Ceramic IF filters for military and commercial equipment where exact interstage alignment and frequency stability are more critical than cost. Recent developments show promise of consumer applications in those cases in which cost is an extremely important factor.

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Ceramic filters offer a number of performance and compatibility advantages in solid-state consumer IF circuits; and they have recently become economically competitive with the conventional LC approach. This article surveys the different types of ceramic filters, as well as their properties and applications, including reports of some new network approaches regarding spurious suppression and application in integrated circuits.

Over the past decade, solid-state technology has changed the design of receiver circuits radically. With the impending introduction of varactor tuning, one of the few conventional components remaining—and certainly one of the most cumbersome because of size and frequency alignment—is the wound inductance in intermediate-frequency (IF) transformers.

Considerable effort has gone into finding new and more compatible methods for IF filtering. Many solutions have been attempted, including RC-active (by feedback, NIC, gyrator, phase-locked loop), electromechanical (resonant gate transistor, tunistor, magnetoostriction), digital, and sampling filtering techniques. At present none of these methods can match the performance and economy of the conventional wire-wound IF transformer at the intermediate frequencies of principal interest, namely, 455 kHz for AM radios, 4.5 MHz for television audio reception, and 10.7 MHz for FM receivers.

The characteristics of ceramic filters are such that, if applied properly, this type of filter performs better than the conventional IF transformer and is more compatible with modern receiver circuitry. Ceramic filters for high-volume consumer application in recent months have become competitive economically with the conventional approach. They produce equal or better performance than conventional devices at equal or lower overall price, and are compatible with present trends in solid-state circuitry with regard to size, reliability, and fixed tuning.

A number of manufacturers, both in the U.S. and elsewhere, are beginning to offer various types of economical ceramic filters. Some have been designed into receivers that are now, or will shortly be, on the market. For reference and comparison with familiar ground, in this article the ceramic filter response frequently will be compared with the well-known performance of tuned LC circuits.

In the following, most of the notes on performance and application relate to filters consisting of one or more individual two-electrode resonators operating at center frequencies between 0.3 and 1 MHz. This type of resonator has been the most promising building block for a true, versatile, and well-proved "economy" filter. Recently, however, various thickness-mode resonators, and especially multiresonator configurations arranged on a single ceramic wafer, are being offered for operation at higher frequencies, especially at the 4.5- and 10.7-MHz IFs. These filters are certain to have a major impact on modern IF circuitry, and although no long-term or large-scale performance data are available to date, their principles of operation and equivalent circuits will be described.

History

The basic building block of ceramic filters is the ceramic resonator. Good ceramic bandpass filters became feasible about a decade ago with the introduction of a stable, high-Q piezoelectric ceramic material that is still the basis for all ceramic filters on the market today.

Until recently, the main application of ceramic filters has been in high-performance military and commercial communication and navigation equipment, where high selectivity, stability, ruggedness, and small size are of prime importance. This type of filter is well established, with standard units ranging in fractional bandwidths from approximately 0.05 to 20 percent and 60/6-dB skirt ratios as low as 1.1. The majority lie in the frequency range between 200 and 700 kHz, but continuing efforts have stretched the range of specific filter types down to a few kilohertz and to as high as 10 MHz.

Since the inception of ceramic filters, attempts have been made to use them in place of conventional IF transformers in broadcast radios. This type of ceramic filter
was described in a number of publications. It offered the genuine advantages of fixed tuning, frequency stability, and small size. However, it never got into high-volume application, for a number of reasons. First, the filters were introduced by a single-source supplier at a time when conventional IF transformer prices were decreasing as a result of the manufacture of these transformers in the Far East. In addition, the ceramic filters lacked a dc path between the filter terminals with a resulting reduction in amplifier gain, which, until a few years ago, represented a significant economic disadvantage; and, finally, they had not yet achieved a proved long-term record of reliability.

None of these drawbacks exist now. Transistor amplifier gain is cheap. Millions of ceramic resonators have been operating in both military and commercial equipment for years without failure or deterioration. New material-resonator and network approaches, as well as automated mass-production techniques, are being applied to make the use of ceramic resonators in broad broadcast receivers and other high-volume applications economically feasible. In fact, the mass production of ceramic IF filters is inherently simpler than that of wire-wound LC circuits, and should ultimately result in reduced prices for IF circuitry.

