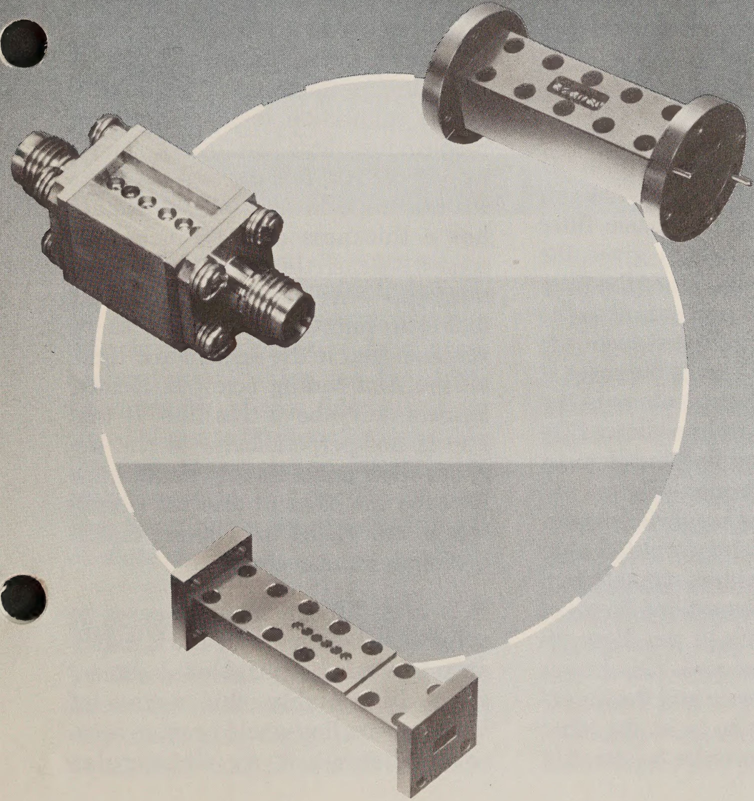
VOL. 17 NO. 6 NOVEMBER/DECEMBER 1990

Combline Filters for Microwave and Millimeter-wave Frequencies

Part 2



WATKINS-JOHNSON COMPANY
Tech-notes

Part 1 of this article[1] describes the basic design and realization of combline bandpass filters. A computerized design procedure has been developed that takes basic design information and generates a mechanical filter design that will give the performance required. Due to the closed-form nature of the mathematical expressions used, the whole design procedure can be executed on a personal computer in a matter of seconds. Due to the speed of the computer design procedure, designs can be iterated in very little time in order to optimize the realizability of the filter. Part 2 of this article gives some examples of these filter designs. Filters with and without end-element impedance transformation that use coaxial interfaces, and techniques used to build these filters for operation at millimeter-wave frequencies will be shown. Also, the design procedure will be extended to cover millimeter-wave filters that have waveguide interfaces.

Coaxial Interface

A filter interfaced to standard 50-ohm coaxial line is often required, and the design procedure given in Part 1 can be used to generate the basic filter design. While this design gives the mechanical configuration of the filter elements, the input and output to the filter are simply represented as impedances. (For the following purposes it will be assumed that 50 ohms is the required termination impedance.) By using either the modified end-element impedance transformer (Figure 9B, Part 1), the constant-bandwidth transformer or the broad-bandwidth design without any impedance transformation, the ground connections for all of the rectangular bars in the filter are located on the same end. This leaves the tuning capacitance and the input/output connections to be at the other end of the bars. In order to simplify assembly and tuning, it is desirable to have access to the tuning screws from the same side as the connection into and out of the filter. The connection to the first element in the filter (either an impedance transformer or resonator) needs to be made to the top of the element from a direction in line with the row of rectangular blocks, but perpendicular to them. This is simply accomplished by using a thin, flat tab of a coaxial connector as an air dielectric section of microstrip transmission line connected to the top of the first element perpendicular to the rectangular bar, as shown in Figure 1. This short section of microstrip line should attach to the top of the first element in the filter structure and have a thickness much smaller than the height of the resonator element. If this thickness and height are comparable, then the large relative dimension of the microstrip feed line tends to act as a tapped impedance transformer at the junction with the first element, thereby changing the terminal impedance at the connection to the filter. This problem manifests itself during tuning as poor VSWR. It is convenient to use the tab of a connector, which usually has a thickness of 0.005 inches, or some 0.002-inch thick gold-plated shim material. This short section of air dielectric microstrip line has to be perpendicular to the element and rigid, as the first tuning screw is located immediately above this line. If this line is not perpendicular to the element, then unacceptably small gaps between the element and the tuning screw can result in a short circuit during thermal cycling.

It is also difficult in most cases to achieve a 50-ohm air dielectric microstrip line using realizable tab dimensions. In actuality, this section of transmission line would be more accurately described as rectangular

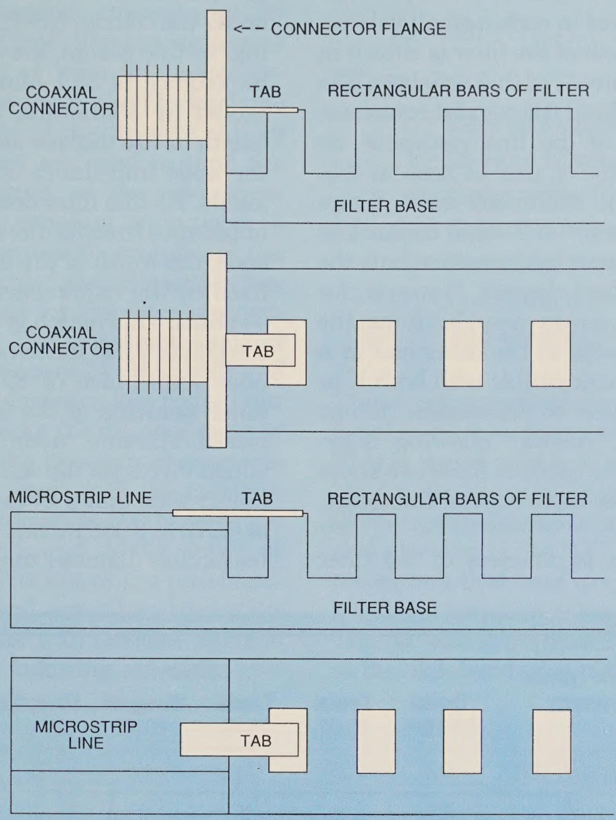


Figure 1. Coaxial and microstrip tab connection to a combline filter.

"trough-line," but air dielectric microstrip closely approximates the physical situation. The impedance of this microstrip line tends to be slightly higher than 50 ohms, but due to the short length of this line, it can be tuned closer to 50 ohms by adjusting the first tuning screw, which adds capacitance to this section of line, thus lowering its impedance closer to 50 ohms.

Microwave Filter Example

To illustrate the design of combline filters for microwave frequencies, a very challenging design for a 75% fractional bandwidth filter is given below. This filter is a non-redundant design, where all the elements in the filter contribute to the filter performance (i.e., there are no impedance-transformers). The basic mechanical construction of this filter is used for the majority of microwave and millimeter-wave combline filters that require coaxial interfaces and can be used for narrower bandwidth filters as well as those that require impedance-transforming elements. The filter is a 19th-order, equal-ripple Chebyshev design that gives a very selective response and is physically small. The size of this filter is 2.32 x 0.66 x 0.5 inches, excluding connectors, and rivals other technologies (i.e., suspended substrate and lumped elements) in the performance achieved in such a small volume. A photograph of the filter is shown in Part 1, Figure 1, of this article [1]. The transition from the coaxial connector to the top of the first resonator, as described above, can be seen in this photograph. Millimeter-wave filters using either K™ or 2.4-mm connectors utilize the same tab transition onto the top of the first element. However, for the millimeter-wave applications, the tab terminates at the other end in a pin that is compatible with both K or 2.4-mm types of replaceable flange connectors, thereby allowing interchangeability between these two styles of millimeter-wave coaxial connector.

