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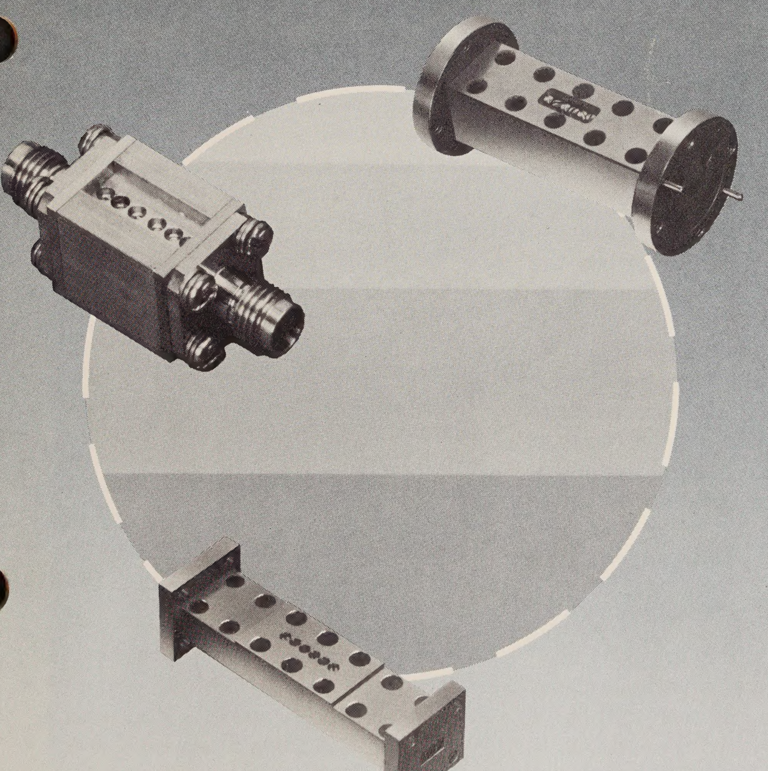
Combline Filters for Microwave and Millimeter-wave Frequencies

Part 1



WATKINS-JOHNSON COMPANY

Tech-notes

Modern microwave and millimeter-wave subsystems require high-performance bandpass filters due to the ever-increasing density of signals and system requirements for maximum sensitivity. This often translates to a bandpass filter requirement that has high selectivity and broad stop-band performance. A proven high-performance filter that can be manufactured to be very small is the combline filter, which uses coupled rectangular bars for the filter elements. Such filters can be designed with confidence that the final measured performance will closely match predicted performance. This type of filter can be built with microstrip, coax, or waveguide interfaces that satisfy the integration needs of modern subsystems.

At millimeter-wave frequencies, conventional designs for waveguide filters suffer from spurious passbands that occur above the required passband range of frequencies. The combline filter has excellent stopband performance, with spurious-free rejection up to about 4.5 times the center frequency of the passband and, in some cases, up to much higher frequencies. Part 1 of this article describes the basic theory and the design of combline bandpass filters. Part 2 of this article gives some practical examples of microwave combline-filter design and extends the design procedure to millimeter-wave frequencies (up to at least 50 GHz), with examples of filters with either coaxial or waveguide interfaces.

Applications of Combline Filters

Broad passband filters are often used at the front end of receiver systems to define the band of interest and reject unwanted signals while simultaneously attenuating any radiated signals from the local oscillators within the receiver. The combline filter has proved to be highly selective and easily manufacturable using the techniques described below. Filters with fractional bandwidths from 2% to 75% with orders up to 19 have been constructed in minimal volume. The design of these filters has been studied and automated with the aid of the personal computer such that filters can be easily designed and constructed in minimal time. One attractive characteristic of these filters is that only one tuning screw is required for each filter resonator, thus allowing the filter to be miniaturized and mechanically simplified. Tuning of the inter-element coupling is not required, and locking screws for the resonator tuning screws are not necessary due to the self-locking nature of the tuning screws used. This simplifies tuning access to only one surface of the filter.

Local-oscillator generation often relies upon the multiplication of some low-frequency, stable signal. Successive stages of multiplication cause unwanted harmonics and sub-harmonics to be generated in the multiplier chain. Therefore, a bandpass filter is needed to pass the required harmonic while simultaneously rejecting all other harmonics or sub-harmonics. This application is ideal for combline technology as it calls for narrow bandwidth filters with low passband insertion loss and high out-of-band rejection. High degrees of integration can be achieved by the use of microstrip interfaces with these filters [1].

Another problem with local-oscillator generation in receiver systems using multiplier chains is that some designs of bandpass filters have poor rejection at frequencies above the required passband, thereby allowing higher harmonics to multiply in the mixer stage used to downconvert the signal. This generates an apparent spurious signal at the frequency applied to the multiplier immediately preceding the filter. An example of this problem is given in reference [2] where a tripler is used to multiply a 14.833 GHz signal up to the required local oscillator frequency of 44.5 GHz. The downconverter described in this reference has an IF range from 4.0 to 18.0 GHz, which places the fundamental oscillator frequency of 14.833 GHz firmly within the IF range. The bandpass filter operating at 44.5 GHz not only has to reject the fundamental and 2nd harmonics, but also the 4th and 5th, the 5th and 6th, 6th and 7th etc. The reason for this is that these harmonic pairs can mix in the multiplier, generating a signal at the original oscillator frequency (the 5th and 6th harmonics at 74.165 GHz and 88.998 GHz can mix, resulting in a difference frequency of 14.833 GHz present at the IF port of the mixer).