The performance of a ceramic filter depends on the quality of the ceramic material and processing, on the geometric configuration of the ceramic resonator, and on the manner of applying the resonator(s) to the IF network. The principal considerations of this article are (1) ceramic resonator structures and their electrical equivalence; (2) single-resonator ceramic filters; (3) multiresonator ceramic filters; (4) inductorless IF networks; (5) circuit applications; and (6) ceramic filters in integrated circuits.

**Ceramic resonators**

Ceramic resonators rely on the piezoelectric effect for a direct interaction between electric and mechanical energy, and particularly for a direct relation between electrical and mechanical resonances. Since mechanical resonances are related to geometric configuration, one may envision many types of ceramic resonators with a large variety of vibrational modes and electrical resonances. Only a few of the more popular configurations in use will be discussed. Demonstrated will be a distinction between "basic resonators," consisting of one homogeneous resonating body, and "composite resonators" that combine two or more basic resonators in one unit.

A popular basic resonator with two electrodes is shown in Fig. 1. It utilizes the radial mode of vibration, i.e., it expands and contracts radially at the frequency of the electric signal applied to its electrodes, whereas the center has no radial motion. In order to minimize damping of its vibrations, the resonator is usually mounted at the center. Electrodes on opposite faces of the disk serve to intercouple electric and mechanical energy.

There are many two-electrode resonator configurations, operating in various extensional, flexural, shear, or thickness modes. They all have the electrical equivalent circuit, circuit symbol, and impedance-frequency characteristics given in Fig. 2. Electrical characteristics are generally specified by either $R$, $L$, $C$, $C_s$, of Fig. 2(A), or $Q$, $C_s$, $f_i$, $\Delta f$, where
where 

\[ f_r = \text{series resonant frequency} = \frac{1}{2\pi \sqrt{LC}} \]

\[ f_\text{a} = \text{antiresonant frequency} = f_r \sqrt{1 + C_0/C_0} \]

\[ \Delta f = f_\text{a} - f_r \]

\[ Q = \frac{2\pi Lf_r}{R} = \frac{1}{2\pi RCf_r} \]

The electrical resonator parameters may be varied over a range dependent upon the ceramic material and processing and on the resonator geometry. For example, Table I lists some typical properties of radial-mode resonators for two widely used ceramic materials.

The value of the resonator shunt capacitance \( C_0 \) depends on the dielectric constant of the material and the dimensions \( T \) and \( D' \) of Fig. 1. The frequency \( f_r \) is equal to the radial mechanical resonance frequency of the disk. For the materials listed, \( f_r \approx 254/D \), where the diameter \( D \) is measured in centimeters and \( f_r \) in kilohertz. Hence, for both very high and very low operating frequencies, \( D \) becomes impractically small or large. At present the preferred frequencies for radial resonators lie between 200 and 700 kilohertz.

The quantities \( \Delta f/f_r \) and \( C_0 \) may be adjusted by the manufacturing process to any value within the tabulated ranges. This makes the ceramic resonator a much more versatile electric circuit element than, for example, the quartz resonator, whose equivalent circuit is the same as that of Fig. 2(A) but whose \( \Delta f/f_r \) is a material constant of small and fixed value.

The stability of the filter relies on the temperature dependence of the resonator parameters. The quality factor \( Q \) remains essentially constant from room temperature down to \(-55^\circ\text{C}\), but declines toward higher temperatures. The strong positive temperature coefficient of \( C_0 \) can generally be accounted for in the filter design in such a manner that it has no significant effect on the filter performance. The frequency-temperature coefficient can be controlled by processing to be either positive or negative. Frequencies \( f_r \) and \( f_\text{a} \) track closely over the temperature range to maintain an essentially constant \( \Delta f/f_r \).