The design parameters of the filter are shown in Table 1 in the column under the coaxial 5.8–12.75 GHz heading. In this design, the filter terminal impedance is 50.1 ohms and is obtained by shortening the resonator length to 40.0 degrees and by lowering the node impedance of the first elements. As this filter does not have an impedance transformer element at the ends, the width of the first element is fixed by the requirement to match to 50 ohms. This match is found to be at a width of 0.045 inches and a starting node impedance of 49.4 ohms. The other elements of the filter have different starting node impedances, which increases the width of the resonators such that they are in the range of 0.072 to 0.088 inches. As a result, a reasonable diameter tuning screw can

Filter Interface Type and Operating Frequency Range, GHz	Coaxial 5.8–12.75	Coaxial 33–37	Coaxial 25–40	Waveguide 41–47	Waveguide 33–37	Waveguide 26.5–40.0
Center Frequency, GHz	9.3	35	34	44	35	33.25
Bandwidth, GHz	6.9	4	15	6	4	13.5
Order	19	4	8	4	4	9
Passband Ripple, dB	0.01	0.01	0.01	0.01	0.01	0.001
Bandwidth Correction Factor, %	74	60	60	53	60	65
Filter Terminal Impedance, Ohms	50.1	50.0	50.0	95.0	92.0	98.3
Resonator Phase length, degrees	40.0	45.0	45.0	45.0	45.0	45.0
Approx. Tuning Capacitance, pF	0.35–0.62	0.077	0.070	0.053	0.077	.079–.096
Cavity Width, inches	0.360	0.100	0.180	0.090	0.100	0.110
Cavity Height, inches	0.146	0.042	0.043	0.034	0.042	0.045
Width of Resonators, in.	.045-.088	0.040	0.072	0.036	0.040	0.044
Height of Resonators, in.	0.126	0.032	0.033	0.028	0.032	0.035

Table 1. Equal-ripple chebyshev combline-filter parameters.

be used to achieve the relatively high resonator tuning capacitance of 0.35 pF to 0.62 pF. The tuning screws used require at least 0.1 inches between centers and are self locking due to the nature of the threads. All 19 tuning screws are used to tune the center frequency of the resonators. There are no coupling tuning screws or locking screws to complicate the design. A detailed plot of insertion loss and return loss for this filter is shown in Figure 2. The worst-case in-band return loss is about 14 dB (a VSWR of 1.5:1) and an insertion loss better than 1 dB (nominally 0.5 dB over most of the band). The rejection characteristics are very good, with the response being 40-dB down at 500 MHz below the passband and 40-dB down at about 400 MHz above the passband. The theoretical out-of-band rejection achieved with a Chebyshev filter is given by the following equation:

$$L = 10 . \mathrm{Log}_{10} \left[1 + r . \cosh^2 \left\{ n . \cosh^{-1} \left(|f_R - f_0| . 2/B \right) \right\} \right]$$

Where,

L = Loss at frequency f_R, dB
r = In-band ripple, dB
n = Order of the filter
f_R = Rejection frequency
f_0 = Passband center frequency
B = Passband bandwidth
$|f_R - f_0|$ = The absolute value of $f_R - f_0$

The rejection performance measured is not as good as theory would predict. This is due in part to a smaller passband than theoretically predicted, due to the approximations in the design and also due to the "Q" losses of the resonators that tend to round off the passband response at the band edges. The "Q" loss is particularly noticeable at the high-band edge of the passband

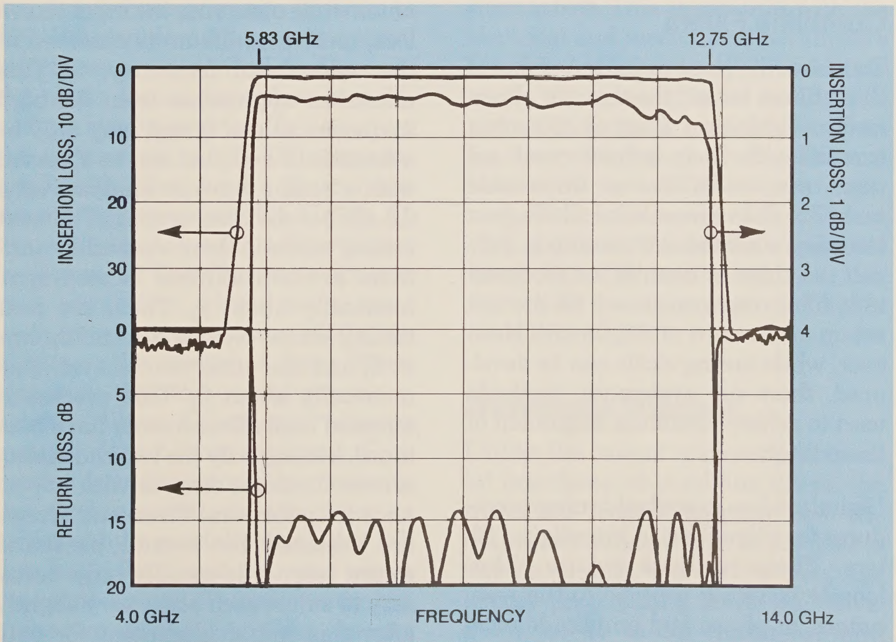


Figure 2. Insertion-loss and return-loss performance of a 5.83 GHz to 12.75 GHz combline filter.

in Figure 2, where the higher frequency makes these losses more pronounced. If the passband is defined as the maximum equal-ripple return-loss bandwidth (i.e., between the 14-dB return loss points), then the above equation would predict a rejection of 53 dB at 5.35 GHz and 41 dB at 13.18 GHz (assuming the revised passband frequencies are 5.83 GHz to 12.83 GHz, a 19th-order design with 0.01 dB ripple). This is almost exactly the rejection measured at the 13.18 GHz frequency on the high side of the band, but about 10 dB off on the low side of the band. As noted in Part 1 of this article, there are approximations in the design and the low-pass-to-bandpass mapping is not always accurate, but the performance of this filter is very good considering the broad fractional bandwidth (about 75%) that this filter exhibits.

Techniques for Tuning Combline Filters

To those who have never tuned one of these filters before, the thought of precisely aligning as many as 19 tuning screws, with their infinite combinations, may seem like an impossible task. For the narrow-band filters, just knowing where to start tuning is difficult, as there is usually no recognizable filter response shown on the test set-up at the start of alignment. However, while tuning skills can be developed, there are systematic methods used to achieve optimum alignment of these filters.

Dishal [2,3] has described tuning procedures for alignment of interdigital filters. These methods require either loosely coupling a probe to the resonators, or phase and amplitude measurement of the return loss. The method described below is used to align the combline filters described in this article and only requires the use of a scalar network analyzer. The following alignment method is simple and has proved very effective for filters of varying bandwidth. There are basically two systematic methods used to tune combline filters; the method used depends on the bandwidth of the filter.

For the narrow-band case, all tuning screws are backed out, away from the resonators, while for the broad-bandwidth case, the tuning screws, except for the two end ones which are disengaged (only if they are impedance-transforming elements), are screwed in all the way, thus shorting to the resonators. The procedure for the narrow-band case is to start turning the first resonator tuning screw (i.e., the first screw on a non-redundant design or the second screw on a design with impedance transformers) so as to reduce the gap at the top of the resonator while observing the input return loss, until a minimum is observed at the center-band frequency, f_0. This minimum approaches from the high frequency side of f_0 and may only be a fraction of a dB, but can be observed with a scalar network analyzer on a 0.5 dB per division setting. The next tuning screw is then screwed in until there are two minima located symmetrically about f_0. Then, the next tuning screw creates 3 minima, one at f_0 and the other two located symmetrically about f_0. This process is repeated until all resonators have been tuned, leaving only the two end tuning screws for those designs with impedance transformers. These end screws are adjusted for overall passband return loss. It is usually only necessary to adjust each screw very slightly after this initial alignment for optimum performance.

The tuning procedure is essentially

the same for the broad-bandwidth case, except that the tuning screws are all first shorted to the top of the resonators, except for the screws over the impedance-transforming elements. Tuning of the resonator is achieved by turning the tuning screw so that the gap at the top of the resonator is increased (i.e., decreasing the tuning capacitance). The return loss minimum approaches f_0 from the low-frequency side. Tuning of the other screws follows in a similar manner as described for the narrow-bandwidth case. In order to obtain the best possible response for broad-bandwidth filters, the fine-tuning step after initial alignment may need to be repeated a few times.

Examples of Millimeter-wave Coaxial Interfaced Filters

Both narrow-bandwidth (centered at 35 GHz) and broad-bandwidth (covering the frequency range 25 GHz to 40 GHz) combline filters with coaxial interfaces are described to illustrate the above tuning procedure at millimeter-wave frequencies. Table 1 gives the parameters of a 35-GHz narrow-bandwidth filter (under the column describing the coaxial 33–37 GHz filter), with a picture of this filter shown in Figure 3. The filter uses 2.4-mm coax connectors, which can be used up to frequencies in excess of 50 GHz. However, the design of these filters was such that they could accommodate the Wiltron K-Connector™ directly instead of the 2.4-mm connector. Connection from the coaxial connector to the first element in the filter is achieved by a specially machined tab that has a 0.012-inch diameter pin at the connector end. This allows the tab connection to the first filter element as described previously, while the pin connects to the replaceable coaxial connector at the other end. The performance of this 35-GHz filter is given in Figure 4, and shows an insertion loss of about 1 dB, with a passband return loss of about 20 dB. The 4th-order characteristic of this filter is clearly shown by the four maximum peaks in the return-loss characteristic of Figure 4.

