Design of Microwave Filters

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Invited Paper

***Abstract*—A survey of the major techniques used in the design of microwave filters is presented in this paper. It is shown that the basis for much fundamental microwave filter theory lies in the realm of lumped-element filters, which indeed are actually used directly for many applications at microwave frequencies as high as 18 GHz. Many types of microwave filters are discussed with the object of pointing out the most useful references, especially for a newcomer to the field.**

***Index Terms*—Bandpass, cavity, ceramic, coaxial, combline, diplexers, evanescent mode, filters, hairpin line, high-pass, high temperature, interdigital, low-pass, lumped element, microstrip, microwave, multiplexers, parallel coupled line, planar, stripline, superconducting, waveguide.**

I. Role of Lumped-Element Filters in Microwave Implementations and Design

SIGNIFICANT developments have taken place since the publication of the previous survey published in the 1984 Special Centennial Issue of this TRANSACTIONS [1]. One important aspect of microwave filters, which was not covered then, is the lumped-element filter, which was starting to make an impact at about that time, actually beginning in the late 1970s. Lumped-element filters are now used at microwave frequencies up to about 18 GHz, and form a large percentage of microwave filters produced by the industry. The unloaded Q, which is realizable depends on frequency, but averages about 200, and values over 800 may be achieved at lower frequencies, e.g., at 170 MHz [2]. Such figures compare favorably with microstrip, and production costs are quite low. Of course, dimensions are much smaller than distributed filters, which is a major advantage. However, there is no escaping the use of larger distributed filters when insertion loss and perhaps power handling are of major concern, unless superconducting filter technology is employed.

An important academic aspect of lumped-element filters is that their study is an essential part of the understanding of distributed filters, which are based to a large extent on lumped-element theory. Thus, many or perhaps even most filter designs commence from a lumped-element low-pass prototype filter, and the concepts of susceptance slope parameters and coupling coefficients unify the theories (see Section III).

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Filters may be classified into categories in several ways, one being into different classes of response functions, defined in terms of the location of the poles of the insertion-loss function and of the zeros within the passband. The zeros are usually spaced throughout the passband to give an equiripple or Chebyshev response since this is far more optimum and superior to the maximally flat or Butterworth response, which is rarely used. As far as the poles are concerned, the most common type of filter response has these located all at dc or infinity and is often described as an all-pole Chebyshev filter, or simply as a Chebyshev filter. When one or more poles are introduced into the stopbands at finite frequencies, the filter is known as a generalized Chebyshev filter or as a pseudoelliptic filter. The special case where the maximum number of poles are located at finite frequencies such that the stopbands have equal rejection level is the well-known elliptic function filter. This is now rarely used since it has problems in practical realization and is not optimum when specific stopbands are required—one seldom needs rejection up to infinite frequency. It is almost always better to place the poles where they are most needed, and also to minimize their number, since each additional finite frequency pole may increase complexity and expense.

The above discussion relates equally to the main categories of filters defined in terms of the general response types of low-pass, bandpass, high-pass, and bandstop. In the case of bandstop filters, the poles are placed in the bandstop region, and the zeros elsewhere, as produced typically by means of a low-pass to bandstop frequency transformation.

There has been rather sparse literature on the topic of lumped-element (*LC*) filters designed for operation at microwave frequencies. This may seem surprising considering the basic role of *LC* filter theory, but textbooks fail to proceed beyond the most elementary design stage, which is, for example, the application of low-pass to bandpass transformations. The problem with this is that if narrow band, then the filters that result are unrealizable because of the resulting wide spread of element values, many of which become quite impractical. It is necessary to introduce loose external coupling networks to transform the impedance levels and to introduce impedance inverters and/or carry out network transformations. The objective is to arrive at designs where typically all inductors, which may be either in series or shunt, or in many designs, both series and shunt, have similar values corresponding to a mid-band reactance in the range 40–100 Ω. This is the same condition desired in the design of coaxial filters, illustrating one of several similarities between *LC* and distributed filter design. Some of these principles are described in two papers that appeared in *Microwave Journal* in the 1980s [3], [4]. Several of the design principles are briefly sketched, but many details are required for a more complete understanding. Commercially available design programs are either very cumbersome in operation, requiring several arcane network transformations to give realizable filters, or do not have the basic capability.

It is interesting that one of the keys to satisfactory design is given in the fundamental 1957 paper of Cohn [5]. The paper actually describes capacitively coupled *LC* filters with all-shunt resonators having equal inductances. However, this is somewhat limiting in having all but one of the poles at dc and, as pointed out in [3] and [4], more general filters are required. One of the means of achieving this is to place the desired number of poles at dc and infinity and to design using exact synthesis, which has the advantages of precision and no bandwidth limitations caused by approximations. The best technique for such synthesis involves the transformed variable as described by Orchard and Temes [6]. Although this is not a simple topic, it is well worth acquiring the techniques. A much simpler paper describing applications to the design of distributed filters was given by Wenzel [7], and it is recommended that it should be studied first. The techniques described in [7] are almost identical to those for the synthesis of *LC* filters, the difference being the initial application of a simple Richards' transformation [1, Sec. III]. In addition, [7] introduces the unit element or cascaded section of commensurate transmission line, which is required for distributed filters and has no counterpart in *LC* filters.

Pseudoelliptic filters are also best designed using exact synthesis. One procedure is to synthesize a low-pass prototype, which is then resonated to form a bandpass filter. Network transformations are used to couple into the filter using, for example, series capacitors, and also to introduce redundant shunt *LC* elements where necessary in order to eliminate "floating nodes." In some cases, it is convenient to synthesize a bandpass filter directly without the low-pass prototype stage, e.g., when the poles are asymmetric with respect to the passband, e.g., when there are poles on just one side of the passband.