Compared with the average consumer-type IF transformer, ceramic materials and filters have a better frequency-temperature stability and quality factor \( Q \), the latter by an order of magnitude. A typical radial resonator with \( f_r = 455 \text{ kHz} \) has a diameter of 0.56 cm and a thickness of 0.038 cm. All equivalent piezoelectric resonator circuits are only valid in the vicinity of the operating frequency. For radial resonators, the equivalent circuit of Fig. 2(A) is accurate for frequencies up to approximately 1.5 \( f_r \). At higher frequencies, other resonances occur due to overtones of the radial mode and to other vibrational modes. By special resonator geometry it is possible to reduce, eliminate, or enhance a particular overtone. This method may be used to obtain higher operating frequencies without reducing \( D \) to impractically small sizes.

For operating frequencies far below and above the radial-resonator range, different resonator configurations and modes are used. For example, ceramic filters at 10.7 MHz use modes whose resonant frequency is determined by the thickness of the ceramic material between the electrodes.

By partitioning one of the electrodes of Fig. 1 into two isolated parts, one obtains a three-electrode resonator whose simplified equivalent circuit and network symbol are shown in Figs. 3(A) and 3(B), respectively. Here \( C_0 \) and \( C_0' \) are the shunt capacitances between terminal pairs 1–1' and 2–2', \( C' \) is the coupling capacitance between input and output electrodes, and the transformer ratio is:

\[ n = \sqrt{\frac{C_0'}{C_0}} \]

Variety in composite resonators is as large as that in basic resonators, but only three versions will be dis-

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### Table I. Typical properties of radial-mode resonators for two ceramic materials

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Ceramic A</th>
<th>Ceramic B</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \Delta f/f_r ), range, percent</td>
<td>1–10</td>
<td>0.2–3</td>
</tr>
<tr>
<td>( C_0 ) range, picofarads*</td>
<td>20–85</td>
<td>10–400</td>
</tr>
<tr>
<td>Relative dielectric constant</td>
<td></td>
<td></td>
</tr>
<tr>
<td>At ( 25^\circ\text{C} )</td>
<td>1150</td>
<td>500</td>
</tr>
<tr>
<td>(-55^\circ\text{C} ) to (+55^\circ\text{C} )</td>
<td>900–1400</td>
<td>450–580</td>
</tr>
<tr>
<td>(-20^\circ\text{C} ) to (+60^\circ\text{C} )</td>
<td>1000–1300</td>
<td>480–590</td>
</tr>
<tr>
<td>( Q ) value</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(-55^\circ\text{C} ) to ( 25^\circ\text{C} )</td>
<td>450</td>
<td>1400</td>
</tr>
<tr>
<td>( +25^\circ\text{C} ) to (+55^\circ\text{C} )</td>
<td>450–250</td>
<td>1400–900</td>
</tr>
<tr>
<td>Maximum total ( f ), shift, percent</td>
<td>0.2</td>
<td>0.2</td>
</tr>
<tr>
<td>Maximum ( f ), time shift, percent †</td>
<td>0.2</td>
<td>0.1</td>
</tr>
<tr>
<td>Maximum ( f ) within five years</td>
<td>0.2</td>
<td>0.2</td>
</tr>
</tbody>
</table>

* For \( f_r \approx 445 \text{ kHz} \).
† Maximum advertised value. Typical value, 0.1 percent.

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**FIGURE 4.** An example of a composite resonator obtained by bonding together two single resonators. Common center electrode is grounded.
cussed. One type consists of a ceramic transducer bonded to a metal resonator. This has the equivalent circuit of Fig. 2(A) with potentially improved $Q$ and stability at the expense of narrow limits on $C_r$ and $\Delta f$.

Another version is obtained by bonding together two single resonators and connecting them in the manner of Fig. 4. This arrangement also has the equivalent circuit of Fig. 3(A). The grounded common center electrode acts as an electric shield between the input and output and greatly reduces the value of $C'$.

A more complex and selective composite resonator is obtained by bonding two basic resonators to an intermediate coupling link that controls the amount of mechanical energy transferred from one resonator to the other. One form of this and the corresponding equivalent circuit are shown in Fig. 5.