The broad-bandwidth filter parameters are given in Table 1 under the column describing the coaxial 25–40 GHz filter. The more challenging task of designing for a broad bandwidth with this style of filter is shown in this example. The response of the filter is given in Figure 5, which shows that the measured passband is from 25 GHz to 40 GHz. This measured passband is wider than expected, indicating the need to artificially reduce the bandwidth prior to running through the design process by more than originally anticipated. The actual bandwidth correction factor for this filter is 60%. This shows that with the electrical and mechanical parameters for this high-frequency broad-bandwidth filter, a different bandwidth correction factor is required than that for lower-frequency filters, where a bandwidth factor of 80% is more common. The performance of this filter is very good over the full Ka waveguide band (26.5 GHz to 40 GHz), with less than 1-dB insertion loss and better than 1.7:1 VSWR.

Waveguide Interface

Unlike the case of microstrip or coaxial interfaces on combline filters, the transition from waveguide into the coupled rectangular bars of the filter is not as simple. For waveguide, the energy propagating down the waveguide needs to be channeled and directed into the coaxial filter structure. One method that can be used would be to make the transition into

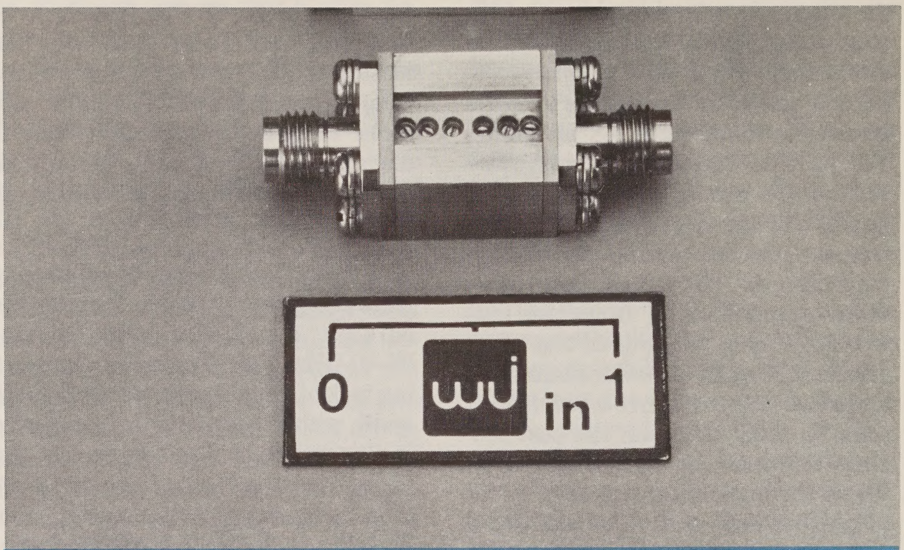


Figure 3. A 4th-order, 35-GHz combline filter with 2.4-mm coaxial connectors.

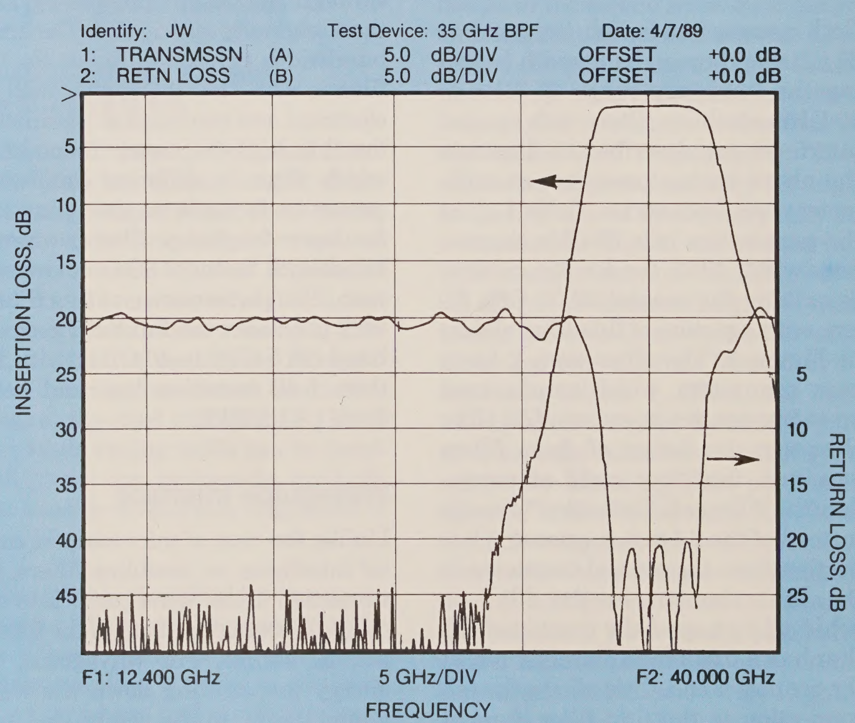


Figure 4. Performance of 35-GHz combline filter with 2.4-mm coaxial connectors.

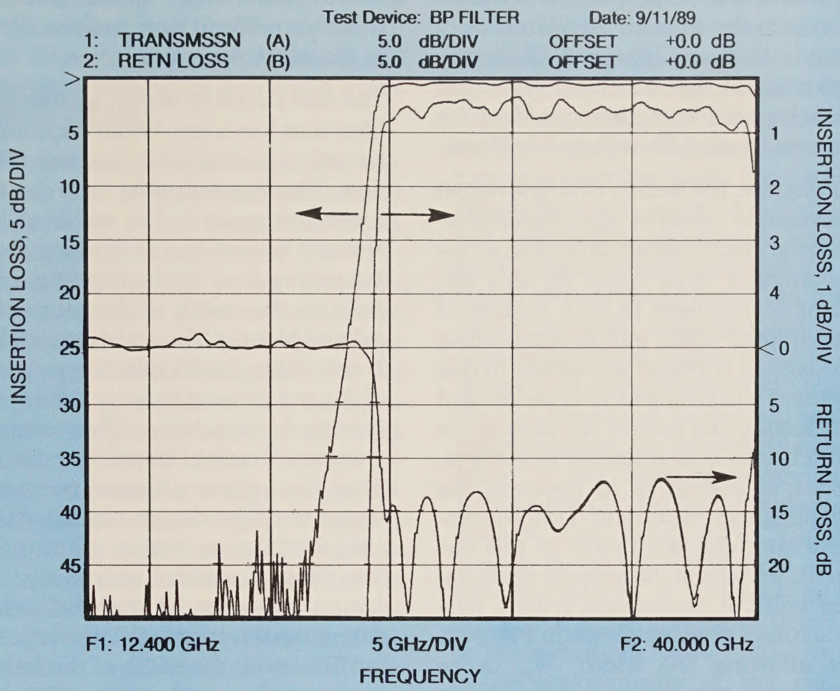


Figure 5. Performance of combline filter covering 25 GHz to 40 GHz with coaxial connectors.

coax by the use of an E-field transition probe and then from the coaxial line into the filter, as described above in the section on coaxial interfaces. Alternatively, by reverting back to the original unmodified configuration for the end transformers (with the ground connection on the opposite side from the resonators), one can conceive of a filter structure that has the E-field probes feeding directly into the first element. Other methods exist that use an intermediate length of coaxial transmission line between the waveguide and the filter elements, but they are all mechanically complex and will require very tight machining and alignment tolerances in order to perform acceptably.

What is needed is a mechanically simple structure for the transition from waveguide into the filter elements that does not require absolutely precise alignment accuracy and is easy to assemble. One solution is to utilize single-ridged waveguide [4] to couple into the first filter element. Quarter-wavelength stepped-impedance single-ridged waveguide transformer sections are used to transform from the terminal impedance of the filter (which is also the impedance of the last section of ridged waveguide) to the waveguide impedance. The subtlety of this approach is in the selection of the terminal impedance of the filter and how the ridged waveguide couples into the first resonator element of the filter.

The first problem to solve is the

method of ending the ridged section of waveguide and making the transition into the filter in a mechanically simple fashion. Figure 6 illustrates one concept that has been developed to solve this transition problem for narrow-bandwidth waveguide filters.

In practice, the section of single-ridged waveguide absorbs the impedance-transforming element at the end of the rectangular coax filter. As this element is supposed to be a section of rectangular coax, simply connecting the section of ridged waveguide to this element will change the impedance of the section and destroy the impedance match that it is intended to achieve. What is needed is to preserve the coupling capacitance and impedance-matching characteristics of this element, while simultaneously reducing the width of the element to zero. This is accomplished, as shown in Figure 6, by allowing the width, W_e, to be reduced to zero while absorbing this element into the ridged waveguide, thus preserving the coupling into the first resonator of the filter. While this procedure is approximate in nature, narrow-bandwidth filters designed this way perform very well, as shown in the examples given below.