Pseudoelliptic filters may be designed to give poles by means of cross coupling between nonadjacent resonators, and further discussion is presented in Section III, with particular reference to cascaded-quadruplet (CQ) and cascaded-trisection (CT) filters [10], [11], [39].

II. Universality of Unloaded Q in Microwave and Lumped-Element Resonators

Returning to the topic of unloaded Q, it is instructive to consider the definition of that for the series *LC* resonator shown in Fig. 1.

The loss of the resonator is due to the resistance R. If the current at the resonant frequency ω is i, then the Q is defined as the ratio of the inductor reactance to the series resistance, i.e.,

$$Q = \frac{\omega L}{R} = \omega \frac{\frac{1}{2}Li^2}{\frac{1}{2}Ri^2} = \omega \frac{\text{Energy stored in magnetic field}}{\text{Energy dissipated in the resistor}}. \tag{1}$$

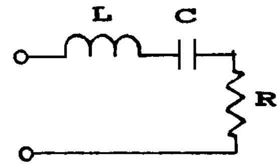


Fig. 1. Lumped-element resonator.

A similar result applies to the case of a shunt resonator. In either case (series and shunt), at resonance, the energies stored in the magnetic and electric fields are equal. We note that (1) is precisely the definition of unloaded Q for any form of resonator, whether lumped or distributed. Actually, there is no need even to make a distinction between the two since the Q of an *LC* resonator is determined primarily by that of the inductor, and its Q depends on the ratio of its volume to surface area as in any electromagnetic cavity resonator. It has been established that the Q of any normal (nonsuperconducting) metallic RF cavity is given by a relationship of the form

$$Q = Kb\sqrt{f} \tag{2}$$

where b is a linear dimension of the cavity, f is the resonant frequency, and K is usually constant for any given type of resonator. It is remarkable that K varies within a surprisingly narrow range for a variety of resonators of the TEM type, including lumped elements, microstrip, stripline, and coaxial lines. It is common in the U.S. to use units of inches for b and gigahertz for f, and then K is of the order 1500–3600 for the various cavities, with b defined variously as the mean diameter of the coil for *LC* resonators, the thickness of the substrate for microstrip, the ground-plane spacing for stripline, and the outer diameter of coaxial line. These figures relate to practical values of Q for copper or silver rather than to the theoretical values, which are almost impossible to realize because of surface imperfections, and a de-rating factor of 0.7 has been incorporated in any theoretical calculations.

In the case of waveguides, the relationship between Q and dimensions is more complex than the simple equation (2), although the Q retains the same type of proportionality to frequency and approximately to a linear dimension. However, the effective K value is much higher for a given volume than for the TEM resonator cases. Later, in Section III-A, it will be seen that the K value in a combline filter increases as the ground-plane spacing increases to an appreciable fraction of a wavelength, supporting the propagation of waveguide modes, hence, leading to higher Q than would be expected from normal TEM-type resonators.

III. Types of Microwave Filters and Design Information

This section describes the main types of distributed filters, where the various types might be used, and the major sources of design information. The categories considered are combline, interdigital, parallel-coupled-line bandpass and bandstop, ring and patch filters, and stepped-impedance filters. The several media for implementation include waveguide, dielectric resonators, coaxial lines, evanescent-mode filters, and various printed circuit filters in microstrip, stripline, and suspended substrate. Acoustic filters are beginning to make inroads into the market for miniature filters for high-volume production, typically for use in wireless telephones, and details will be found in the papers on regarding microwave acoustics in this TRANSACTIONS.

Superconducting filters are also coming into vogue when the ultimate in performance is required, i.e., taking advantage of the extremely high unloaded Q due to the virtual elimination of resistive losses.

In some instances, the cited references may not be to the earliest invention, but to a more convenient or simpler description. In such later papers, the earliest work is almost always cited.

Design theories for distributed filters are similar to those for *LC* filters, and are based either on exact synthesis or narrow-band approximation. The latter is summarized rather succinctly in [8, pp. 432–433]. The important concepts here are the definition of the coupling coefficients between resonators and between the source and load resistors and the end resonators, and also the susceptance slope parameter of a shunt resonator, which may be of any type, lumped or distributed, i.e.,

$$\text{Susceptance slope parameter } b = \frac{\omega}{2}\frac{\text{dB}}{d\omega}\bigg|_{\omega=\omega_0} \tag{3}$$

where B is the susceptance of the resonator, e.g., for a lumped resonator

$$B = \omega C - 1/(\omega L) \tag{4}$$

which, after applying the condition $\omega_0 = 1/\sqrt{(LC)}$, gives

$$b = \omega_0 C. \tag{5}$$

For a quarter-wave short-circuited transmission line of characteristic admittance Y and electrical length θ

$$B = Y \tan\theta \tag{6}$$

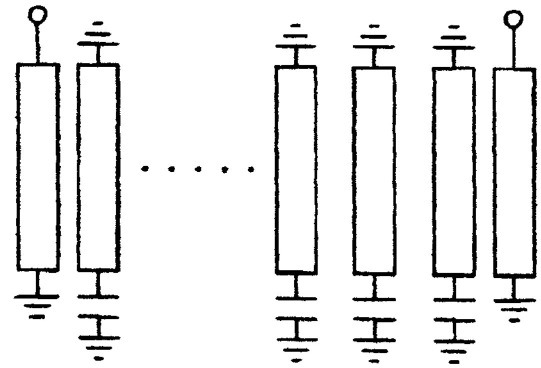
and substituting this into (3), we find

$$b = Y\theta_0/2 = Y\pi/4. \tag{7}$$

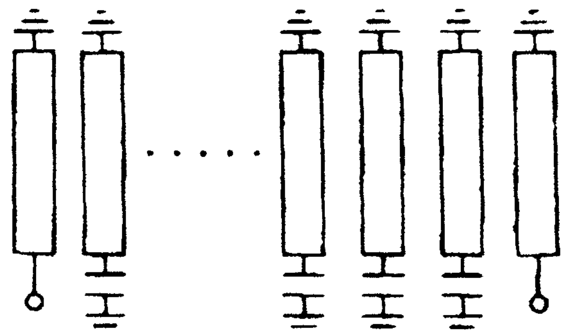
Many examples for a variety of mixed lumped and distributed resonators are given in [8].