The same principle and equivalent circuit hold for the so-called “coupled mode” or “monolithic” filter, which is becoming especially important for thickness-mode operation in the MHz range. It consists of the combination of two or more separate but acoustically intercoupled resonators on a single, thin ceramic wafer. For $n$ resonators one obtains $n$ resonant circuits—separated by shunt capacitors—in the equivalent circuit. This concept has been applied to quartz filters as well, and has been described in a number of recent publications.\footnote{11}

**Single-resonator filters**

In general, several resonators are required to shape the IF response. This may be done by arranging individual resonators at separate points in the IF amplifier or by lumping several resonators into composite filter structures.

For ease of understanding and for comparison with the LC approach, the results are related to the performance of a single-tuned LC circuit. We recall that the 3-dB bandwidth of such a circuit is $B = f_r/Q_L$—where $f_r$ and $Q_L$ are the center frequency and loaded $Q$, respectively—and that its frequency response is given by the universal resonance curve.\footnote{12} A typical value for the loaded $Q$ of 455-kHz IF transformers is $Q_L = 30$, resulting in a bandwidth of $B = 15$ kHz and the filter response shown in Fig. 10 (in color). For performance comparison, the ceramic filters will be referred to the same 3-dB bandwidth response.

The filter applications of the basic two-electrode resonator all relate to the fundamental networks, Fig. 6(A) or 7(A), where $Z$ represents the resonator impedance and $R_1$ and $R_2$ the source and load impedances, which, for our purposes, are assumed to be resistive. Figures 6(B) and 7(B) illustrate pertinent filter responses, where $f_r$ and $f_c$ are the resonator frequencies identified in Fig. 2(C). The passband response approximates that of a single-tuned LC circuit if the value of $\Delta f/f_r$ is relatively large.

For example, in the network of Fig. 6(A), the passband center coincides with the series resonant frequency $f_r$ of the $LRC$ branch of $Z$. The loaded $Q$ for series resonance is

$$Q_L = \frac{1}{4\pi\Delta f/C_{c_{\text{series}}}}$$  \hspace{1cm} (1)

where, for all practical purposes, $R_{\text{series}} = R_1 + R_2$. If

$$\frac{\Delta f}{f_r} \leq \frac{1}{Q_L}$$  \hspace{1cm} (2)

FIGURE 9. Collector-shunt coupler network (A) and base-shunt coupler circuit (B) are additional examples of two-electrode resonators used in a common-emitter amplifier stage.

FIGURE 10. Selectivity of a series coupler compared with that of a single-tuned circuit of equal 3-dB bandwidth.

FIGURE 11. Typical emitter-bypass responses for various values of $R_c$ compared with the selectivity of a single-tuned circuit having an equal 3-dB bandwidth.

then the 3-dB bandwidth is

$$B = \frac{f_a}{Q_L}$$  \hspace{1cm} (3)

The capacitance $C_r$ in Fig. 6(A) contributes the insertion loss peak at $f_a$ (due to parallel resonance of $Z$) and the declining stopband insertion loss toward higher frequencies.

The network shown in Fig. 7(A) is the dual of that illustrated by Fig. 6(A), and the response, Fig. 7(B), may be explained in an analogous manner. In particular, it can be shown that Eq. (3) holds if (2) is satisfied, and that

$$Q_L = \frac{\pi C_r R_2}{\Delta f}$$  \hspace{1cm} (4)

where, for all practical purposes,

$$R_p = \frac{R_1 R_2}{R_1 + R_2}$$  \hspace{1cm} (5)

Figures 8 and 9 show four different examples of incorporating two-electrode resonators in a common-emitter amplifier stage. With proper design, the circuits illustrated in Figs. 8 and 9 produce the responses shown by Figs. 6(B) and 7(B) respectively.

The series coupler [Fig. 8(A)] conforms directly to the previous analysis of the network shown in Fig. 6(A), except that $R_c$ now corresponds to the combined resistance of the transistor base input and the associated biasing.
circuit. Figure 10 compares the actual selectivity of a series coupler with that of a single-tuned circuit designed for the same 3-dB bandwidth of $B = 15$ kHz.