Description of Combline Filters

The filters described in this article are realized as coupled rectangular coaxial transmission lines. The choice of this type of transmission line is due to the ease of machining and the wide variation in coupling coefficients (and, hence, bandwidths) that can be realized with these rectangular bars. Round rod designs and iris-coupled designs are also possible, but these designs tend to be narrowband types due to the limited coupling capacitance that can be achieved between the resonators. The added mechanical complexity of these round rod and iris designs is also a factor in their rejection in favor of the rectangular bar elements. Figure 1 shows one of the

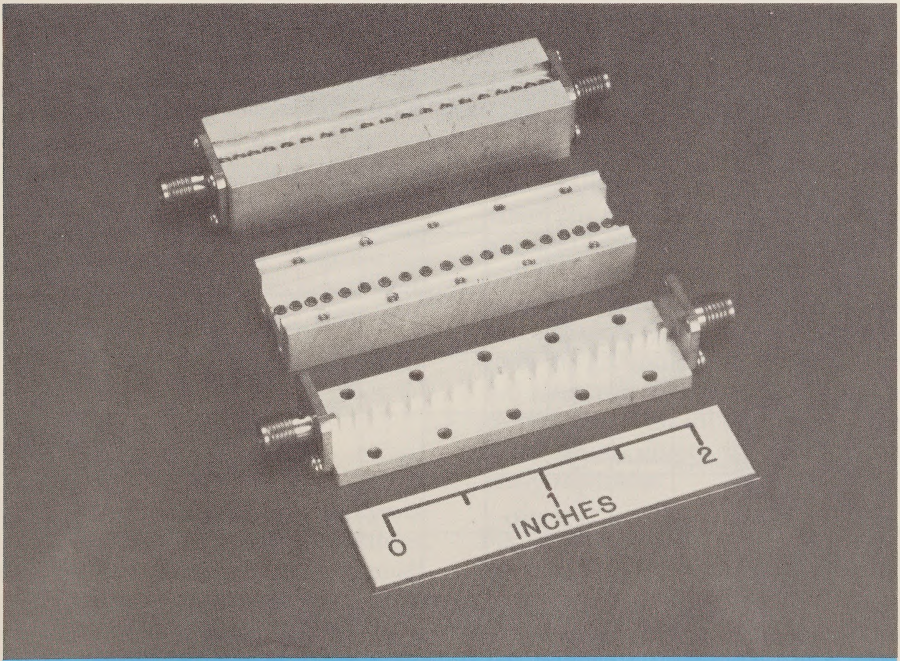


Figure 1. Construction of a 19th order, bandpass combline filter covering 5.83 GHz to 12.75 GHz.

combline filters that was designed as a 19th order, 0.01-dB equal-ripple Chebyshev filter. The bandwidth that this filter operates over is 5.83 GHz to 12.75 GHz (a fractional bandwidth of 74.5%), with a maximum passband VSWR of 1.5:1. This filter is a non-redundant design (all elements are resonators) that does not require any impedance transformation at the ends of the filter. Appropriate choice of phase length of the resonators (40 degrees), Chebyshev passband ripple and resonator node impedance produce a filter terminal impedance of 50.0 ohms. A very attractive feature of this filter is that there is only one tuning screw for each resonator which is intended to only adjust the resonant frequency of each rectangular bar. The design procedure is accurate enough so that inter-element tuning is not required. This greatly simplifies the mechanical construction and tuning procedure of the filter. A plot of the insertion loss performance of this filter is given in Figure 2.

Another example of a combline filter is the 3rd order, 0.01-dB equal-ripple Chebyshev filter, with a bandwidth of 200 MHz centered at 8.0 GHz (2.5% fractional bandwidth) that is shown in Figure 3. The filter exhibits an insertion loss of about 1.0 dB with better than 1.5:1 VSWR, and is designed as a drop-in, microstrip-compatible device that is used to achieve very high rejection of unwanted harmonics at the output of a multiplier chain [1]. The raised section on the filter housing is used to form a section of the inner wall of a subassembly that is used to electrically isolate the cavity containing the filter from the cavity containing the multipliers. This

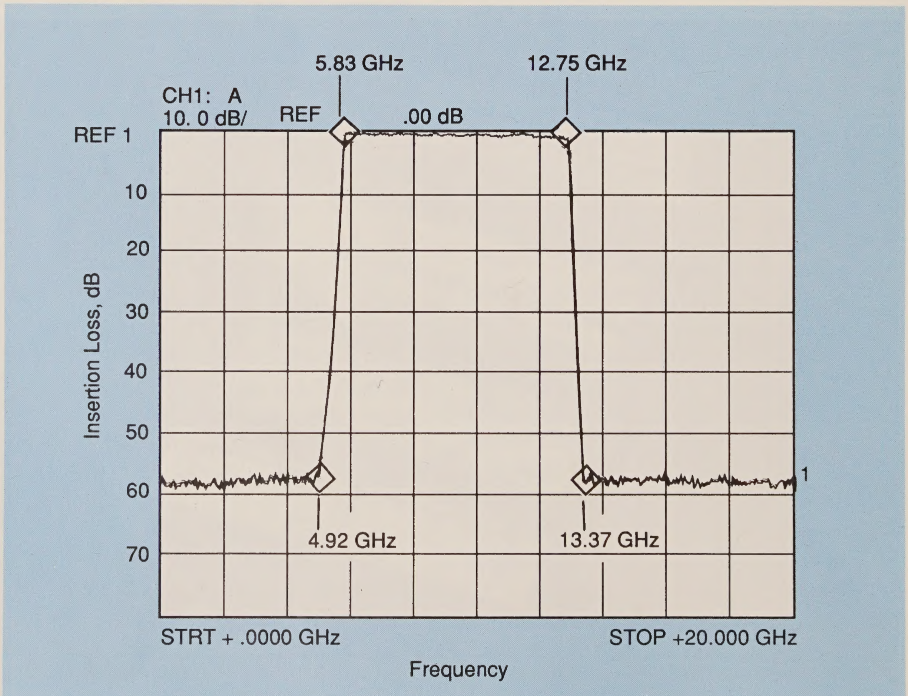


Figure 2. Insertion-loss performance of 5.83 GHz to 12.75 GHz combline filter.