The first step is to design a basic coax filter that has a zero-width impedance-transformer element at the ends of the filter. This can indeed be done due to a correction made to the width of this element caused by the fringing fields (assuming that all the fringing fields exist as the width of the element is reduced to zero). The situation required is one where the fringing capacitance of the end element cancels out its own even-mode capacitance. This occurs at a unique terminal impedance for the filter for a given physical geometry. Using the filter design already developed for the coax version of the combline filter described previously, the design program is iterated while changing the terminal impedance of the filter until the width of the impedance-transformer element vanishes to zero. For the filters described below, this occurs at a terminal impedance of about 100 ohms. The dimensions of

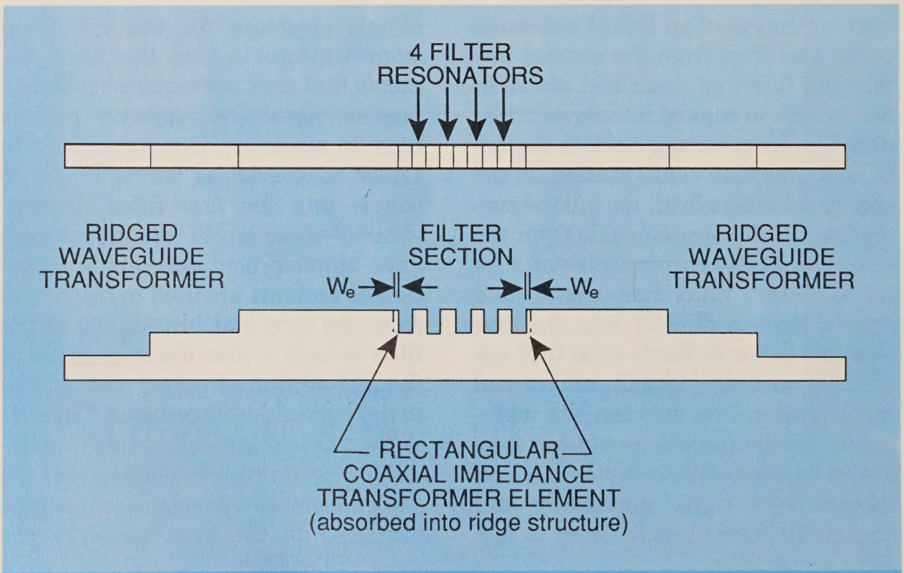


Figure 6. Ridged waveguide quarter-wave transformer transition to coaxial combline filter.

the filter resonators and thickness of the ridge in the ridged waveguide has to be selected such that the end face of the ridge has the same thickness and height as that of the zero-width transformer element.

The next stage in the design is to construct the single-ridged impedance transformer to transform from the filter terminal impedance to the wave impedance of the waveguide. (The wave impedance is defined as the ratio of the transverse electric to transverse magnetic field in the guide.) For the examples given below, this amounts to transforming from approximately 100 ohms to 500 ohms. This can be done by designing a stepped quarter-wavelength single-ridged waveguide impedance transformer. The design procedure that follows is very approximate due to the inhomogeneous nature of the waveguide sections and no claims to its accuracy are made. However, it has proved acceptable for the narrow-bandwidth filters described below. A different, and possibly more accurate, treatment of this problem is covered in the design of the broad-bandwidth waveguide filter.

For any given bandwidth, there exists an optimum set of step impedances, for a given number of steps, that yields a Chebyshev-type equal-ripple match that minimizes the VSWR over the required frequency range selected. This procedure is described by Cohn [5] and readily lends itself to be solved on the computer. The design assumes that if these impedances are created in the ridged waveguide quarter-wavelength sections, then the required transformation will be achieved. There is uncertainty as to which definition of impedance in the ridged waveguide to use and this will be discussed later.

In order to keep the length and loss of the single-ridged transformer sections to a minimum, the number of sections in the transformer should be minimized. While multiple sections over the full waveguide bandwidth can be constructed, the added complexity is unnecessary for narrow-bandwidth designs (a full waveguide-band transformer is described later). For the filter examples given below, a 3-step (i.e., two intermediate impedances between the waveguide impedance and the terminal impedance of the filter) transformer covering approximately a 1.2 bandwidth ratio will give excellent return loss performance and achieve the transformation in minimal length. For waveguide, the bandwidth ratio is given by the lowest frequency guide wavelength divided by the highest frequency guide wavelength. Unfortunately, the guide wavelength at the two frequencies is different in the sections of ridged waveguide than that in the rectangular waveguide. As the guide wavelengths of the ridged-waveguide sections are not known at the start of the design, the ratio of guide wavelengths in the rectangular waveguide can be used to determine the bandwidth ratio.

Examples of Millimeter-wave Waveguide Interface Filters

To illustrate the above design procedure, a rectangular coax combline bandpass filter with a center frequency of 44 GHz and a B-Band (WR 22, 33 GHz to 50 GHz) rectangular waveguide interface is described below. The required electrical specification for the filter is entered into the design equations on the computer with an initial guess at the terminal impedance for the filter. The physical realization of the filter is in a 0.090-inch cavity using resonators that are 0.036-inch thick (i.e., a resonator thickness to cavity width ratio of 0.4). This small cavity size is chosen to increase the cutoff frequency of any spurious waveguide modes, thus further assuring that the stop-band performance of the filter is maintained. The program is iterated, changing the terminal impedance, until the width of the end-transformer section decreases to zero. This has been found to occur with a terminal impedance of 95.0 ohms. Simultaneously, during these iterations, the node impedances of the filter are adjusted to make all the resonators have the same width. The complete specification for the filter is given in Table 1, under the waveguide 41–47 GHz filter column, and the physical realization of this design is shown in Figure 7.

Knowledge of the required terminal impedance for the filter allows the impedance transformer to be designed. First, the waveguide impedance needs to be calculated. For this example, the WR 22 (0.224-inch x 0.112-inch) rectangular waveguide was tapered to WR 19 (0.188-inch x 0.094-inch) in order to raise the cut-off frequency of the waveguide and to reduce the discontinuity susceptance in the transition region between the ridged waveguide and the coax filter. Using the following relationships for TE_{10} modes in air dielectric waveguide:

$$\frac{1}{L_g^2} = \frac{1}{L_0^2} - \frac{1}{L_c^2}$$

and,

$$Z_0 = 377.L_g/L_0$$

Where,

- L_g = Guide wavelength
- L_0 = Free-space wavelength
- L_c = Waveguide cut-off wavelength
- Z_0 = Waveguide characteristic impedance

The definition of waveguide impedance as the ratio of transverse electric and magnetic fields (wave impedance) appears to be a logical choice to use in matching to the terminal impedance of the filter. However, using this definition makes the determination of the ridged waveguide dimensions rather complex. A different definition of waveguide impedance is based on the power carried by the waveguide (power impedance) and applies to infinite frequency. In order to determine the impedance at a specified frequency, the impedance must be multiplied by L_g/L_0. This impedance, Z_{inf}, is defined as:

$$Z_{inf} = V^2 / (2.P)$$

Where,

- V = Peak voltage across the center of the ridge
- P = Average power carried by the guide

To illustrate the above, the narrow-bandwidth filters are designed using the wave impedance definition, while the broad-bandwidth filter uses the power definition of impedance.

At a frequency of 44 GHz, the guide wavelength and the wave characteristic impedance are found to be 0.383-inch and 538.0 ohms, respectively. Cohn's procedure [5] is then used to design a 3-step (two intermediate impedances) quarter-wavelength stepped impedance transformer to match 538.0 ohms to 95.0 ohms over the frequency range of 40 GHz to 48 GHz. This procedure gives 347.2 ohms and 147.2 ohms as the two intermediate impedances, with a better than 1.05:1 VSWR.