In the emitter-bypass circuit [Fig. 8(B)], the usual emitter-bypass capacitor has been replaced by a resonator. The additional resistor $R'$ in the emitter branch may or may not be necessary to adjust for the desired bandwidth. If $R_e$ is the equivalent resistance of the signal source as seen from the transistor base, one obtains for $R_{ser}$ of Eq. (1)

$$R_{ser} \approx R' + \frac{R_e + h_{ie}}{h_{fe}}$$

(6)

where $h_{ie}$ and $h_{fe}$ are the hybrid transistor parameters. For the voltage transfer function one obtains

$$\frac{e_2}{e_1} \approx \frac{R_{ser} + R_e Z}{R_e + Z}$$

(7)

Figure 11 shows typical emitter-bypass responses (for various values of $R_e$) compared with the single-tuned-circuit selectivity.

The performance of the collector shunt coupler [Fig. 9(A)] can be derived from Eq. (7). Since the emitter resistor is now bypassed by a capacitor rather than a resonator,

$$\frac{e_2}{e_1} \approx \frac{R_d + h_{ie}}{Z h_{fe}}$$

(8)

is obtained, where $Z_c$ is the total load impedance (including $R_c$ and $Z$). The frequency response is proportional to $1/Z_c$. The loaded $Q$ is given by (4) if $R_e$ is substituted for $R_l$ in (5).

If the base shunt coupler, shown in Fig. 9(B), is driven from a constant-current source (this may not be a practical assumption for conventional circuits, but can be achieved by special means), its performance is also described by Eqs. (8) and (4), provided that $r_2$ in Eq. (5) is replaced by the combined resistance of the transistor base input and the associated biasing network.

Figure 12 shows a shunt-coupler response in comparison with that of a single-tuned circuit.

In summary, the applications of Figs. 8 and 9 result in asymmetric passband responses that individually are not well suited for IF filtering. However, networks of the type shown in Fig. 8 can be combined with those of the type shown in Fig. 9 (either in one amplifier stage or in consecutive stages) such that the individual curves superimpose and yield a symmetric response. For example, Fig. 13 shows the superposition of Figs. 10 and 12; it compares the combination of a series and a shunt coupler [curve (a)] with that of two single-tuned circuits [curve (b)]. Although curve (a) has steeper passband skirts, curve (b) offers better stopband rejection. Neither curve (a) nor (b) is optimal, but a combination of the two would produce a desirable IF response.

Next, consider the basic three-electrode resonator having the equivalent circuit of Fig. 3(A). Generally, $C'$ is negligibly small. If, furthermore, $n = 1$, the network may be redrawn as in Fig. 14, which also includes the generator and load resistances $R_1$ and $R_2$.

Without going into a detailed analysis, one observes that the network corresponds essentially to a single-tuned series circuit whose selectivity can be evaluated in terms of its loaded $Q$.\(^9\)\(^1\)\(^2\) Note, however, that the center frequency $f_c$—in contrast to the conventional IF transformer—is critically affected by the loading at both the input and output of the filter. If, for example,

$$R_1 \ll \frac{1}{2\pi f_c C_{eq}}$$

and if
\[ R_t \ll \frac{1}{2\pi f_s C_a} \quad (10) \]

then

\[ f_s \approx \frac{1}{\sqrt{L/C}} \quad (11) \]

whereas if

\[ R_t \gg \frac{1}{2\pi f_s C_a} \quad (12) \]

then

\[ f_s \approx \frac{1}{\sqrt{L C_a/(C + C_a)}} \quad (13) \]

If \( n \neq 1 \), the basic filter response is not changed except for an input/output impedance transformation.

The preceding analyses of the two- and three-electrode circuits are based on several simplifying assumptions; in particular, the termination impedances are assumed to be purely resistive. If significant and known reactive components are present, they can be accounted for by adjusting the resonator parameters.

Another general observation regards the lack of a dc path in ceramic resonators and the resulting gain reduction. In the circuits shown in Figs. 8 and 9, for example, the collector direct current is supplied through \( R_c \). For minimum dc voltage drop, \( R_c \) should be small; for maximum signal output, however, \( R_c \) should be of the order of the collector impedance, which may be several hundred kilohms. This dilemma occurs with both single- and multiresonator ceramic filters, and may be solved by a compromise for the value of \( R_c \) and the achievable gain. An alternative is the substitution of \( R_c \) by a choke or IF transformer, but this method is becoming less attractive because of the low cost of transistor gain.