Narrow-band filter designs for simple Chebyshev response are carried out by forming the appropriate susceptance slope parameter, and then substituting into the formulas given in [8]. Filters composed of series resonators are less common. They are simply the dual of the shunt case, and are also treated in [8].

In the case of pseudoelliptic filters, designs may be carried out by deriving a suitable low-pass prototype, followed by a low-pass to bandpass (or bandstop, or high-pass) transformation, and the same susceptance slope parameter method applied. Typically, the finite frequency poles are produced by a shunt *LC* "tank" circuit connected in series. This can be realized directly in some cases, obviously so in the case of lumped filters, and by employing a coupling capacitor in the case of a combline resonator, e.g., as in [9]. However, it is more usual to produce finite frequency poles by transformation of the circuit into one that requires cross coupling between nonadjacent resonators, as described in [10], [11], and [39].



(a)



(b)

Fig. 2. Combline filter. (a) Opposite-sided transformer coupling. (b) Same-sided transformer coupling.

A. Combline Filters

These are the most widely used types of coaxial filters, at least for frequencies below 10 GHz or so, since capacitive end-loading of the resonators gives a useful size reduction compared with those that are based on a quarter-wave resonance. They consist of an array of parallel resonators that are short circuited at one end with a loading capacitor at the other end. The resonators are oriented so that the short circuits are all on one side of the filter, and the capacitors all at the other side. A narrow bandwidth theory was published initially in a *Stanford Research Institute Technical Report* and appeared in the open literature in 1963 [12], with republication in textbook form in [8, Sec. 8-13]. A definitive exact theory valid for any bandwidth appeared in 1971 [7].

The theory given in [12] requires external couplings consisting of an "opposite sided" transformers, as indicated in Fig. 2(a). Here, redundant unit elements are introduced at the input and output ports in order to give a realizable internal impedance level, and network transformations applied to give self and mutual admittance of the transformers, which give practical physical dimensions. However, such transformers are now infrequently used in the industry since they are somewhat inconvenient mechanically, and also increase the length of the filter. It is more usual to employ direct tapping of the end resonators, as described, for example, in [13]. The main problem with this form of coupling is that it is unsuitable when the tap point is close to the shorted end of the resonator, as with narrow bandwidths at high frequencies, since it is then difficult to control. A good alternative then is to use some form of magnetic loop coupling, which may be adjusted externally.

Another external coupling technique that is often very useful is to use "same-sided transformers" [14], which look like extra resonators at each end of the filter, as shown in Fig. 2(b), but actually have the role of impedance transformers, and do not contribute to the insertion-loss function of the filter. The filter is lengthened compared with direct tapping, but this may not always be a problem, especially at high frequencies where the dimensions are small, and gives a robust coupling technique. Such filters also have a wide tuning range with no adjustments required to the couplings and rather little change in bandwidth [14].

A drawback of combline filters lies in the asymmetry of the insertion loss, which is much weaker on the low-frequency side, especially for broad bandwidths. It is possible sometimes to introduce an attenuation pole on the low side of the passband, but this may be difficult with broad bandwidth filters due to dispersion effects—the cross-coupling capacitance has a linear frequency dependence—and the couplings between resonators are large, making cross coupling between nonadjacent resonators mechanically difficult. Other types of distributed filters may then be preferred.

When the ground-plane spacing becomes an appreciable fraction of a wavelength at mid-band, waveguide modes become the predominant form of coupling, and the filter can no longer be considered as a pure TEM line structure, but rather is an evanescent waveguide filter (see [15] and Section III-H of this paper). This type of filter has a very good unloaded Q, and is widely used in telecommunications base-stations, especially in pseudoelliptic form with cross coupling to give more optimal response functions. Here, the rejection for the simple Chebyshev combline filter has better symmetry compared with combline filters of smaller ground-plane spacing [15] since propagating waveguide filters have steeper rejection on the low side of the passband, the opposite of combline filters. The "transitional" filter [15] has a characteristic somewhere between these two extremes.

B. Interdigital Filters

These consist of parallel-coupled quarter wavelength or 90° parallel-coupled lines, which alternate between the short- and open-circuited ends, as shown in Fig. 3.

The equivalent circuit is a cascade of shunt shorted stubs spaced by transmission lines, and unlike combline filters, the lines may be a full 90° in electrical length. In practice, the lines are shorter because of inevitable fringing capacitances and, in some cases, it is advantageous to shorten the lines still further, typically to 60°, in order to obtain smaller filters with wider upper rejection bands, comparable to that obtained with combline filters [7].

