since inductive admittance inverters with series branch inductance of LS also have -LS shunt branches, additional shunt branches of +LS and -LS can be added with no net change (they cancel each other). This trick enables the recognition of the segment of copper pipe that can be viewed as an inductive admittance inverter with shunt inductors on both ends. Tuning screws can be added in the E plane to introduce shunt capacitance, thereby creating the shunt resonators. For less inter-resonator coupling, the length of the pipe segment between tuning screws can be increased (LS is increased).
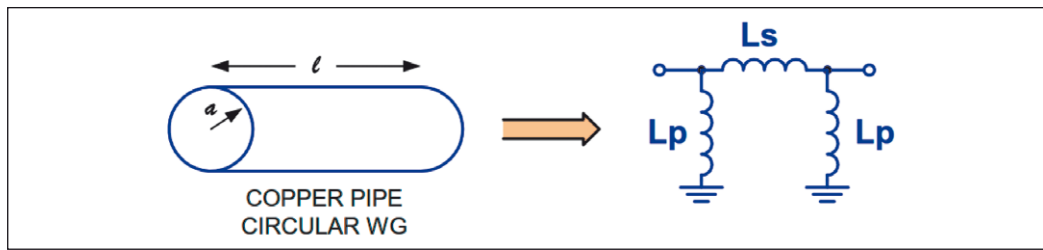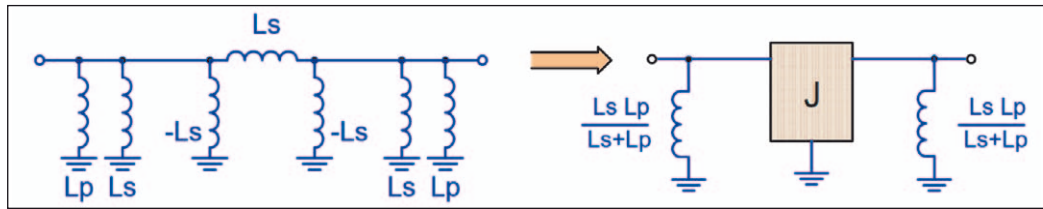
Figure 7: Below cutoff equivalent circuit.

Figure 8: Below-cutoff equivalent circuit - alternate interpretation

The equations in Table 1 describe the circuit behavior of a segment of half-inch diameter copper pipe.

These are numeric values for the circuit representations found in Figures 7 and 8. Clearly, by changing the length, the series inductance and thereby the coupling (admittance of inverter) can be set. The excess shunt inductance of Figure 8 can be resonated by adding a tuning screw in the E plane. This provides a variable capacitance to form the shunt resonator.

Design Procedure

The preceding paragraphs provide a procedure for analyzing a length of circular waveguide operating below cutoff. To design a filter in this media, the designer could start from LP prototype tables, perform an LP to BP transformation, and map the series resonators into shunt resonators cascaded with inverters. This is the classic approach found in Craven and in Howard and often involves some iterative optimization.

Another approach might be to fully describe a physical filter with inter-resonator lengths and tuning capacitors as variables in a circuit analysis environment and simply use optimization to obtain a solution. This is a somewhat forceful approach and might not yield an optimum solution.

If some knowledge of desirable filter solutions is used to simplify and constrain the structure, circuit analysis with optimization can efficiently obtain a solution that is well conditioned to converge to a desired solution. Recall the symmetry shown in Figure 4. If the physical structure and thereby the circuit description is explicitly defined to be symmetrical, then the number of variables is cut in half. Also, recall the mention of Dishal's tuning method for synchronously tuned filters. Each resonator should be tuned to the center frequency when all adjacent resonators are short circuited. This constraint provides a closed form solution for all tuning capacitors as a function of surrounding inductors and eliminates the capacitors as variables, providing a significant advantage in preconditioning the optimizer for a desired solution. Ultimately, the designer decided to use this approach for the filter design project.

NI AWR Design Environment, specifically Microwave Office, was used for the circuit analysis and optimization. Circuit values were all described algebraically within the software, a feature that enabled easy implementation of symmetry and synchronous tuning variable elimination.