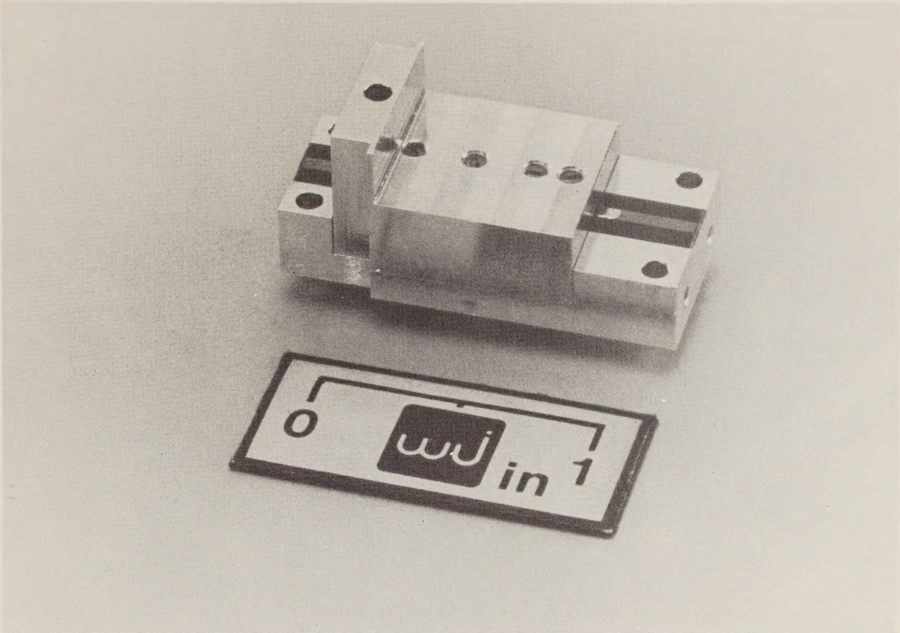


Figure 3. Combline filter with microstrip interfaces.

technique helped maintain the high suppression of the unwanted harmonics to better than 60 dBc.

In order to make a filter that has a good input/output VSWR and is easily tuned, care must be taken to design the filter, with or without impedance transformation, so that a good match can be maintained with the circuitry at the filter interfaces. Both coaxial and waveguide interfaces that operate in the microwave and millimeter-wave frequency range can be built using the following design principles.

Basic Filter Theory and Realization

The basic combline filter design procedure follows that given by Matthaei, Young and Jones [3], but with modified end transformer elements. The design procedure described below generates normalized self-capacitance-to-ground and mutual-coupling-capacitance for the array of transmission lines in the filter. The standard design given in [3] has the end transformer elements shorted on the opposite side from the resonator elements within the filter, as shown in Figure 4. This creates a mechanical alignment problem as well as problems in tuning accessibility. To overcome the problem of the ground connection on the end elements being on the opposite end from the resonator ground connections, the impedance transformers are modified. This modification inverts the ground connection on these end elements resulting in a common ground plane for all elements in the filter, which allows the filter to be easily realizable.

Figure 4 shows a representation of the combline filter, in conventional form, with extra end elements (elements 0 and n+1 with their ground connection on the opposite side from the resonators) that are used to transform the terminal impedance to the

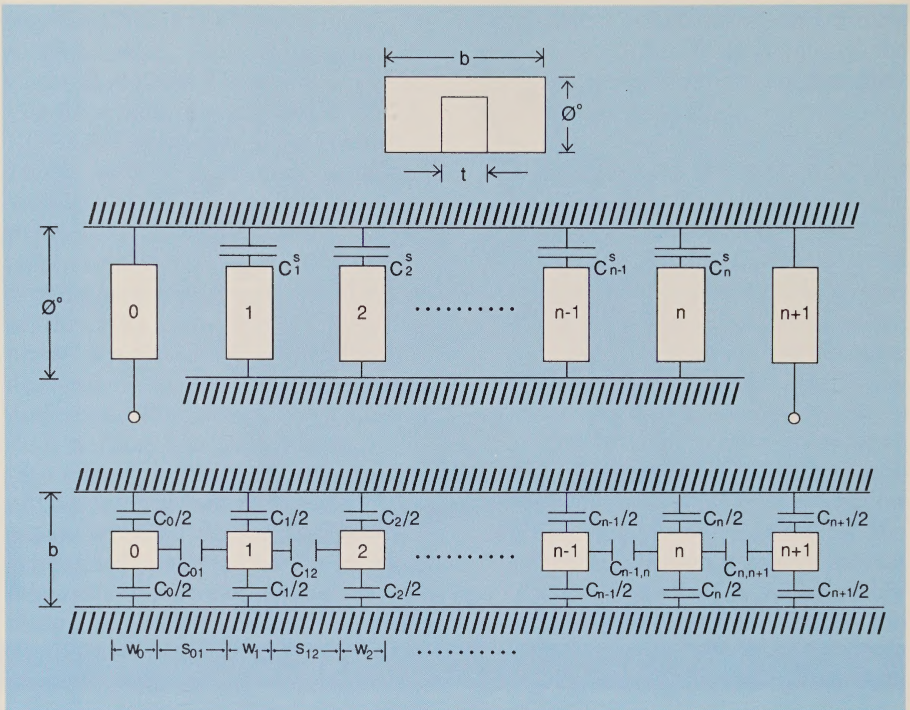
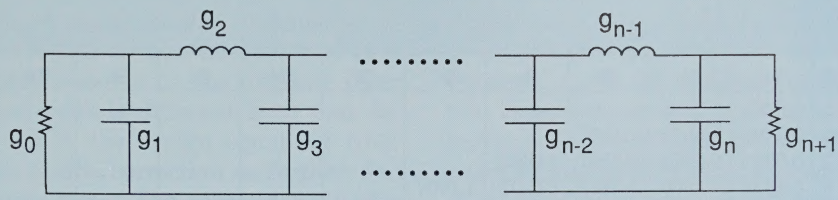