**Multiresonator ceramic filters**

Most ceramic filters in consumer applications consist of multiresonator structures. The trend toward integrated circuits and toward the reduction of the number of leads to the IC package will probably further enhance the lumped-filter concept.

A very important aspect of filters is their sensitivity to tolerances and stability of their network elements. This is especially true for ceramic filters, whose center frequencies cannot be adjusted like those of IF transformers. Obviously, less critical tolerances mean better manufacturing yield and efficiency. For this reason, an economical filter should not only have a minimum number of elements, but its resonators should be as simple as possible. Nevertheless, some multiresonator filters are using composite or three-electrode resonators, and they offer advantages in some applications. However, it seems that with the exception of coupled-mode filters in the MHz range, the simple two-electrode resonator is the most economical and versatile building block for ceramic filters.

The multiresonator filters subsequently described are all directly or indirectly related to ladder or lattice configurations. These are three- or four-terminal structures in which the 3-dB bandwidth is no longer a simple function of center frequency and loaded Q, and cannot be varied in a predictable manner by merely adjusting the termination impedances. Instead, the filter is designed for the desired performance and must be matched to the predetermined termination impedances.

It is worth noting that for application in integrated circuits, \( LC \) filters are also being converted from the separate IF transformer approach to lumped structures. Because of lower \( Q \) values, their performance with regard to insertion loss and steepness of filter skirts cannot match that of ceramic filters.

Composite filters using three-electrode resonators have been described. They are essentially ladder structures consisting of cascades of resonators coupled either directly or by series or shunt capacitors. Figure 15 shows a two-resonator version.

If the interelectrode capacitance \( C' \) of Fig. 3(A), is negligible, the circuit corresponds to two interconnected single-tuned circuits and consequently produces the selectivity of a double-tuned \( LC \) circuit. If \( C' \) is not negligible, insertion-loss peaks appear at finite frequencies in the upper or lower stopband. Beyond these peaks, the insertion loss declines, and rises to peaks again at zero and infinite frequency.

**Ladder filters**

To understand the characteristics of ceramic ladder filters, consider the basic two-resonator ladder (also referred to as L section) of Fig. 16, which includes the load \( R_t \) and the generator with its internal resistance \( R_s \). Let the series and parallel resonant frequencies of resonators \( Z_1 \) and \( Z_2 \) be \( f_{r1}, f_{r2} \), and \( f_{s1}, f_{s2} \), respectively.

If, for example, \( f_{r1} = f_{r2} = f_s \), one obtains maximum signal transmission to the load at \( f = f_s \). At \( f_s \), which is smaller than \( f_0 \), and at \( f_a \), which is larger than \( f_s \), one obtains minimum signal transmission (corresponding to insertion-loss peaks). Because of the limits on \( f_s \), these peaks will be in the vicinity of \( f_s \) and assure good passband selectivity. At frequencies below \( f_s \) and above \( f_a \), both resonators act like a capacitive voltage divider. The larger the ratio \( C_a/C_s \), the larger is the insertion loss in the stopband.

Figure 17 shows the response for given resonator frequencies \( f_s \) and \( f_r \) and for \( C_a/C_s \) ratios of 100, 30, and 10 [curves (a), (b), and (c), respectively]. The stopband rejection is relatively constant over a large frequency range and rises to peaks again at very low and very high frequencies.

Ceramic ladders with \( n \) series resonators and \( m \) shunt resonators may have a response with up to \( n \) and \( m \) separate insertion-loss peaks in the lower and upper stopband, respectively.

The main advantage of ceramic ladders is their relative insensitivity to element variations. For a quantitative illustration, the nominal parameters of the filter with the response shown in Fig. 17(B) were changed as follows: (1) \( R_1 \) ± 50 percent; (2) \( R_t \) ± 50 percent; (3) \( R_t \) ± 50 percent and \( R_2 \) ± 50 percent; (4) \( R_t \) ± 50 percent and \( R_2 \) ± 50 percent; (5) \( f_{r1} \) ± 0.5 percent; (6) \( f_{r1} \) ± 0.5 percent; (7) \( f_s \) ± 0.5 percent; (8) \( f_s \) ± 0.5 percent; and (9) \( Q \) ± 50 percent.