The cutoff frequency, Fc, of a circular wave guide is a function of the guide's inside radius, a. The measured inside diameter of a half-inch copper pipe actually is 0.565", so a = 7.1755 mm.

$$\textit{Cutoff Frequency:} \quad F_c = \frac{1.8412\, C}{2\pi a} = \frac{(1.8412)(300)}{2\pi(7.1755)} \approx 12.252\; GHz$$

$$\textit{Center Frequency:} \quad F_0 = 10.4\; GHz$$

$$\textit{Evanescent Mode Propagation Constant:} \quad \gamma = \frac{\omega_0}{C}\sqrt{\left(\frac{F_c}{F_0}\right) - 1} \approx 0.1356$$

$$\textit{Wave Impedance:} \quad X_0 = \frac{120\,\pi}{\sqrt{\left(\frac{F_c}{F_0}\right) - 1}} \approx 605\; \Omega$$

The equivalent circuit for a length, l, of half-inch copper pipe can be calculated as follows:

$$\textit{Series Inductance:} \quad L_S = \frac{X_0 \sinh(\gamma l)}{2\pi F_0}$$

$$\textit{Shunt Inductance:} \quad L_P = \frac{X_0 \coth\left(\frac{\gamma l}{2}\right)}{2\pi F_0}$$

Table 1: Behavior of a segment of half-inch diameter copper pipe

```
Zn=50      (port reference Z)
Qo=200

a=7.1755 (radius mm)
Fc=1.8412*300/6.2832/a
Fc: 12.25

Fo=10.4    (GHz)
b=(6.2832*Fo/300)*sqrt((Fc/Fo)^2 -1)
b: 0.1356

X0=377/sqrt((Fc/Fo)^2 -1)
X0: 605.4

l1=5.47674781483291            l2=21.9933981090275
b: 0.1356                      b: 0.1356
XLs1=X0*sinh(b*l1)             XLs2=X0*sinh(b*l2)
XLs1: 492.3                    XLs2: 5963
Ls1=XLs1/6.28/Fo              Ls2=XLs2/6.28/Fo
Ls1: 7.537                     Ls2: 91.3

XLp1=X0*cosh(b*l1*0.50)/sinh(b*l1*0.50)  XLp2=X0*cosh(b*l2*0.50)/sinh(b*l2*0.50)
XLp1: 1704                     XLp2: 670
Lp1=XLp1/6.28/Fo              Lp2=XLp2/6.28/Fo
Lp1: 26.1                      Lp2: 10.26

Lc1=1/(1/Ls1+1/Lp1+1/Ls2+1/Lp2)   Lc2=1/(2/Ls2+2/Lp2)
C1p=1000/(4269.9856*Lc1)           C2p=1000/(4269.9856*Lc2)
C1p: 0.06544                       C2p: 0.05079

x1=l1/25.4                     x2=l2/25.4
x1: 0.2156                      x2: 0.8659
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Figure 9: Algebraic description in Microwave Office - |1 and |2 are tune variables (in mm).

The only independent variables were the distances between tuning screws. These are shown as l1 and l2 in the example algebraic description in Figure 9. All inductors and capacitors were calculated in closed form from these two values.

Figure 10 shows a corresponding circuit description. This is a three-resonator filter with tuned port transitions. Note that loss has been introduced through the inductor unloaded Q parameter, Q0. After measuring a single unloaded resonator, the unloaded Q was empirically determined to be approximately 200.