Figure 4. Conventional combline bandpass filter realized using coupled rectangular rods.

required impedance for the filter resonators. It is assumed that the input and output terminal impedances, Z_o, are the same (usually equal to 50 ohms). The resonators have an electrical length of $\phi°$, which is less than 90° at the resonant frequency (typically 45° at the center frequency of the filter). The lumped capacitors, C_j^s, are used to resonate the filter elements. The filter is described in terms of the normalized self capacitance to ground per unit length, C_j/E, and the normalized mutual coupling capacitance per unit length, $C_{j,\,j+1}/E$ (where $E = E_oE_r$ is the dielectric constant of the medium). These capacitances are then used to determine the width, w_j, and the gaps, $s_{j,\,j+1}$, of the filter.

The design starts with the n^{th} order low-pass prototype filter parameters ($g_0, g_1, \ldots g_{n+1}$) as shown in Figure 5, with the derivation of the "g" values for an equal-ripple Chebyshev design given in Figure 5. Due to its superior passband performance, it is normal to select a Chebyshev design although any of the standard sets of "g" values can be used. Table 1 gives the "g" values for the most commonly designed filter (a 0.01 dB equal-ripple Chebyshev).

As can be seen from the "g" values in Table 1, the "g" values are not symmetrical and the terminating admittance (or conductance) values, g_{n+1}, do not equal 1 for even values of n. In a low-pass design, this would result in an asymmetric design with different terminating impedances at the ends of the filter, but for the bandpass designs described in this article, the transformation from low pass to bandpass creates a symmetrical structure and equal terminations. However, should different terminating imped-



Maximally Flat:

$$g_0 = g_{n+1} = 1$$
$$g_j = 2\sin[0.5(2j-1)\pi/n]$$

Chebyshev :

$$g_0 = 1$$
$$g_{n+1} = 1 \text{ (for n odd)}$$
$$g_{n+1} = \coth^2(B/4) \quad \text{(for n even)}$$
$$g_j = (4a_{j-1}a_j)/(b_{j-1}g_{j-1})$$
$$r = \text{In-band ripple, dB}$$

where,

$$B = \ln[\coth(r/17.37)]$$
$$a_j = \sin[0.5(2j-1)\pi/n]$$
$$b_j = \sinh^2[0.5\ln\{\coth(r/17.37)\}/n] + \sin^2(j\pi/n)$$

Mapping Function:

$$f_0 = (f_2+f_1)/2$$
$$W = (f_2-f_1)/f_0$$
$$f'/f_1' = 2[(f-f_0)/f_0]/W$$

where,

f_0 = Bandpass center frequency
f_1 = Bandpass low cut-off frequency
f_2 = Bandpass high cut-off frequency
W = Fractional bandwidth
f = Bandpass frequency
f' = Lowpass frequency
f_1' = Lowpass cut-off frequency

Figure 5. Low-pass prototype filter and low-pass-to-bandpass mapping function.

ances be required, then this can easily be accommodated in the design of the end transformer elements.

All the filters described in this article are Chebyshev designs that have equal ripple passbands. This results in peaks in the return loss of the filter (the worst-case return loss caused directly by the ripple mismatch) with a number of maximum return-loss peaks equal to the order of the filter. Due to the approximations used in the design, it is unusual to obtain the design return loss except for narrow bandwidth designs. A 0.01-dB equal-ripple Chebyshev filter should have a theoretical in-band return loss of 26.4 dB, and a 0.1-dB ripple design a 16.3 dB return loss. However, the design return loss is rarely achieved for broad bandwidth designs and, therefore, the filters are usually designed for 0.01-dB ripple (26.4 dB return loss) with the expectation that the return loss as measured on a tuned filter will have degraded by several dB (usually to about 15 dB).

n	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}
1	0.0960	1.0000								
2	0.4488	0.4077	1.1007							
3	0.6291	0.9702	0.6291	1.0000						
4	0.7128	1.2003	1.3212	0.6476	1.1007					
5	0.7563	1.3049	1.5773	1.3049	0.7563	1.0000				
6	0.7813	1.3600	1.6896	1.5350	1.4970	0.7098	1.1007			
7	0.7969	1.3924	1.7481	1.6331	1.7481	1.3924	0.7969	1.0000		
8	0.8072	1.4130	1.7824	1.6833	1.8529	1.6193	1.5554	0.7333	1.1007	
9	0.8144	1.4270	1.8043	1.7125	1.9057	1.7125	1.8043	1.4270	0.8144	1.0000

Table 1. Low-pass "g" values for a 0.01 dB equal-ripple Chebyshev filter ($g_o = 1$ and low-pass cutoff freq. = 1 radian)

The low-pass-to-bandpass mapping is also given in Figure 5, which is essentially an approximate, narrow-band function, but has produced acceptable passband responses for fractional bandwidths up to 75%. This low-pass-to-bandpass mapping does not accurately predict the role-off of both sides of the filter response due to this approximation, the high side of which is also affected by the electrical length of the resonators.