In cases (1) to (6), the passband response between the 20-dB limits is essentially unchanged, and the only significant variation occurs for cases (1) and (2), with changes of approximately ±2 dB in stopband insertion loss and for cases (3) and (6) with a 0.5 percent shift of the insertion-loss peaks at \( f_{r1} \) or \( f_{r2} \), of Fig. 17. Cases
(7) and (8) result in approximately ±0.2 percent shift in center frequency (affecting only the response between approximately the 15-dB levels) and in case (9) the insertion loss at $f_0$ doubles and leaves the remaining response unchanged. Variations of the shunt capacitances $C_m$ and $C_n$ have the same effect as changing the termination impedances. For example, an increase in both $C_n$ and $C_m$ by 50 percent corresponds to changing both $R_1$ and $R_2$ by $-50\%$.

As mentioned before, the ceramic frequency-temperature coefficient can be controlled by processing to be positive or negative. In ladder filters, the series resonators normally have positive and the shunt resonators negative temperature dependence. This tends to improve the center-frequency stability. Further, it counteracts the band-narrowing effect of the declining $Q$ toward higher temperatures.

The main drawback of ladders with only a few resonators is the low stopband rejection. Presently, the practical upper limit on the $C_m/C_n$ ratio of Fig. 16 is about 40. For a two-resonator ladder, this corresponds to a stopband rejection of about 28 dB. There are a number of ways of alleviating this problem.

1. Increase the number of resonators. This depends on economic feasibility. For a $C_m/C_n$ ratio of 40, each additional resonator would add about 14 dB to the stopband rejection. If the number of resonators is odd, the passband response will be slightly asymmetric. If, for example, the network of Fig. 16 were completed to a T section by adding another series resonator $Z_3$, the insertion-loss peak at $f_0$, would be higher (due to parallel resonance in the two series resonators) than the insertion-loss peak at $f_m$.

2. Use additional capacitors. For example, by connecting a capacitor $C = C_n$, in parallel to $Z_2$ of Fig. 16, the parallel capacitance of the shunt resonator could effectively be doubled, thereby raising the stopband rejection. However, this procedure has limits because it reduces the effective $\Delta f$ of the shunt resonator. Other resonator-capacitor ladders that are not limited in this way will be discussed later.

3. Use one or more LC circuits in conjunction with ceramic ladders. This can produce a very efficient filter since it combines the steep-skip characteristics of the ceramic ladder with the high stopband rejection of the LC circuit.

Figure 18 shows a simple version of such a network. Curve (A) of Fig. 19 is its response when designed for the same capacitance ratio, $C_m/C_n$, and the same frequencies $f_0$, and $f_m$, used to obtain the response shown in Fig. 17(B) of the transformerless L section. The comparison shows that the introduction of the transformer raised the minimum stopband rejection by about 16 dB. Curve (B) in Fig. 19 shows, for reference, the response of a single-tuned circuit with the same 3-dB bandwidth as curve (A).

In terms of interchangeability in conventional IF circuitry, the network of Fig. 19 offers several advantages, including dc continuity at the input terminals, flexible impedance transformation, and suppression of spurious resonances.

**Lattice filters**

Figure 20(A) shows a full symmetric lattice network, Fig. 20(B) its half (or hybrid) lattice equivalent whose
The main drawback of the lattice is its sensitivity to network parameter variations. The networks of both Fig. 20(A) and 20(B) are bridge circuits. Their performance depends on the balance of the impedances $Z_1$ and $Z_2$ and on the symmetry of the transformer. As a result, the filter response is especially sensitive to variations in the transformer balance or the capacitance ratio $C_0/C_{in}$. The latter is illustrated in Fig. 24, where curve (A) is identical to the class A response of Fig. 23 and curves (B) to (E) were obtained from curve (A) by changing $C_0/C_{in}$ from 0.9 to 0.96, 1, 1.04, and 1.1, respectively [this presumes $f_0 > f_1$, for the resonant frequencies of resonators $Z_1$ and $Z_2$ of Fig. 20(B)].