The subminiature version A (SMA) transition can be seen in Figure 11. A shorted loop forms an H plane transition. This launcher is tuned with a shunt capacitance from a tuning screw. For simulation purposes, the port transition tuning was expressed was expressed as C0p. It was calculated using Dishal's method. This can be seen algebraically as:

Lc0 = 1/(1Ls1 = 1/Lp1)

C0p = 1000/(4269.9856 Lc0)

C0p:0.04005

Practically, the end transition tuning will differ slightly due to the H plane transition inductance. The optimized swept response for this example filter is illustrated in Figure 12. This is a five percent bandwidth (BW) solution centered on 10.4 GHz.

Results

A set of solutions (Table 2)was run over BW steps for 2, 3, 4,

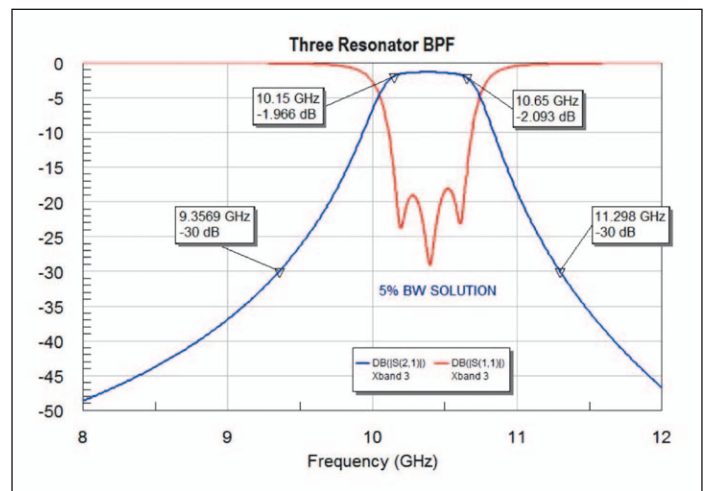*Figure 11: Five-resonator BP filter with tuned H-plane launchers*
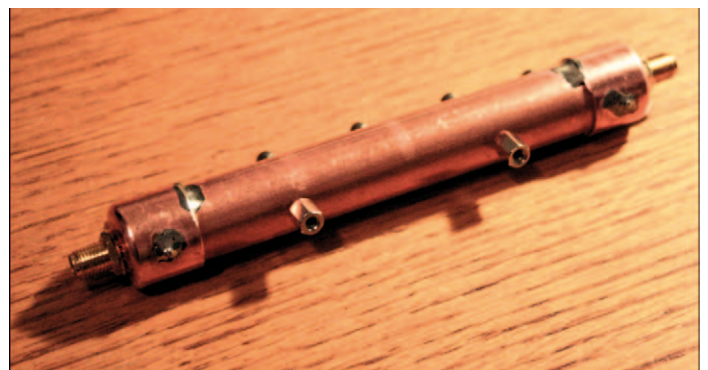

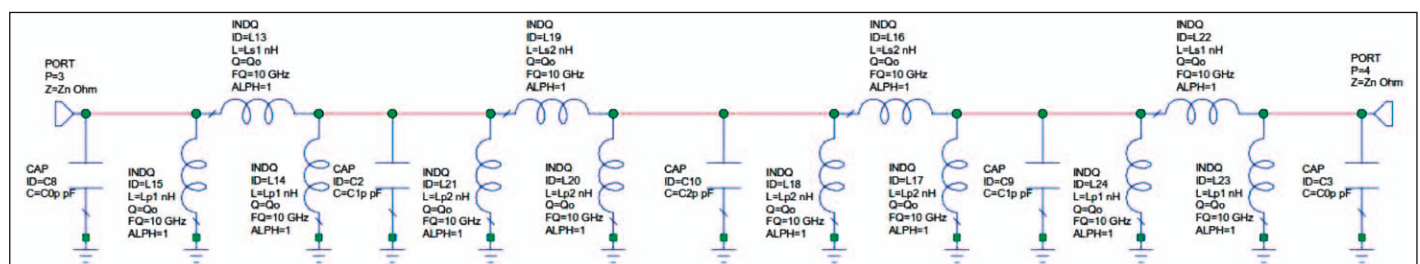*Figure 12: Optimized response*

Figure 13: Stand-off mounting hardware

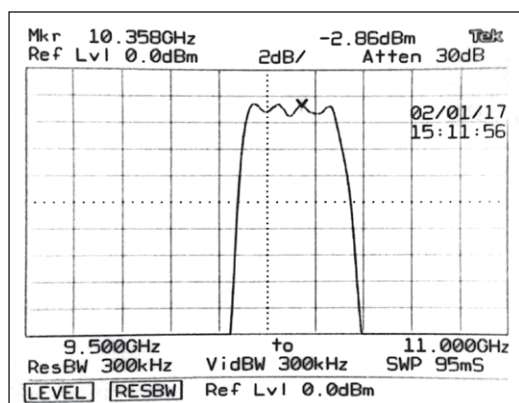*Figure 10: Three-resonator filter with tuned port transition.*

Figure 14: Measured response of three percent BW filter with 2 dB per step

Figure 15: Measured 40 dB BW response with 5 dB per step

and 5 resonator filters and create frequency of 10.4 GHz. These results follow: Tuning screw locations, passband loss, and reflection and -30 dB frequency points are displayed. A photo of the five resonator filter that was used in the 10 GHz test set is shown in Figure 13. Brass nuts soldered to the copper tube provide solid support and grounding for resonator tuning screws. Additional lock nuts are on each tuning screw.

Resonator screws are 4-40 and SMA transition tuning screws are 2-56 thread. Mounting stand-offs are 4-40 and are attached to the filter body with 'button head' screws.

The measured response of fabricated filters was very good. Figure 14 is a screen shot of swept response from one of the image reject filters in the 10 GHz test set shown in Figure 2. The measured passband is 10.18 GHz to 10.48 GHz with nominal insertion loss of 2.9 dB and ripple less than 1 dB peak. This 300 MHz BW was obtained from a three percent BW design. The 40 dB BW is 720 MHz, as shown in Figure 15.

Conclusion