We are now in a position to synthesize three sections of single-ridged waveguide with impedances of 347.2 ohms, 147.2 ohms and 95.0 ohms. A sketch of the stepped impedance transformer is shown in Figure 8 and all that remains is to determine the step heights

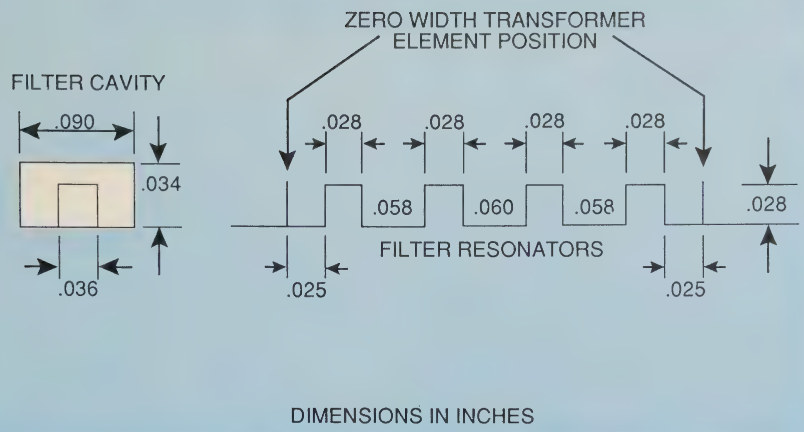


Figure 7. A 44-GHz combline filter design (excluding waveguide interfaces).

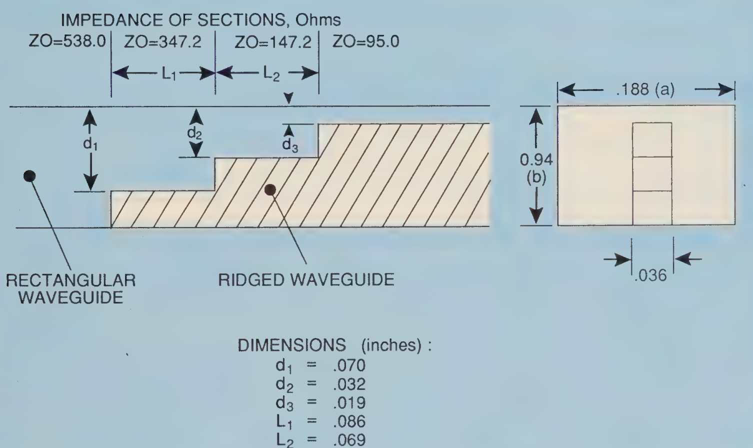


Figure 8. Stepped impedance single-ridged waveguide transformer design for the waveguide interfaces on 44-GHz combline fitler.

(d_1, d_2 and d_3) and the lengths (L_1 and L_2) that make the intermediate steps appear electrically 90° long.

The design procedure used for the single-ridged waveguide is based on that described by Hopfer [4]. Using the 95.0-ohm section of ridged waveguide as an example, the following parameters are used to evaluate the unknown parameter, d_3.

$$a/b = 0.5$$

$$s/a = 0.1915$$

$$Y_o = 1/Z_0 = 1/95.0 = 0.01053$$

Where,

a = Wide waveguide dimension

b = Narrow waveguide dimension

s = Thickness of the ridge

First, an estimate of d/b is made, where d is the distance between the top of the ridge and the waveguide (d_3 for this section). In this case, 0.1 is used as the initial guess for d/b and, using Figure 9, the value for the normalized cut-off wavelength, L_c/a, is determined. Although this set of curves is valid for b/a = 0.45, they can be used if a correction to L_c/a is made for the actual value of b/a = 0.5. The parameter F_o is read off from Figure 10 and the correction made as follows:

$$L_c/a \text{ (corrected)} = L_c/a + F_o.(b/a - 0.45)$$

Using $L_c/a = 5.16$ and $F_o = 1.85$, the corrected value for L_c/a is found to be 5.253. The guide wavelength, L_g, is then determined using the relationship between L_g, L_0 and L_c shown above. The design procedure [4] uses the power definition of impedance, and it is assumed that the wave definition of impedance can be used in place of the power definition. Knowing L_g allows the admittance at infinite frequency to be determined as follows:

$$Y_{inf} = Y_o.L_g/L_0$$

Using this value for Y_{inf} and $s/a = 0.1915$, the value $d/b = 0.22$ can be read from Figure 11. It is assumed that the error introduced by using this design curve (based on a power definition of impedance) is minimal. The value determined for d/b represents a better guess at the final value, but is

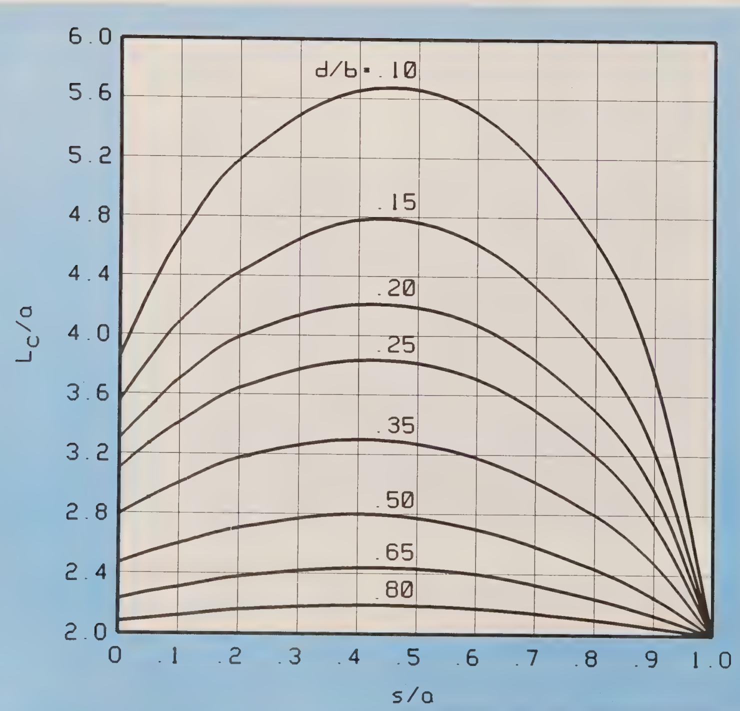


Figure 9. Determination of normalized cutoff wavelength (L_c/a).

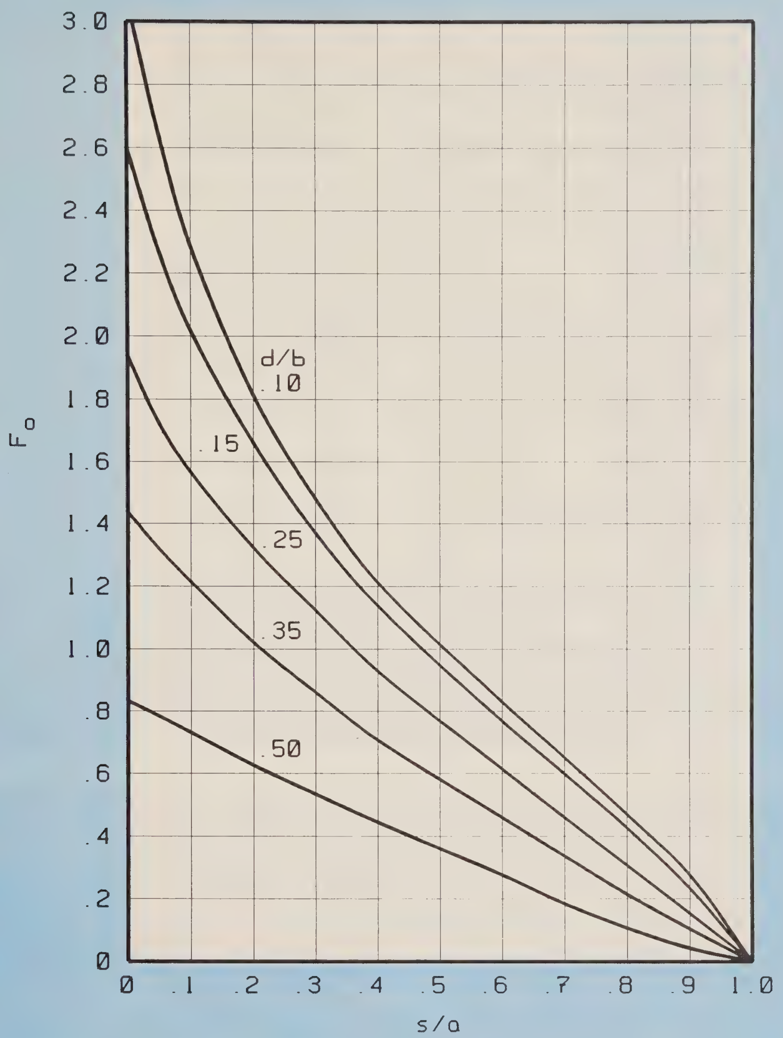


Figure 10. Waveguide aspect-ratio correction factor (F_o).

in error due to the slightly incorrect value of L_g that comes from using the initial value of d/b = 0.1. Using the new value for d/b = 0.22 and iterating the above procedure allows the design to converge on the final value of d/b = 0.20. In practice, the number of iterations needed is only 2 or 3, and in the above example it takes only one more iteration to arrive at the final value of d/b = 0.20. The above procedure is then repeated for the other sections of ridged waveguide.