After fixing the terminating impedance of the filter (usually 50 ohms) the starting node impedances of the resonator elements should be selected. This node impedance is that impedance which would result from the self capacitance to ground with nearest neighbors grounded. This node impedance has a very large effect on the realizability of the filter as well as the unloaded Q of the resonator (and, hence, the loss of the filter). It has been found that 60 ohms is a very good starting node impedance that results in filters that can be mechanically realized and that have low insertion loss. It is a simple case to iterate the filter design using different values of starting node impedance to optimize the realizability of the filter. However, for relatively large bandwidths of about an octave or more, the first and last elements (the impedance transformers) tend to be unrealizable. The best solution to this problem is to discard the impedance transformer elements and adjust the resonator phase length, passband ripple and node impedances until a match to the terminating impedance is achieved (usually 50 ohms). For intermediate bandwidth filters, some combination of all of the above may be necessary to achieve a mechanically realizable design. It is also common to iterate the design so that the widths of all the resonators are equal, thereby simplifying the mechanical details and measurement after machining. The lumped tuning capacitance, normalized self and mutual capacitances are then calculated as described in [3].

Determination of the Widths and Gaps of the Bars

The normalized self and coupling capacitances are used to determine the width and gaps of the rectangular bars in the filter. The following procedure is only approximate as it assumes that there is no coupling other than to nearest neighbors. In reality this is not the case and is a prime reason why the ultimate filter bandwidth is different from that depicted in the design equations (this bandwidth correction is further discussed later). The parameter t/b (the ratio of resonator thickness to cavity width) as shown in Figure 4 is first selected. Experience has shown that this t/b ratio should be between 0.2 and 0.4, and the following information covers t/b = 0.2 or 0.4, as these two values can be used for the majority of filter designs. Care must be taken in the selection of t because transverse resonances can occur that distort the filter response. A good rule of thumb is to try to keep both the thickness, t, and the width, w, less than the cavity height.

The normalized gap, s/b, is determined using the appropriate value of t/b and the normalized coupling capacitance, $C_{j,\ j+1}/E$. Figure 6 shows a graphical representation of s/b versus $C_{j,\ j+1}/E$ for various values of t/b. This curve, as well as all the other curves used in determining the dimensions of the filter elements, can be fitted to a polynomial expression of the form:

$$y = a_1 + a_2 \cdot \log(x) + a_3/x + a_4/x^2 + a_5/x^3$$

In this expression, the independent variable, x, is replaced by $C_{j,\ j+1}/E$, and the dependant variable, y, is replaced by s/b. Using this expression allows all the required curves to be expressed in a closed form by the appropriate selection of the coefficients a_1 through a_5, resulting in better than 0.5% accuracy. The curves published in references [3,4,5] were curve-fitted by a least-squares fitting routine that utilized Gauss-Jordan elimination. Table 2 gives the values

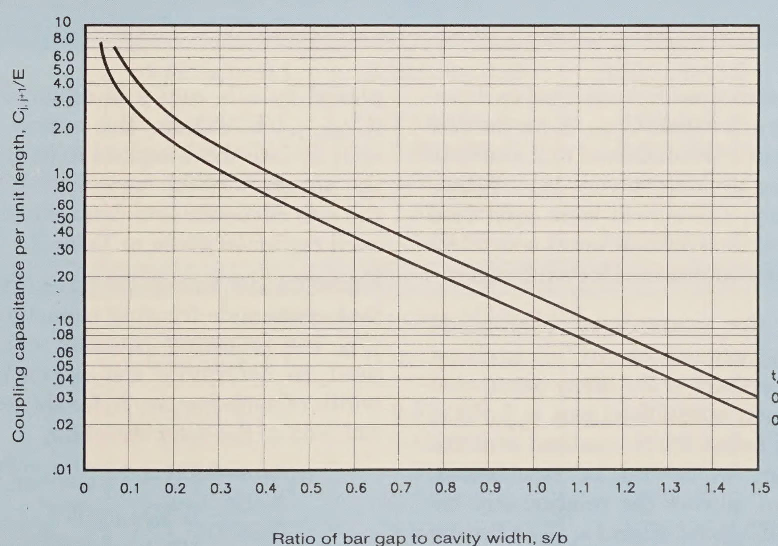


Figure 6. Normalized inter-bar coupling capacitance per unit length vs normalized bar gap.

$$y = a_1 + a_2 \cdot \log(x) + a_3/x + a_4/x^2 + a_5/x^3$$

Where, $x = C_{j,\,j+1}/E$

$y = s/b$

Coefficients for case of t/b = 0.2			
	$20 \geq C_{j,\,j+1}/E > 7.0$	$7.0 \geq C_{j,\,j+1}/E \geq .8$	$0.8 > C_{j,\,j+1}/E \geq .03$
a_1	1.09754E-2	-1.46086E-1	2.95293E-1
a_2	-5.16596E-4	4.91802E-2	-2.79955E-1
a_3	-6.62018E-2	6.42149E-1	1.07105E-2
a_4	2.95989E 0	-2.18154E-1	-3.67023E-4
a_5	-8.13567E 0	3.43861E-2	4.75300E-6