Interdigital filters find most application at higher microwave frequencies above 8 GHz or so, especially for broad bandwidths. The ideal interdigital filter has characteristics having perfectly arithmetical symmetry, which can be of considerable advantage compared with combline filters. Such symmetry gives better phase and delay characteristics, and it is simpler to design linear

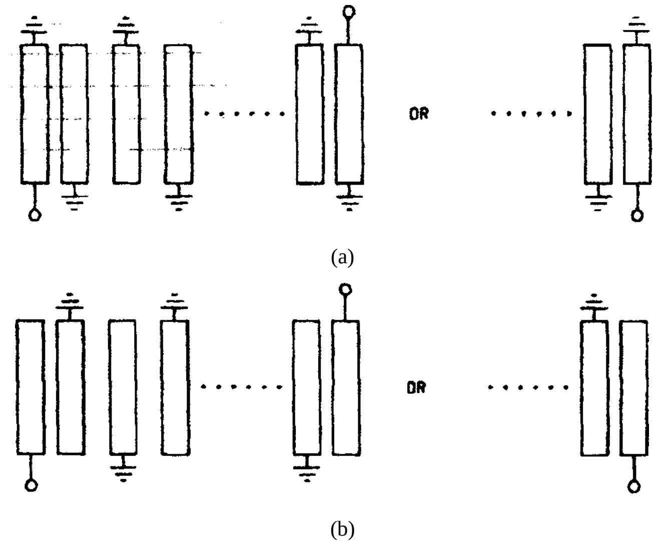


Fig. 3. Interdigital filter. (a) Short-circuited transformer coupling. (b) Open-circuited transformer coupling.

phase filters which use cross coupling between nonadjacent resonators. They have looser inter-resonator couplings compared with combline filters, i.e., the gaps between resonator bars may be larger and, hence, simpler to produce at high frequencies and wide bandwidths where dimensions become quite small. Interdigital filters have been designed for such broad bandwidths and high frequencies as 8–18, 20–28, and 28–40 GHz, with up to 23 resonators. Note that, when very broad bandwidths are specified, it is then difficult in practice (but not in theory) to achieve a return loss of better than 15 dB or so (0.15-dB ripple level) over the passband.

The design theory given by Matthaei in [8, Sec. 10.06] is based on wide-band approximations, which give results that are sufficiently accurate for most wide- and narrow-band situations. Any deviations from an ideal response may be corrected by optimization, or the exact design theories of Wenzel [7], [16] may be used.

It is important to mention that the upper stopband of full quarter-wavelength interdigital filters does not extend as high as the third harmonic as frequently stated because of waveguide moding. The first waveguide mode is the TE_{10} mode with a cutoff wavelength of twice the width of the filter housing, corresponding to just twice the fundamental frequency for the full quarter-wavelength case. A similar result holds for capacitively loaded interdigital or combline filters, where if the housing width is, say, 60°, the waveguide moding occurs at three times the mid-band frequency, i.e., prior to the first main harmonic passband, which is the fourth harmonic.

C. Parallel-Coupled, Hairpin-Line, Patch, and Ring Filters

The original type of parallel-coupled-line filters, as depicted in Fig. 4, are described in [8, Sec. 8.09]. They are realized mainly in microstrip or occasionally stripline at higher microwave frequencies since excessive length precludes their economic application in low-loss airline situations. In microstrip, it is necessary to take account of the differing phase velocities between the even and odd modes in the coupled-line regions. There are several papers on this topic, but few appear

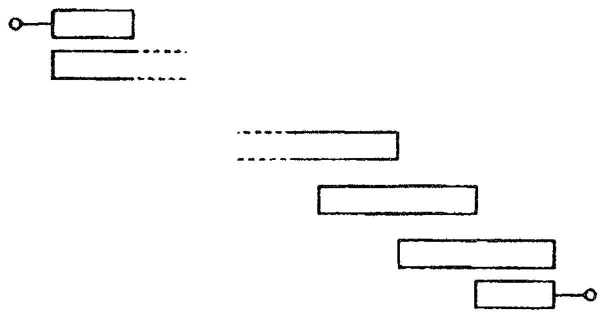


Fig. 4. Parallel-coupled-line bandpass filter.

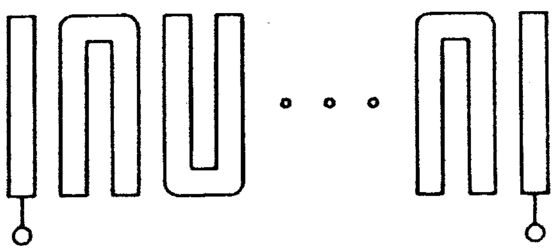


Fig. 5. Hairpin-line filter.

to give simple design procedures that can be applied without excessive work. References may be found in the indexes of past issues of this TRANSACTIONS under the subject of "microstrip filters." One of the simpler procedures is given in [17], and further discussion may be found in [18] and [19].

A very useful paper describing various types of folded parallel-coupled-line filters, known as hairpin filters, is given in [20]. One such type is shown in Fig. 5. It is interesting that [20, Fig. 5(b)] has some similarity to the very useful ring-type filter introduced by Hong and Lancaster at about the same time [21], although the former has larger capacitive loading to reduce the dimensions.

Bandstop filters also may take advantage of the strong coupling provided by parallel-coupled lines, as described in [22] and [23].

Continuing with the subject of printed circuit bandpass filters, one of the more useful developments since 1984 has been the introduction of *dual-mode* patch and ring filters [24]. The original paper describes a variety of resonator configurations and filter responses, and includes results for a superconducting filter.

This work was followed a few years later by the introduction of ring filters having capacitive gaps [25], structures which lead to enormous flexibility and variety in design. Since the fields near the capacitive gaps are mainly electric, then coupling between two such resonators with these gaps in close proximity is capacitive. The fields on other portions of rings far from the gaps is magnetic, and coupling between these portions in similarly magnetic. Hence, both positive and negative couplings are realizable with proximity coupling only, which is a considerable advantage over other types of filter structures that require capacitive probes to realize electric couplings. Another interesting development by the same authors is to use a two-layer structure with both electric and magnetic coupling apertures [26].