Having determined the heights of the ridged sections, all that remains is to determine the lengths of the two quarter-wavelength intermediate sec-

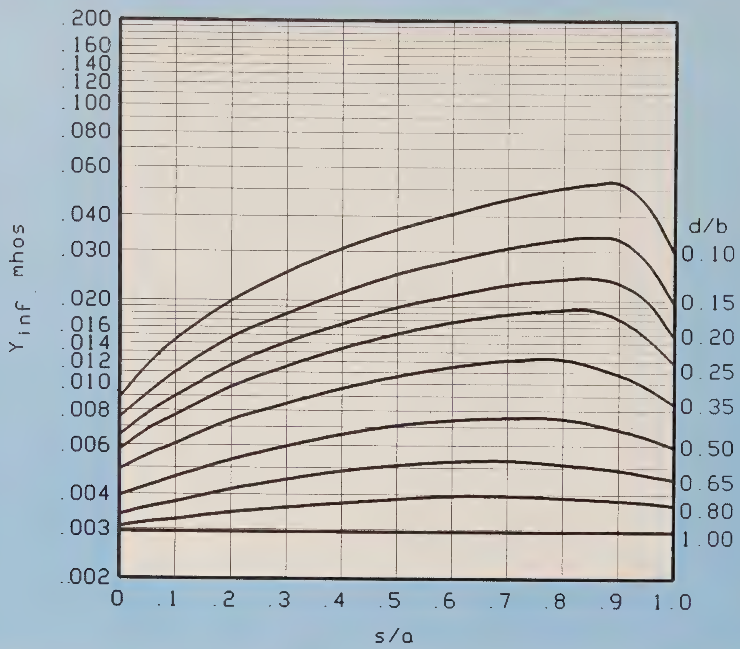


Figure 11. Waveguide admittance at infinite frequency (Y_{inf}).

tions of ridged waveguide (L_1 and L_2 in Figure 8). The initial length of these sections is calculated from the following expression:

$$\text{Length} = \frac{L_{g40}.L_{g48}}{2.(L_{g40} + L_{g48})}$$

Where,

L_{g40} = Guide wavelength at 40 GHz

L_{g48} = Guide wavelength at 48 GHz

This length is then corrected to account for the fringing fields at the step in the ridge. The procedure used is covered in Section 6.08 of reference [6]. The value for the shunt susceptance at this junction is taken from Figure 5.07–11 of reference [6]. The equivalent circuit used to represent this junction is that for an asymmetric junction. As the majority of the fields are concentrated above the ridge, the values for the shunt susceptance obtained are assumed to be the same as if the ridge extended across the whole width of the waveguide, thus allowing the use of the above information directly. The value of this shunt susceptance and the use of the procedure given in reference [6] is highly questionable due to the inhomogeneous sections of waveguide, which have different guide wavelengths that are physically different compared to a step that extends across the whole width of the waveguide. However, as long as the steps are relatively small, then this correction is minor and even though it is not entirely accurate, it gives acceptable results.

The final design of this 44-GHz filter is given in Figure 8 and Table 1 under the heading for the waveguide 41–47 GHz filter. This 44-GHz filter is shown in Figure 12, and its performance in Figure 13. The insertion loss is less than 1 dB, with greater than 20-dB return loss across the passband of the filter.

A similar filter was designed for a center frequency of 35 GHz that uses the same central combline section as the filter shown in Figure 4, only with waveguide interfaces instead of the coax interfaces. The design parameters for this filter are given in Table 1 under the heading for the waveguide 33–37 GHz filter. Figure 14 shows the split-block construction of this filter, with the performance shown in Figure 15.

Broad-Bandwidth Millimeter-wave Waveguide Filter Example

The design procedure described above produces acceptable results when applied to narrow-bandwidth designs. However, when designing for broad bandwidths, the errors introduced due to approximations and discontinuities cannot be easily tuned out as is the case with narrow-bandwidth designs. The following example is for a combline filter with a pass bandwidth that covers the whole Ka waveguide band (26.5 to 40 GHz).

As stated earlier in this article, broad-bandwidth combline filters tend to have a low filter-terminal impedance. Thus, it is not only unnecessary to try to absorb a zero-width impedance transformer element into the ridged waveguide, but it is also undesirable, due to the approximations and assumptions in the design approach. By designing the broad-bandwidth filter without impedance-transformer elements, the connection into the filter can be made by simply connecting a metal tab from the top of the ridged waveguide to the top of the first filter element in a manner similar to the

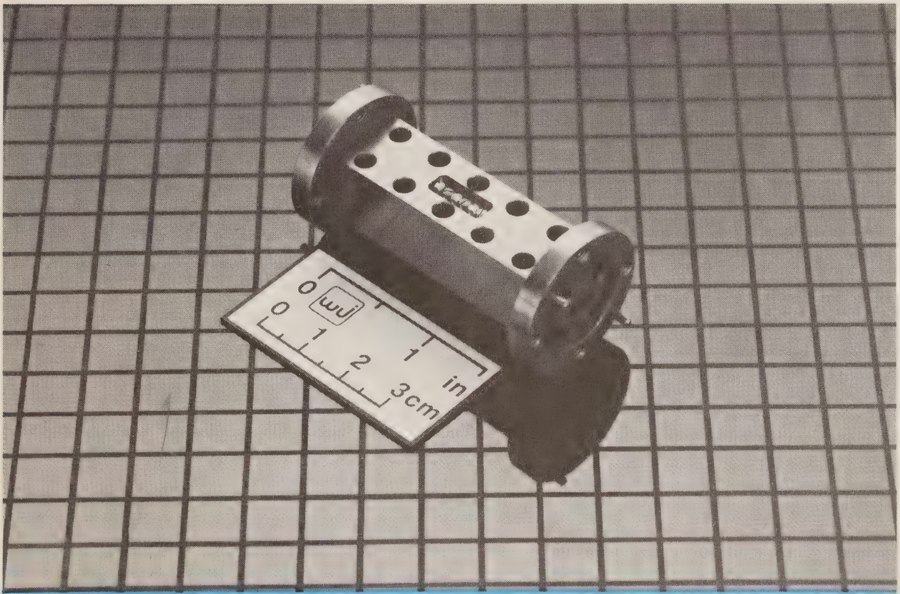


Figure 12. A 4th-order, 44-GHz combline filter with waveguide interfaces.

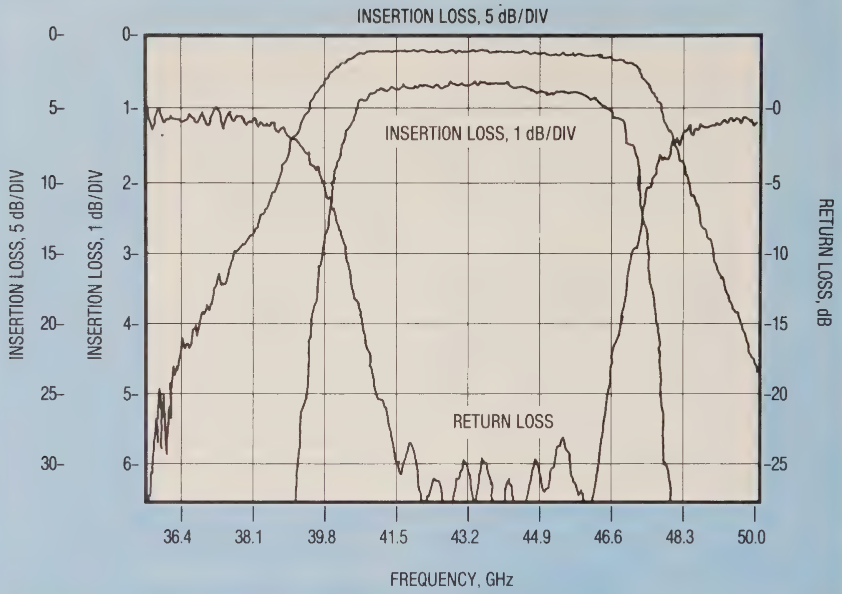


Figure 13. Performance of 44-GHz combline filter with waveguide interfaces.