Coefficients for case of t/b = 0.4			
	$20 \geq C_{j,\,j+1}/E > 7.0$	$7.0 \geq C_{j,\,j+1}/E \geq .8$	$0.8 > C_{j,\,j+1}/E \geq .03$
a_1	-9.92689E-2	-6.39497E-2	3.99850E-1
a_2	2.24205E-2	1.63302E-2	-2.97861E-1
a_3	1.27579E 0	7.23784E-1	6.04207E-3
a_4	-4.47165E 0	-3.22789E-1	-2.57145E-4
a_5	1.12117E 1	7.28344E-2	4.05462E-6

Table 2. Coefficient values used to determine s/b vs normalized coupling capacitance per unit length for t/b = 0.2 and t/b = 0.4

of the above coefficients used in determining s/b versus $C_{j,\,j+1}/E$ for the two values of t/b considered (0.2 and 0.4). In order to achieve very high fitting accuracy, the curves were split into three regions and different sets of coefficients determined for each region.

Next, the normalized even-mode fringing capacitance, $(C'_{fe})_{j,\,j+1}/E$, of the inner bars of the array associated with each normalized gap, $s_{j,\,j+1}/b$, is needed before the normalized width of the bar, w_j/b, can be established. Figure 7 shows the relationship between $(C'_{fe})_{j,\,j+1}/E$ and $s_{j,\,j+1}/b$ for the two cases of t/b = 0.2 and 0.4. The curves shown in Figure 7 are also curve-fitted in the same manner as described above. In this case, x is replaced by s/b, and y is replaced by $(C'_{fe})_{j,\,j+1}/E$. Again, the curves are split up into three regions to enhance the accuracy of the curve fitting and the sets of coefficients determined for each region as given in Table 3.

Knowing the values for the normalized even-mode fringing capacitance, C'_{fe}, the following equation can be used to determine the normalized width of each bar, w_j/b, for the internal bars of the filter structure.

$$w_j/b = 0.5\,(1-t/b)\,[0.5\,(C_j/E) - (C'_{fe})_{j-1,\,j}/E - (C'_{fe})_{j,\,j+1}/E]$$

If the resulting width of the bar given by the above equation is so small that $w/b < 0.35\,(1-t/b)$, then the

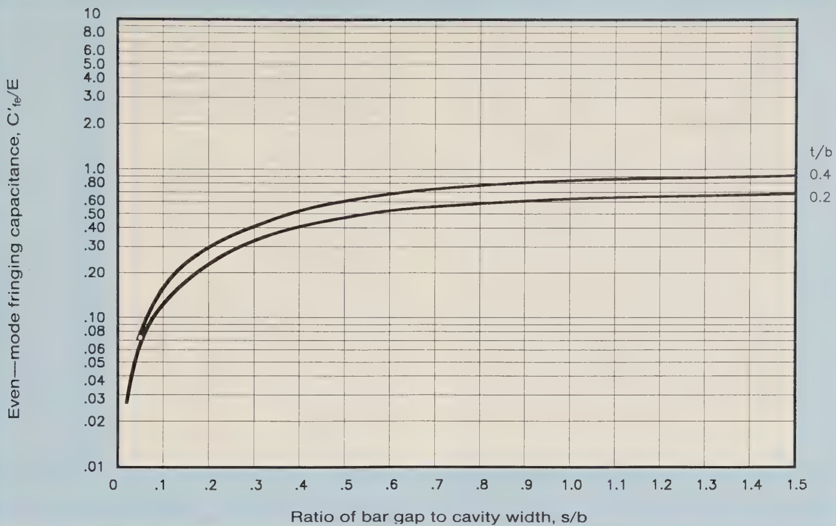


Figure 7. Normalized even-mode fringing capacitance per unit length vs normalized bar gap.

$$y = a_1 + a_2 \cdot \log(x) + a_3/x + a_4/x^2 + a_5/x^3$$

Where, $x = s/b$

$y = C'_{fe}$

Coefficients for case of t/b = 0.2			
	$0.02 \leq s/b < 0.1$	$0.10 \leq s/b < 0.55$	$0.55 \leq s/b < 1.5$
a_1	5.48992E–1	7.14999E-1	6.97928E–1
a_2	2.56114E–1	2.59646E–1	-2.86718E–2
a_3	1.92969E-2	-4.51493E–2	1.04566E–1
a_4	-2.98863E–4	7.92733E–3	-2.32257E–1
a_5	2.10594E–6	-3.39716E–4	6.04042E–2

Coefficients for case of t/b = 0.4			
	$0.02 \leq s/b < 0.1$	$0.10 \leq s/b < 0.55$	$0.55 \leq s/b < 1.5$
a_1	6.70979E–1	8.37990E–1	-7.83216E–1
a_2	2.91714E–1	5.42268E–1	1.01193E 0
a_3	1.89944E–2	7.61665E–2	2.64422E 0
a_4	-2.62532E–4	-2.14965E–3	-1.25913E 0
a_5	1.71062E–6	1.87461E–5	2.30692E–1

Table 3. Coefficient values used to determine the normalized even-mode fringing capacitance vs s/b for t/b = 0.2 and t/b = 0.4

normalized width is corrected to a new normalized width, w′/b, as follows:

$$w'/b = [0.07\,(1-t/b) + w/b]/1.2$$

All that remains is to determine the width of the end bars of the filter structure. As these end bars only have adjacent bars on one side, then one of the normalized even-mode fringing capacitance terms needs to be replaced by the normalized fringing capacitance, C'_f/E, of an isolated bar. Figure 8 shows a plot of C'_f/E vs t/b, but for the cases given here, we are only concerned with the value for C'_f/E at $t/b = 0.2$ and $t/b = 0.4$. The following values for C'_f/E are determined from Figure 8.

$$C'_f/E = 0.690 \text{ for } t/b = 0.2$$
$$C'_f/E = 0.915 \text{ for } t/b = 0.4$$

This completes the design procedure for conventional end-element combline filters.