D. Ceramic Block and Ceramic Puck Resonators

An extremely important advance in microwave filters that has taken place since the 1984 survey [1] is the development of ceramic resonator filters of the following two main types:

- ceramic resonator or "puck" filters;
- TEM-mode coaxial cavity dielectric-resonator filters.

The theory for the first type was actually described quite early, e.g., in [27], but implementation was delayed for several years because of the lack of availability of suitable dielectric materials having good temperature stability. These were under development at a number of establishments, and became generally available in the 1980s [28]. Ceramic resonator filters of the first type have very low loss and give a substantial reduction in the size of conventional waveguide filters. They may be either single or dual mode [29], but recent preference is to use single-mode designs because of better temperature stability and somewhat better performance characteristics.

There are numerous implementation of filters of the second type and, rather than giving specific references, it is suggested that the reader should research papers by Nishikawa, Wakino, and several co-authors. Much information is available also from the manufacturers of dielectric resonators. The two main advantages of these filters are small size, suitable for use in mobile telephones, and very low production cost. They basically consist of coaxial cavities coupled either by series capacitors or by magnetic-field coupling through the dielectric. The theory is usually similar to that for air-line filters, such as combline filters in the magnetically coupled case. In some instances, dielectric may be removed from the coupling region, and an inhomogeneous structure results, requiring specialized field theory computations [30].

E. Suspended Substrate Stripline (SSS) Designs

SSS was first described by Rooney and Underkoefler [31] and has proven to be essential to the design of comparatively low-loss very broad-band filters and multiplexers. It is highly suited to the design of pseudoelliptic low- and high-pass filters. An adequate description of the techniques will be found in [1], and perhaps the only subsequent paper in the field that should be added is [19]. This is apart from the considerable body of work carried out on micromachined filters, which utilize SSS in the form of a very thin membrane, a topic covered elsewhere in this TRANSACTIONS.

F. Waveguide Filters

The basic paper here is the well-known 1957 paper by Cohn [5], which is suitable for narrow-bandwidth rectangular waveguide filters. If broader bandwidth filters are required, then more exact theories should be employed as described in [1]. A high-pass filter in waveguide is usually best designed as a broad-band bandpass filter, where the upper stopband is above the desired operating band and, in any case, is often almost nonexistent. In general, this results in a shorter and lower loss filter than one relying on the cutoff frequency of the waveguide.

Reference [1] also gives a description of dual-mode waveguide filters that are widely used in satellites and in some terrestrial systems. Triple-mode filters are mentioned also, but these are usually too complicated for other than highly specialized applications. The main problem is difficulty in tuning and, in common with dual-mode filters, an inability to tune such filters over even a narrow tuning range.

There has been much literature on filters having highly optimum (canonical) elliptic-function characteristics with the maximum number of attenuation poles, but practical considerations lead one to the preferred use of filters of the cascaded-synthesis type [10], [11], [39] rather than the canonical type. The CQ, CT, or combined CQ/CT filters may require an additional resonator, but this is almost always justified by the ease of tuning.

Waveguide low-pass filters are important, being used for harmonic rejection. Waffle-iron filters, as described in [8, Chs. 7, 15], are often used, and are suitable for very broad passbands. When the passbands are narrower, then tapered corrugated low-pass filters [32] are usually preferred since they have lower loss, smaller size, and higher power-handling capacity, as discussed further in [1]. It is important to taper both broad and narrow dimensions of the waveguide to give high rejection of higher order modes in the stopbands.

G. Coaxial Low-Pass Filters

This is another topic dealt in [8, Ch. 7], but a more satisfactory design technique is to use the very accurate mixed lumped and distributed theory given in [33], which guarantees good performance of the filter in the entire operating band, extending to the cutoff frequency. The same theory may be applied to filters designed in any medium, such as coaxial line, stripline, microstrip, SSS, or coplanar waveguide.

H. Evanescent-Mode Filters

A common assumption when designing combline or interdigital filters is that coupling between nonnearest neighbor lines can be ignored. For configurations in which the ground-plane spacing becomes larger than about 30° at the midband frequency (with 360° representing one free-space wavelength at midband), and when the bandwidth is such that the spacing between any two lines is less than about $1.5\times$ the diameter of either line (a condition most often required for broadband filters), propagation can no longer be considered as TEM. The computation of coupling between lines must include the effects of at least the cutoff dominant waveguide mode (due to the large ground plane spacing), and for bandwidths typically in excess of 40% or less than 2%, must include the effects of other cutoff modes as well. If one does include these effects, then filter bandwidth and response can be computed with great accuracy. The approach to filter design incorporating initial consideration of the cutoff waveguide modes is termed “the evanescent mode design technique.”