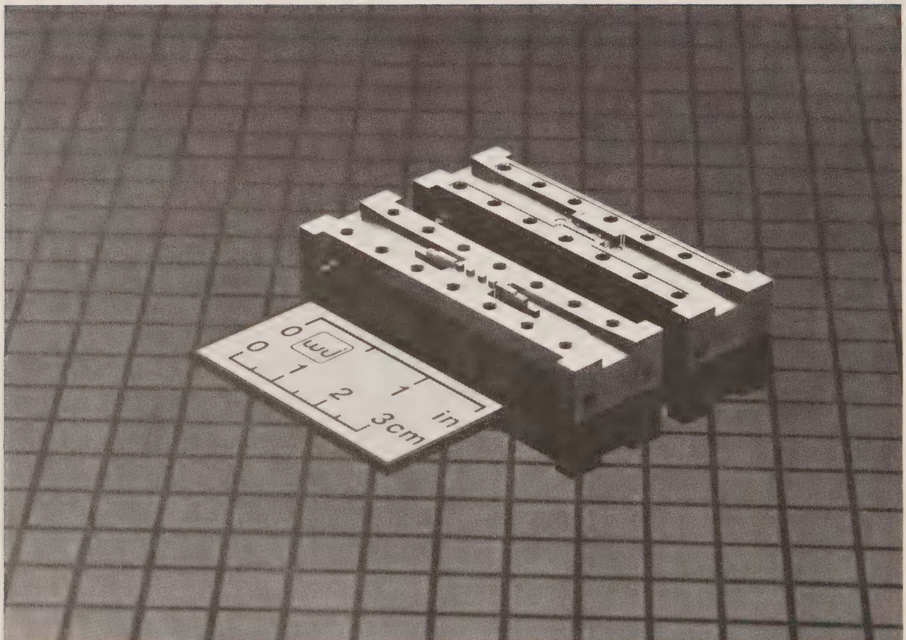


Figure 14. A 4th-order, 35-GHz combline filter with waveguide interfaces showing split-block construction and ridged waveguide transition.

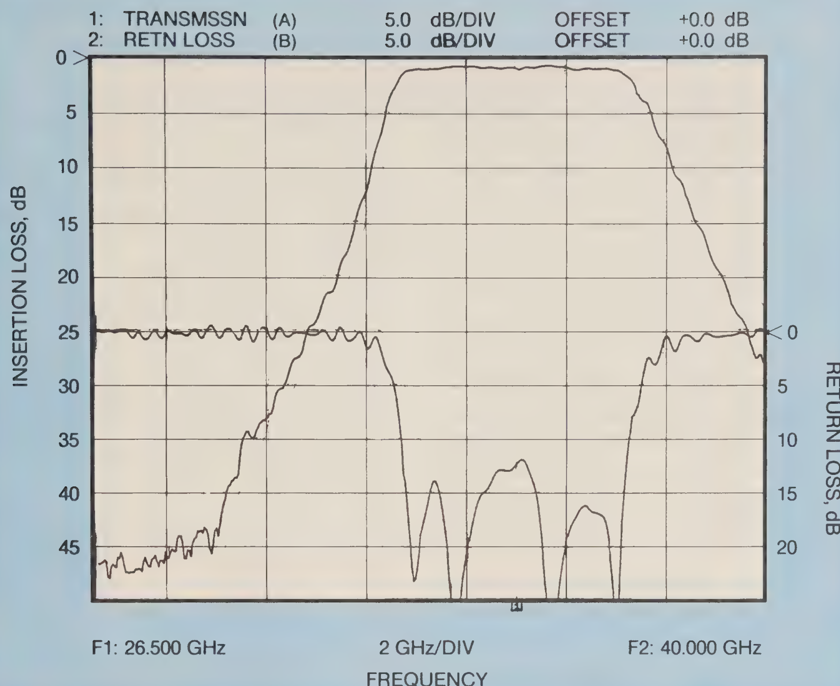


Figure 15. Performance of 35-GHz combline filter with waveguide interfaces.

microstrip tab transition shown in Figure 1. By keeping the length of this tab short, the resulting discontinuity can be minimized and acceptable performance achieved.

The filter is designed to cover 26.5 GHz to 40 GHz as a 9th-order, 0.001-dB equal-ripple Chebyshev type, with 45-degree resonators using a 65% bandwidth correction factor. The starting node impedances are adjusted so that all of the resonators have a width of 0.030 inches. The combination of all of these inputs to the design, results in a required terminal impedance of 98.3 ohms. A summary of the parameters of this filter are given in Table 1 under the heading for the waveguide 26.5–40.0 GHz filter.

The problem now is to design a quarter-wavelength single-ridged waveguide transformer that can provide the impedance transformation from the WR28 rectangular waveguide into the 98.3-ohm terminal impedance of the filter. With traditional waveguide filters, the absolute value of waveguide impedance is not important, since all filter impedances can be normalized to the characteristic impedance of the waveguide no matter which definition of impedance is used. This is not the case for combline filters with waveguide interfaces discussed in this article. An absolute value of the terminal impedance at the combline filter is known and this needs to be transformed to the waveguide impedance.

There are several ways to approach this problem, all with varying degrees of accuracy. For this filter example, the use of the power definition of impedance in the waveguide sections is adopted. From Figure 11, it can be seen that d/b=1 is the case for waveguide with no ridge present, and an infinite frequency impedance of 339 ohms results (i.e., an infinite frequency admittance of 0.00295 mhos). This is equivalent to the power definition of impedance for free space (the wave definition of impedance of free space is 377 ohms). If it is assumed that the power definition of impedance at the input to the filter is 98.3 ohms, then a simple iteration of the ridged-waveguide design procedure, as described for the narrow-bandwidth case, gives a required terminal impedance at infinite frequency of 93.1 ohms.

The design procedure using the power definition of impedance at infinite frequency, as given in [4], can now be used to match the rectangular waveguide impedance of 339 ohms to 93.1 ohms over the 26.5 to 40 GHz frequency range. While this does produce acceptable results, it would be more accurate to transform the terminal impedance at the filter (assumed to be equivalent to a wave impedance) to the power impedance before determining the guide wavelength, and then calculate the power impedance at infinite frequency. We already know the relationship between the wave and power definitions of impedance at infinite frequency (i.e., free space), and this may be used to scale the terminal impedance of the filter prior to determining the impedances of the quarter-wavelength transformer sections.

To convert from wave impedance to power impedance, multiplying the wave impedance by the ratio of the power impedance and wave impedance at infinite frequency (i.e., by 339/377) may produce more accurate results. The effect and applicability of this impedance transformation has yet to be determined, and the design presented below does not include this step.

The ridged waveguide transformer uses 4 steps and is designed to cover the frequency range from 25 GHz to 41 GHz, in order to give some design margin. This gives a fractional bandwidth of 2.62, as determined by the ratio of guide wavelengths, and results in a theoretical maximum return loss of 24.5 dB over the band of interest. The values of infinite frequency impedance from the rectangular waveguide to the terminal impedance of the filter are, therefore:

Waveguide Section		Impedance
Rectangular	=	339.0 ohms
1st ridged	=	267.9 ohms
2nd ridged	=	177.7 ohms
3rd ridged	=	117.8 ohms
4th ridged	=	93.1 ohms

For the given value of s/a=0.157, the value for d/b can be derived from Figure 11 by using the above values of infinite frequency impedance (note that $Y_{inf} = 1/Z_{inf}$). Given d/b, Figures 9 and 10 can be used to determine L_c and L_g, and the lengths of the intermediate sections of ridged waveguide are determined in the same way as described in the section on narrow-bandwidth designs.

Figure 16 shows the insertion loss and return loss achieved with this filter over the frequency range 20 GHz to 40 GHz. As the Ka-band waveguide approaches its cut-off frequency below 22 GHz, the response shown from 20 GHz to 22 GHz is not valid because an accurate calibration of the test equipment cannot be made. The filter performs very well over the 26.5 to 40 GHz frequency range with a worst-case VSWR of about 1.5:1.