This application note has described the design of two evanescent mode circular waveguide 10 GHz filters, which was undertaken to extend ham radio operation from typical operation below 30 MHz to 10 GHz. Microwave Office was used to design the filters for a range of bandwidths up to five percent. The ability of the software to optimize the design while simultaneously accounting for resonator spacing, insertion loss, and return loss, was key in the success of this design.

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N=6

%BW	BW(MHz)	l1(mil)	l2(mil)	l3(mil)	l4(mil)	C1(pF)	C2(pF)	C3(pF)	Loss(dB)	RetLoss(dB)	F30L(GHz)	F30H(GHz)
5	500	218	887	1020	1038	0.06517	0.05069	0.05062	6.32	20.21	9.98	10.81
3	300	221	950	1116	1142	0.06470	0.05062	0.05057	6.72	29.45	10.10	10.70
2	200	223	983	1172	1204	0.06441	0.05060	0.05060	6.89	39.20	10.15	10.65
1	100	235	1035	1245	1285	0.06302	0.05058	0.05054	7.68	48.89	10.20	10.59
0.5	50	279	1162	1385	1436	0.05919	0.05055	0.05053	11.83	48.44	10.28	10.51

N=5

%BW	BW(MHz)	l1(mil)	l2(mil)	l3(mil)	C1(pF)	C2(pF)	C3(pF)	Loss(dB)	RetLoss(dB)	F30L(GHz)	F30H(GHz)
5	500	171	793	961	0.07322	,05081	0.05088	3.10	27.41	9.77	10.99
3	300	218	940	1098	0.06514	0.05063	0.05058	4.63	28.00	10.10	10.77
2	200	256	1050	1209	0.06103	0.05058	0.05055	6.43	29.20	10.14	10.66
1	100	279	1151	1344	0.05922	0.05055	0.05054	8.64	30.03	10.23	10.57
0.5	50	300	1225	1452	0.05788	0.05054	0.05053	11.13	31.48	10.28	10.52

N=4

%BW	BW(MHz)	l1(mil)	l2(mil)	l3(mil)	C1(pF)	C2(pF)	Loss(dB)	RetLoss(dB)	F30L(GHz)	F30H(GHz)
5	500	194	841	966	0.06865	0.05075	2.66	21.25	9.66	11.07
3	300	219	938	1070	0.06491	0.05064	3.17	25.46	9.87	10.90
2	200	259	1051	1180	0.06077	0.05058	4.41	25.12	10.04	10.75
1	100	298	1173	1314	0.05797	0.05055	6.08	30.26	10.17	10.63
0.5	50	308	1215	1369	0.05743	0.05054	6.64	40.98	10.20	10.60

N=3

%BW	BW(MHz)	l1(mil)	l2(mil)	C1(pF)	C2(pF)	Loss(dB)	RetLoss(dB)	F30L(GHz)	F30H(GHz)
5	500	216	866	0.06500	0.05079	2.10	15.69	9.36	11.30
3	300	240	961	0.06259	0.05066	2.31	20.38	9.63	11.09
2	200	280	1072	0.05913	0.05059	3.13	20.51	9.87	10.89
1	100	320	1196	0.05680	0.05055	4.22	23.45	10.05	10.74
0.5	50	361	1310	0.05513	0.05054	5.80	25.83	10.16	10.63

N=2

%BW	BW(MHz)	l1(mil)	l2(mil)	C1(pF)	Loss(dB)	RetLoss(dB)	F30L(GHz)	F30H(GHz)
5	500	214	753	0.06576	1.22	13.26	7.33	12.26
3	300	253	880	0.06136	1.43	15.70	8.53	11.77
2	200	272	947	0.05974	1.54	18.72	8.90	11.56
1	100	319	1079	0.05686	2.10	21.34	9.44	11.21
0.5	50	352	1176	0.05545	2.69	23.01	9.69	11.02

Table 2: Set of solutions