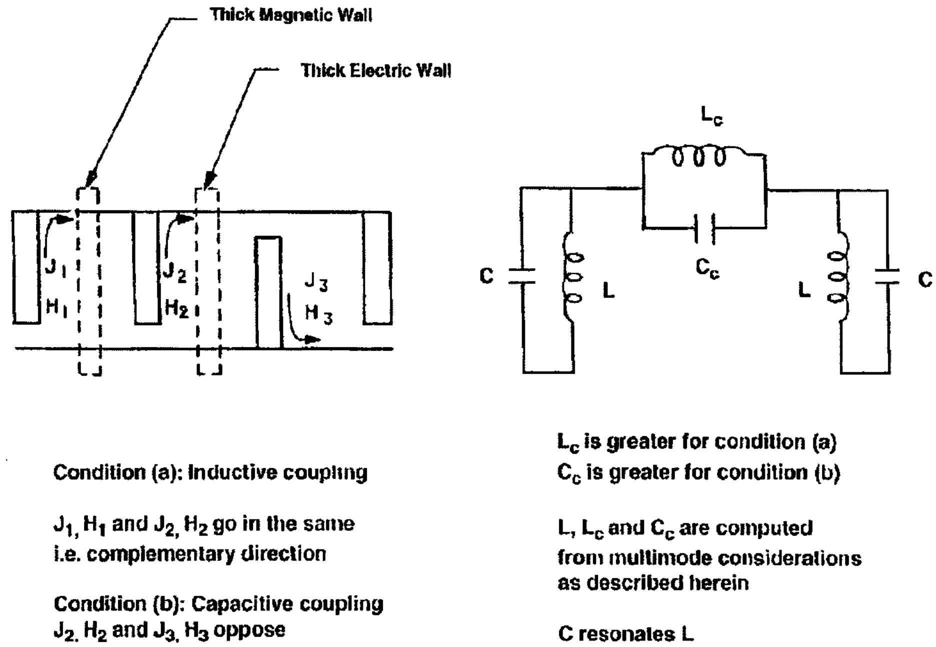
This technique uses an equivalent-circuit approach in which the circuit elements represent the natural frequency variation of below-cutoff waveguide modes. The assumption that widely spread perturbations (the rods) are coupled by below-cutoff waveguide sections is reasonable for narrow to moderate bandwidths because the spacing of the rods is at least several times the rod diameter for these cases. A wave is coupled into a below cutoff section (cutoff to the dominant mode in the particular cross section), and begins to decay asymptotically. Incidence of this “dominant” wave upon a capacitively loaded rod results in scattering of energy into other modes with different propagation constants, and consequent energy storage. Conventionally, the below cutoff section is represented as a trio of inductors (Pi or Tee). This is in accordance with the observation that, for a given length of cutoff section, the further below the cutoff frequency is the operating frequency, the higher the input impedance as measured at the input to the cutoff section. Reversal of direction, or inversion, of some or all of the resonators in an evanescent-mode bandpass structure, allows for achieving wide-bandwidth filters, with some transmission zeros located at dc and some at infinity [34]. Close spacing, as occurs when bandwidth is large, implies significant higher order mode coupling and, thus, accurate design requires consideration of these modes. The resulting filters still display the small size, low loss, and wide stopbands common to evanescent mode structures in general. Resonator inversion is equivalent to replacing the electric wall equivalent between conventional evanescent resonators with a magnetic wall, which allows for some degree of net capacitive coupling.

Fig. 6(a) illustrates the effect of obstacle direction. If the magnetic fields oppose, due to the opposing directions of current in adjacent resonators, the net coupling between obstacles tends to become capacitive. If the fields do not oppose (resonators pointing in the same direction), the net coupling is inductive. The total coupling is a composite of inductive and capacitive components. Closely spaced obstacles require inclusion of both capacitive and inductive coupling effects. Fig. 6(b) illustrates single-mode and multimode equivalent circuits for capacitive obstacles, which point in the same direction, in below-cutoff waveguide. As the obstacles are spaced closer and closer, the higher order modes couple more strongly, thus reducing the net inductive coupling between the obstacles.

Evanescent mode filters can be implemented with coaxial or propagating waveguide input and output. The filters are practical for bandwidths from about 1% to at least 70%. Inclusion of the higher order modes allows for quite exact prediction of filter bandwidth. The filters can be folded, with cross-coupling used to implement pseudoelliptic or (quasi-elliptic) elliptic designs [35]. In fact, the evanescent filters can be used to provide the actual cross-coupling network. Finally, this filter type can be combined with lumped, dielectric, or propagating waveguide resonators to form composite or hybrid structures with low-loss and wide stopbands.

IV. MULTIPLEXERS

Multiplexers are combinations of filters connected in such a way as to provide access to the passband and stopband characteristics of each filter from a common connection. Such a device permits the use of a single antenna with several receivers or transmitters, for example. The common port must display a low voltage standing-wave ratio (VSWR) and isolation must be maintained between each of the component filters. A two-channel version is called a diplexer, a three-channel-triplexer, etc. (Fig. 7). If the adjacent passbands of each channel “cross-over” at a level of about -3 dB, the device is called a “contiguous” multiplexer.



(a) (b)

Fig. 6. (a) Illustration of inductive coupling (comb and evanescent filters, condition a) and capacitive coupling, condition b, as in interdigital filters. (b) Coupling equivalent circuit when higher order modes are included.

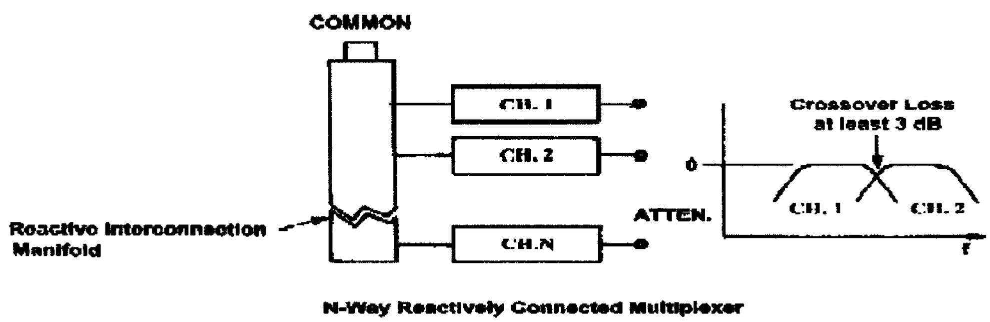


Fig. 7. N-way reactively connected multiplexer.