End-Section Impedance Transformer Modification

As discussed previously, it is desirable to have the grounding of the impedance-transforming end elements on the same side as the grounds for the resonators in the filter. The modified end-transformer section is formed by grounding the first element in the filter on the same side as the other elements in the filter, adding a shunt

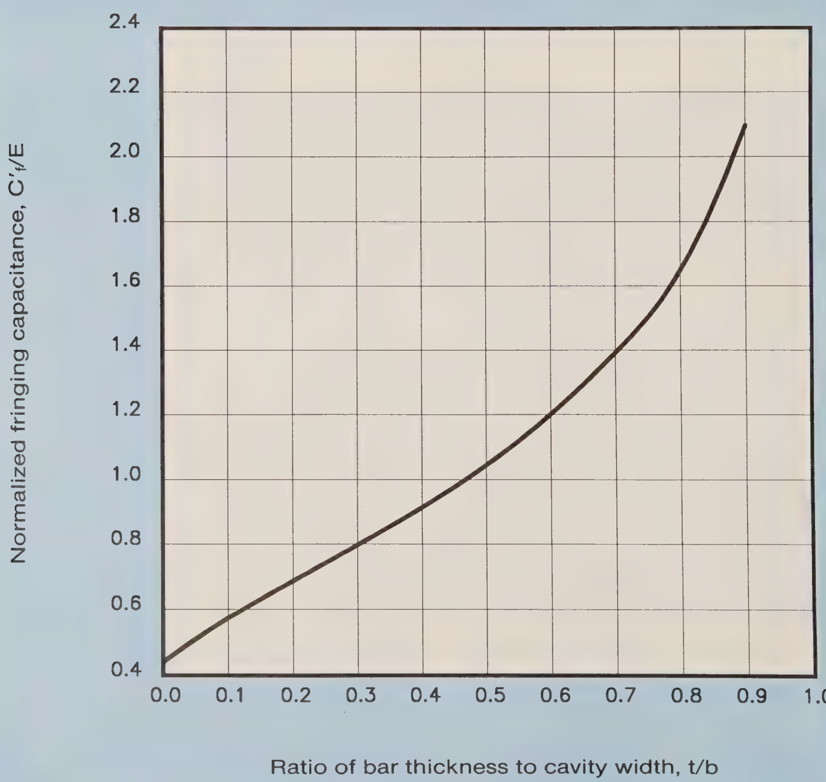


Figure 8. Normalized fringing capacitance per unit length vs normalized bar thickness.

capacitance to the other end of the line and feeding the input to the connection of the new capacitor and the top of the first element. Figure 9 shows the conventional and modified end-transformer sections and their two-wire equivalent circuits.

Using this new end section, the admittances of the two shunt shorted lines, Y_0 and Y_1, do not change from the original design. The series transmission line is transformed into a series short-circuited line, but has the same admittance as the original line, Y_{01}. Thus, the widths and gaps of the elements do not change from the old to the new design. Physically, the circuit is completed merely by reversing the first element and introducing a lumped capacitor to the open end of this element where the input connection exists and by slight adjustment of the first resonator lumped tuning capacitance, C_1. The modified end impedance transformer has been used on filters with fractional bandwidths from 2% to 75% with very good results. The simplified mechanical structure means that all the posts and gaps are machined from one side and very small dimensions and tight tolerances can be maintained. The limit to the size of these filters is commonly dictated by the size of the tuning screw, as the filter elements can easily be miniaturized.

When using the above design procedure, the performance of the resulting filter always has a wider passband than the design equations would theoretically indicate. This is due to coupling to other than nearest-neigh-

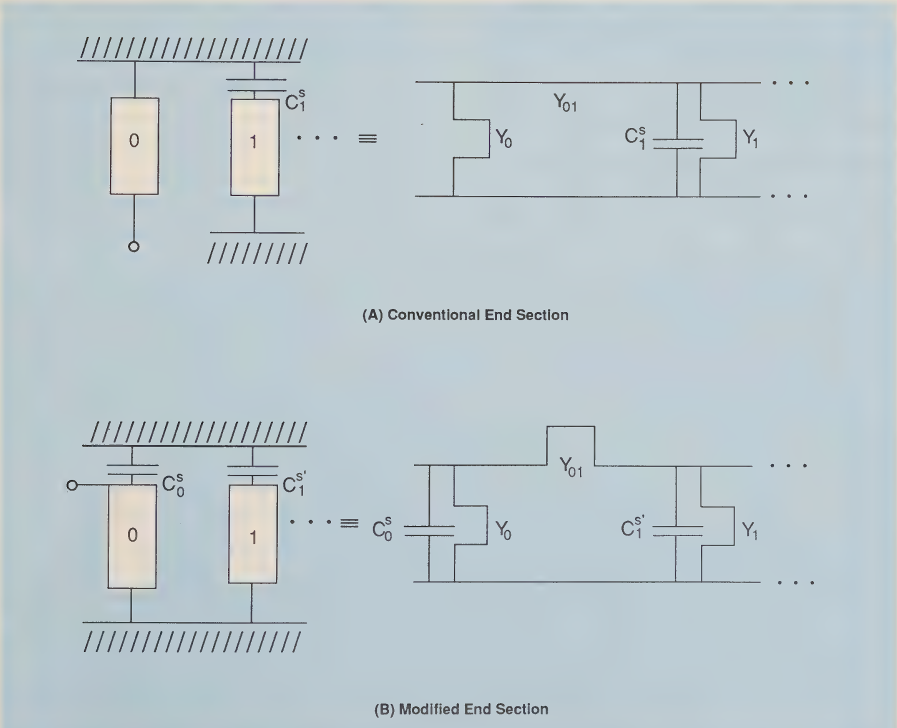


Figure 9. Conventional and modified combline impedance transformer end sections.