This filter is designed with 9 resona-

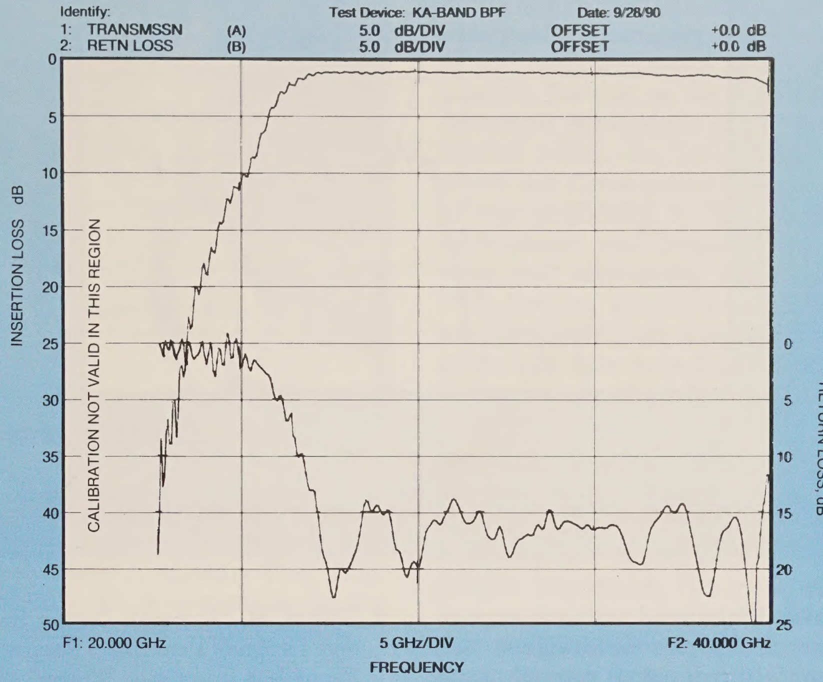


Figure 16. Insertion loss and return loss of waveguide Ka-band filter.

tors in order to assure that the rejection at 44 GHz is greater than 30 dB below the passband response. The test equipment used to measure the response shown in Figure 16 was limited to an upper frequency of 40 GHz. To test the high frequency rejection response, a B-band (33 GHz to 50 GHz) waveguide test set-up is used that employs a WR22 to WR28 tapered waveguide transition to connect to the filter. The results of measuring the insertion loss and return loss over the 33 GHz to 50 GHz frequency range are given in Figure 17, and show that the rejection of the filter at 44 GHz greatly exceeds the requirement to be 30 dB below the passband response.

Conclusion

The design and manufacture of microwave and millimeter-wave combline filters with coaxial interfaces has been shown. Examples of coaxial filters for operation up to 40 GHz have been given. It appears that the frequency limitation of these coaxial filters is determined by the mechanical accuracy required. Coaxial interface combline filters for operation in excess of 40 GHz can be manufactured, but the limited use of coaxial equipment above 40 GHz shows that there is only a limited need as yet.

Millimeter-wave combline filters with waveguide interfaces have also been

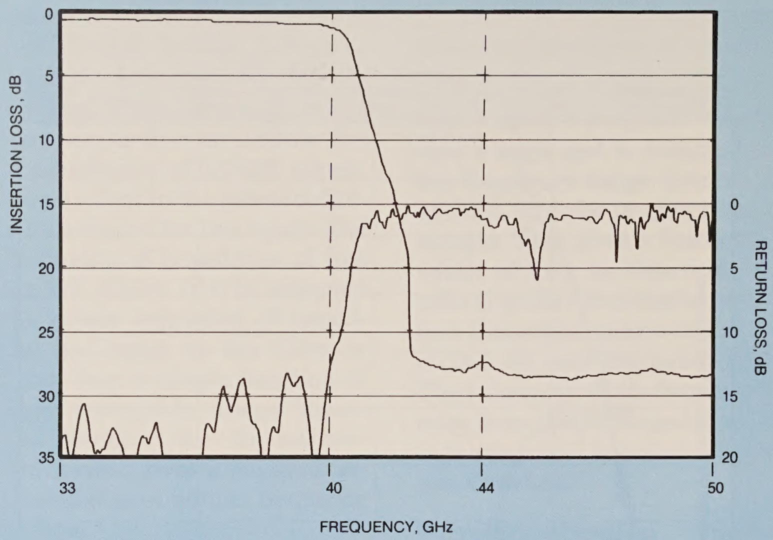


Figure 17. Performance of Ka-band waveguide filter over the 33 GHz to 50 GHz frequency range.

shown. Both broad and narrow passband filters can be designed and manufactured using combline-style technology. This type of filter has proved to be a high-performance design that is relatively easy to manufacture. The lack of spurious responses close to the required passband make these filters an attractive alternative to other styles. The broad-bandwidth waveguide design can cover a full waveguide bandwidth and provide excellent rejection characteristics and low insertion loss.

While combline-style bandpass filters may not be ideal for every filter requirement, they are very versatile and have proved to be adaptable to most filter requirements.

References

1. Hey-Shipton, G.L. "Combline Filters for Microwave and Millimeter-wave Frequencies: Part 1," Watkins-Johnson Co. Tech-Notes, Vol. 17, No. 5, September/October 1990.
2. Dishal, M. "A Simple Design Procedure for Small Percentage Bandwidth Round-Rod Interdigital Filters," *IEEE Trans. on Microwave Theory and Techniques,* Vol. MTT-13, September 1965, pp. 696–698.
3. Dishal, Milton. "Alignment and Adjustment of Synchronously Tuned Multiple-Resonant-Circuit Filters," *Proc. IRE,* Vol. 39, June 1952, pp. 154-164.
4. Hopfer, S. "The Design of Ridged Waveguides," *IRE Trans. on Microwave Theory and Techniques,* Vol. MTT-3, October 1955, pp. 20–29.
5. Cohn, S.B. "Optimum Design of Stepped Transmission Line Transformers," *IRE Trans. on Microwave Theory and Techniques,* Vol. MTT-3, April 1955, pp. 16–21.
6. Matthaei, G.L., Young, L. and E.M.T. Jones, "Microwave Filters, Impedance-Matching Networks, and Coupling Structures," Artech House Inc., 1980.

Author

Gregory L. Hey-Shipton

Dr. Hey-Shipton is a Department Manager, Subsystems Division, Watkins-Johnson Company, Palo Alto, California. He is responsible for the development of the complex subassemblies being built in Palo Alto, which cover the frequency range from dc to 110 GHz and include the products from the Spacekom operation formerly located in Santa Barbara, California.

Prior to his assignment in Palo Alto, Dr. Hey-Shipton was the Engineering Manager at the Watkins-Johnson Company Spacekom Facility. At Spacekom, he was responsible for the research, development, design and management of devices and products mainly for operation at millimeter-wave frequencies.

Dr. Hey-Shipton has been involved in the design of many of the subsystems products currently in production as well as in the design of new components. Serving as the head of the Advanced Subsystems Development Section within the Subsystems Research and Development Department, he was responsible for the research, development, and design of new devices and subsystems using GaAs FET devices, amplifiers, doublers, microstrip, stripline, suspended substrate, coplanar waveguide, mixers, filters, switches, couplers, and associated circuitry.

As a member of the technical staff at Watkins-Johnson Limited in Windsor, England, he designed amplifiers, multicouplers, control circuitry, and receiver accessories. His main work was as project leader and chief designer of an airborne ECM system, which he took from the initial feasibility stage through design and into flight testing.

Dr. Hey-Shipton has written many technical articles. He received a B.Sc. from the University of Manchester, England, and a Ph.D from the University of Leeds, England.

WATKINS-JOHNSON COMPANY
3333 HILLVIEW AVENUE
PALO ALTO, CA 94304-1204
(415) 493-4141

ADDRESS CORRECTION REQUESTED

BULK RATE
U.S. POSTAGE
PAID
PERMIT
#1
SAN JOSE
CALIFORNIA

WATKINS-JOHNSON

Facility Locations

United States

CALIFORNIA

Watkins-Johnson
3333 Hillview Avenue
Palo Alto, 94304-1204
Telephone: (415) 493-4141

Watkins-Johnson
2525 North First Street
San Jose, 95131-1097
Telephone: (408) 435-1400

Watkins-Johnson
440 Kings Village Road
Scotts Valley, 95066-4081
Telephone: (408) 438-2100

MARYLAND

Watkins-Johnson
700 Quince Orchard Road
Gaithersburg, 20878-1794
Telephone: (301) 948-7550

Watkins-Johnson
8530 Corridor Road
Savage, 20763
Telephone: (301) 497-3900

NORTH CAROLINA

Watkins-Johnson
1000 West Powell Road
Fuquay-Varina, 27526-9399
Telephone: (919) 552-6161

International

UNITED KINGDOM

Watkins-Johnson
Dedworth Road
Oakley Green
Windsor, Berkshire SL4 4LH
Telephone: 753 869241
Telex: 847 578
Cable: WJUKW-WINDSOR
Telefax: 753 841534

ITALY

Watkins-Johnson S.p.A.
Piazza G. Marconi 25
00144 Roma-EUR
Telephone: 6 592 4554
6 591 2515
Telex: 612 278
Cable: WJ ROM I

GERMANY, FEDERAL REPUBLIC OF

Watkins-Johnson
Boschstr. 10
8039 Puchheim
Telephone: 89 802087/88
Telex: 529 401
Cable: WJDBM-MUENCHEN
Telefax: 89 803044

The Watkins-Johnson *Tech-notes* is a bi-monthly periodical circulated to educational institutions, engineers, managers of companies or government agencies, and technicians. Individuals may receive issues of *Tech-notes* by sending their subscription request on company letterhead, stating position and nature of business to the Editor, *Tech-notes,* at Watkins-Johnson's Palo Alto, California address. Permission to reprint articles may also be obtained by writing the Editor.

Printed in U.S.A.