The frequency separation of channels is called the "guardband." Frequently a diplexer is termed a duplexer. The distinction is that a duplexer combines a transmitter and a receiver. Thus, a duplexer can be designed as a diplexer, but duplexing can be carried out in other ways, e.g., using TR devices that reflect high-power signals, but transmit low-power ones.

A multiplexer is normally used if a wide spectrum must be accessed equally and instantaneously. Conventionally, multiplexers have had the disadvantage of requiring at least 3-dB excess loss ("crossover" loss) at frequencies common to two channels. Thus, the passband characteristics for contiguous (crossing over at common -3-dB points) structures always showed an insertion-loss variation over the passband of at least 3 dB, unless power dividers are used as discussed below. Noncontiguous multiplexers are those with guardbands between adjacent passbands.

To construct any multiplexer, it is necessary to connect networks to the constituent filters such that each filter appears as an open circuit to each other filter. While this is simple for narrow-band channels, it is difficult for broad-band or contiguous filters. Normally, the filters and the multiplexing network are synthesized as a set, with computer optimization being used to simulate the results before construction begins. Some of the more common multiplexing techniques include line lengths, circulators, hybrids, and transformers.

More recently, the multiplexer filter channels have been combined using power dividers (Fig. 8).

In the case of two-way combining, conservation of energy means that the 3-dB insertion loss is still experienced . . ., but on a flat-loss basis. Although each channel is subject to the additional 3-dB loss, it is essentially constant loss over each channel and, thus, the excess passband loss variation is less than 1 dB. Excess loss is defined as that loss not attributable to the individual channel filter rolloff. This power-divider-based combining can be extended to triplexers (4.7-dB flat loss), quadruplexers (6-dB flat loss), etc. Since the loss variation is minimized, the overall insertion loss can frequently be made up using post amplifiers ("gain blocks"), displaying flat gain versus frequency. Such gain blocks are currently inexpensive and provide flat gain and low noise over wide bandwidths, thus compensating for the power-divider losses.

Filters can be multiplexed by parallel combination at both ends. For example, if two bandpass filters are diplexed at both input and output, the resulting network provides one input and

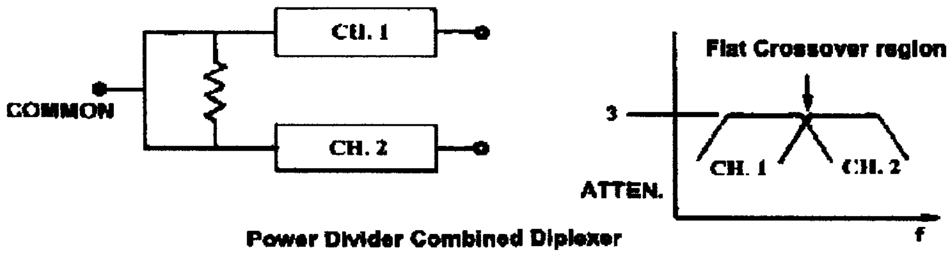


Fig. 8. Power divider combined diplexer.

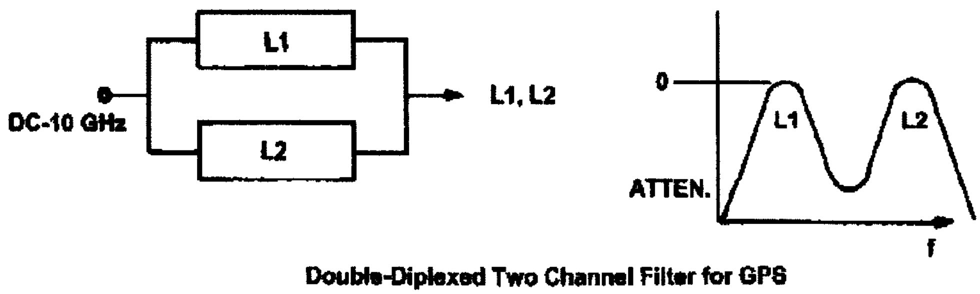


Fig. 9. Double-diplexed two-channel filter for GPS.

one output, has two isolated passbands, essentially attenuating everything else. Such assemblies are useful in systems such as the global positioning system (GPS), which have two or more operating frequencies, with the requirement for isolation between the operating channels and the adjacent cluttered regions of the spectrum (Fig. 9).

Another approach employs switched selection of filters. Hybrid combinations using multiplexers with power dividers, switches, and amplifiers are now possible. The interactions of these essentially reactive components can cause undesirable degradation of stopbands or passbands, without proper precautions.

Multiplexer theory is probably best approached by studying lumped-element multiplexers, similarly to the approach suggested for the study of stand-alone microwave filters. The constituent filters may also be quantified using susceptance slope parameters and coupling coefficients, as discussed in Section III.

The major result in the basic theory concerns the design of a contiguous low-pass/high-pass diplexer. The technique is succinctly described in [36]. Other descriptions and references will be found in [1, Sec. X], and there have been more recent developments resulting in multiplexers with many contiguous channels. Much of this work involves computer-aided optimization, frequently combined with field theory to optimize the actual dimensions. Space precludes detailed description, and the reader is referred to the index issues of this TRANSACTIONS, these days to be found most conveniently in CD ROM format. However, [37] does provide additional useful detail.

V. HIGH-TEMPERATURE SUPERCONDUCTOR (HTS) FILTERS

A. General Properties of HTS Resonator Structures

Early superconducting devices made use of material such as niobium at liquid helium temperature (4.2 K). The requirement for this very low operating temperature introduced a very severe limitation on the practical use of superconductivity. However, in 1986 so-called "high-temperature superconductivity" was discovered, which could operate in the 60–80-K range. This made possible adequate cooling with liquid nitrogen, and also the use of small practical electromechanical "cryocoolers." This has opened up a large new range of applications for superconductivity, and of special interest herein, the design of very compact low-loss microwave HTS filters.