bor elements that is not considered in the derivation, the choice of low-pass to bandpass mapping function and other approximations used in the design. The resulting bandwidth is a function of both electrical and mechanical design parameters, and an analytic description of this change-in-bandwidth phenomenon is not known. Experience has shown that the required passband needs to be reduced to approximately 80% prior to running through the design process. This results in a design that would theoretically have a narrower bandwidth than required, but when the filter is actually constructed, the measured bandwidth will be close to the required bandwidth. Empirical data is used to more accurately predict the bandwidth reduction required, as this 80% correction can be in error by as much as 10% (or more in certain circumstances), depending upon the particular electrical and physical details.

Wideband Designs Without Impedance Transformation

The filter shown in Figure 1 is derived in a slightly different manner. One limitation of the above design procedure is that the impedance transformer elements at the ends of the filter become too thin to realize when wide bandwidth (around an octave or more) filters are required. The reason for this is that as the bandwidth increases, the filter impedance drops until it approaches the terminating impedance (usually 50 ohms). Accordingly, the impedance-transforming end element's width diminishes, as little or no impedance-transformation is necessary. The solution is to simply discard the impedance-transforming end element and connect directly to the first resonator element. An alternative schematic for the end elements of the filter is shown in Figure 10, where the impedance matching is achieved through the use of a transformer. In order to eliminate the end elements, the filter impedance, Z_{T1}, is set to the terminal impedance, Z_0. This occurs when the impedance transformer has a 1:1 turns ratio and can, thus, be discarded. Effectively, line 0 and n+1 shown in Figure 4 are removed and the input and output con-

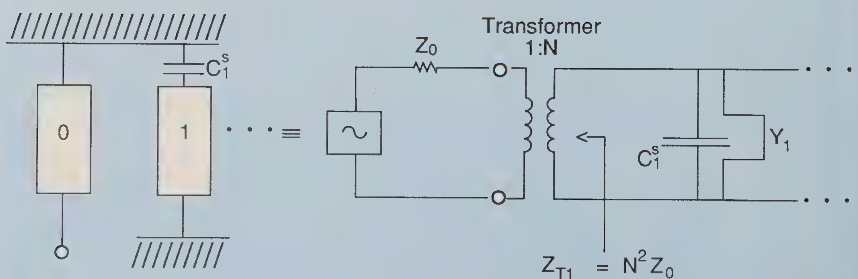


Figure 10. Transformer representation of combline end elements. For broad bandwidth designs where a 1:1 turns ratio applies ($Z_{T1} = Z_0$), the end elements can be discarded.

nections are made directly to the tops of resonators 1 and n, respectively.

For any particular bandwidth, the terminating impedance that this new filter wants to see is fixed by the phase length, in-band design ripple and node impedance of the resonators. If this filter terminating impedance is not equal to 50 ohms, then one or all of these parameters can be adjusted in order to achieve a 50-ohm impedance. Care has to be taken with adjustments to the phase length and ripple specifications. As the phase length of the resonators is reduced, then more tuning capacitance needs to be provided. For very small resonator cross sections and small tuning screws, only limited capacitance can be achieved before the gap at the top of the resonator becomes unrealistically small, and the filter becomes untunable. The ripple specification determines the passband VSWR of the filter, and increasing this ripple too far can produce unacceptable values of VSWR. Reduction of the ripple specification is preferred but this does reduce the selectivity of the filter and extra resonators may be needed to maintain the required rejection characteristics.

Another interesting feature of the design without impedance transformation is that the absolute bandwidth of the filter is approximately constant over a reasonably wide range of frequencies [6]. This occurs where the phase length of the resonators is set to 53° at the center frequency for the range considered. Another design approach, which is more applicable to narrow-band cases, has been shown [7] to also have constant bandwidth tuning centered on a resonator phase length of 53°. However, in this case, the narrow bandwidth necessitates the use of an impedance transformer. This impedance transformer takes the form of the modified end section shown in Figure (9B), but the electrical parameters of the components are different from those used to simply modify the end elements of the conventional design shown in Figure (9A). This constant bandwidth design is limited in bandwidth to narrow-band designs as the impedance-transformation end element quickly becomes too thin and, therefore, unrealizable as the bandwidth is increased.

Conclusion

The design of various forms of combline filters for microwave and millimeter-wave use has been described in this part of the article. From a practical sense, the conventional design given in [3] with the modified end elements described above (Figure 9B), is used for the majority of applications. The two other types of designs discussed above, the broad and narrow constant bandwidth designs, are also used for those applications where the standard design is impractical. All three of these design procedures have been written into a computer program that operates very quickly due to the closed form of the expressions used. Examples of actual designs using the above methods will be given in the second part of this article, and the modifications necessary to interface these filters to millimeter-wave waveguide will be covered, with examples given.

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Notes

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