Most HTS microwave filters in use today are of microstrip form using a thin-film HTS ground plane at the bottom of the substrate and HTS photoeteched circuitry on the top, the substrate being mounted in a normal metal housing. HTS filters are not limited to thin-film technology, however, as at least one company has developed a range of nonplanar filters using HTS thick films. It is quite common for very small HTS microstrip resonators to yield Q's of 30 000–50 000 in the 1–2-GHz range, much higher than that for relatively bulky conventional waveguide resonators in the same frequency range. Some forms of HTS resonators may yield Q's of 100 000 or more. Of course, the need for a cryocooler offsets the small size advantage of microstrip HTS filters to some extent, but in many applications, several filters may be placed in a single cryocooler. For example, numerous HTS filter systems for cellular communications use 12 filters in a single cryocooler.

The substrates used with HTS circuits must have a good crystal lattice match with the HTS in order to obtain a good bond. This plus the need for very low dielectric loss means the number of substrate materials usable are very limited. The most commonly used HTS materials are yttrium-based material YBCO and a thallium-based TBCCO. The commonly used substrate materials are lanthanum aluminate $LaAlO_3$ ($e_r = 24$), magnesium oxide, MgO ($e_r = 9.7$), and sapphire ($e_r \sim 10$) with a ceria CeO_2 buffer layer to provide a lattice match. These substrates are usually available in only relatively small sizes with diameters of only 2 or 3 in. This limits the size of HTS circuits, as does the size of the cryocooler, which, of course, is kept as small as is feasible. An important property of HTS circuits, besides their very low loss, is that when the current density in the HTS is raised to a certain level, nonlinear effects begin to appear. These cause the conductivity to decrease with increasing current density, thus lowering the resonator Q, and can cause intermodulation of signals.

B. Design Considerations for HTS Filters

The various requirements and limitations placed on HTS filters often make their design very challenging. For example, the applications for which users are willing to pay for the extra costs of HTS are typically quite difficult ones that may require up to a dozen resonators or more and a number of cross couplings. Very compact circuitry must be devised to realize such complex filters using such small substrates. Due to the crowding of the circuitry, in many designs, stray couplings beyond nearest neighbor resonators must be accounted for in addition to the desired couplings. To get the best system performance, regions of high current density in the circuitry must be minimized in order to keep the system power level at which significant nonlinear effects begin to occur as high as possible. In conventional microstrip circuits, losses due to radiation or currents induced in the cover plate are small compared to the circuit conductor losses and can be totally ignored. However, these losses can be very important in high-Q HTS circuits. For practical reasons, most HTS microstrip circuits use normal metal cover plates, and it is highly desirable to minimize any currents induced in the cover by use of circuitry that tends to keep the fields confined to the substrate or its near vicinity. For example, a microstrip meander line tends to confine the fields much better than does a line in the form of a loop or spiral.

Due to the nonlinear effects in HTS, planar HTS filters are useful mainly for low-power applications, say, of the order of a milliwatt or less. However, if relatively large nonplanar HTS structures are used, considerably higher power can be handled. For example, one manufacturer has marketed a nonplanar thick-film HTS power combiner for relatively high-power cellular applications (power of the order of tens of watts). Another approach that uses HTS thin-film technology for relatively high powers is to incorporate very low-loss dielectric material in such a way that most of the energy is carried in regions away from the HTS. Resonators have been demonstrated that use a circular puck of sapphire spaced between two parallel thin-film HTS ground planes. This keeps the energy confined with most of it in the sapphire, and the fields are relatively weak near the HTS. Q's of several hundred thousand have been realized in this way. (Having very high Q's in HTS, high-power resonators is important for additional reasons besides permitting sharp filter cutoffs. If too much energy is dissipated in the resonator, the resonator will heat up, and the superconducting operation may be lost.) This type of resonator has many spurious resonances so it would only be of interest for filter applications where the unwanted signals are quite close to the passband. Also, because of practical size limitations on the thin-film ground planes and the puck, this approach is probably of interest mainly for relatively high frequencies (such as 10 GHz) rather than, say, cellular frequencies of around 800 MHz or so.

As can be seen, the design of high-quality HTS filters can be a very challenging task, but they offer quite attractive advantages for some system applications where very high-Q circuitry is required in a relatively small space. An excellent comprehensive paper on the general subject of microwave superconductivity applications can be found elsewhere in this TRANSACTIONS [38].

VI. CONCLUSIONS

The topic of filters in general and microwave filters in particular is a vast one, and has generated at least one textbook in excess of 1000 pages [8]. In this paper, the object is to hopefully simplify the entry points for newcomers to the field, and to point out how the theory may be unified. Considerable areas of filter activity have not been covered here, such as tunable filters and the role of active filters, but these may be considered as specialized topics and somewhat less fundamental. It is shown that a sound basis for comprehension of microwave filters lies in a study of lumped-element filters, but in so doing, it is necessary to overcome the restrictions of standard textbook presentations and to delve deeper into the literature.

Another interesting aspect of filter theory concerns its application to the design of many different kinds of passive components, such as directional couplers, power dividers, and phase shifters, topics that are covered elsewhere in this TRANSACTIONS.

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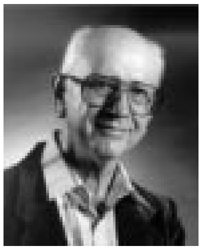

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