The results of Fig. 24 could be duplicated by maintaining the nominal $C_0/C_{in}$ ratio and upsetting the balance of the center-tapped transformer of Fig. 20(B). Since the ratio $C_0/C_{in}$ is critical, the temperature dependence of the filter does not suffer as long as $C_{in}$ and $C_0$ have the same temperature coefficient.

Figure 24 demonstrates that even if the transformer is perfectly balanced, changes of $\pm 2$ percent in the branch impedances produce response variations that are normally unacceptable in IF circuitry. Generally, the branch resonators will have to be sorted and matched in pairs. This characteristic represents a significant economic handicap.

Another limitation of the lattice in its conventional form, Fig. 20(B), is the requirement for a wire-wound and balanced inductance. Several ways of using active phase inverters instead of transformers have been suggested. This type of approach is especially attractive for integrated-circuit applications.

**Inductorless IF circuits**

The strong modern trend toward eliminating wound inductors from discrete, and especially integrated, IF circuitry was discussed previously; also mentioned were the reasons why, in some cases, a combination of $LC$ circuits with ceramic filters may be desirable or (as in lattice networks) necessary. However, the overriding reason for using $LC$-ceramic combinations is the need for the suppression of spurious responses.

Every resonating body has harmonic or nonharmonic overtones of its desired and undesired modes of vibration. For example, the resonator of Fig. 1 has radial-mode overtones around $k_1$ times the fundamental frequency $f_1$, where $k_1 = 2.5, 4, 5.5,$ and 7, respectively, for the first to fourth radial overtones. Figure 25 shows the impedance vs. frequency response of a 430-KHz radial-mode resonator with unsuppressed overtones.

Further, the same resonator has thickness-mode responses. The fundamental thickness-mode frequency depends on the resonator and electrode geometry. For the materials of Table I this value is about $f_t = 254/T$, where $f_t$ is measured in kilohertz and the thickness $T$ is measured in centimeters. Figure 25 refers to a resonator with $T = 0.0254$ cm whose thickness resonance lies beyond the plotted frequency range.

The undesired resonances are called spurious responses. To account for them, the approximate equivalent circuit shown in Fig. 2 should be expanded to that illustrated by Fig. 26.

Due to spurious responses, ceramic filters have spurious passbands. With radial resonators operating at the fundamental radial frequency $f_1$ and with $D > T$, all
FIGURE 20. Full symmetric lattice network (A), with its half-lattice (or hybrid-lattice) equivalent (B) whose transformer is ideal.

FIGURE 21. Response of the hybrid lattice circuit with ideal (black curve) and intentionally nonideal (colored curve) transformers.

FIGURE 22. Schematic equivalent of hybrid lattice network with ideal transformer using a resonator in one branch and a capacitor in the other.

FIGURE 23. Typical responses for two-resonator hybrid lattice filters, using ideal (black curve) and nonideal (colored curve) transformers at identical 3-dB bandwidths.

FIGURE 24. Response of a two-resonator hybrid lattice filter with ideal transformer, compared with responses derived from the curve by varying $C_v/C_m$ from (A) 0.9 (original value) with (B) 0.96, (C) 1.0, (D) 1.04, and (E) 1.10.

FIGURE 25. Relative admittance vs. frequency response of a 430-kHz radial-mode resonator with unsuppressed overtones.
spurious passbands lie at frequencies \( f > f_r \). Figure 27 shows the wide-spectrum response of a commercial ceramic ladder filter. Note the spurious peak at \( f = 2.1 \text{ MHz} \), which is due to thickness resonance of the series resonators whose \( T \) is 0.12 cm.

There are a number of ways of dealing with responses due to undesired vibrational modes. We mentioned, for example, that radial-mode resonators are spring-mounted at their node of radial vibration. This mechanically dampens the thickness vibrations. Further, by making \( T \ll D \), the thickness responses can be located at frequencies far removed from the desired passband.

Until recently, little had been done about the overtones of the desired vibrational mode. The reason was that a single-tuned circuit with a typical loaded \( Q \) of 35 and tuned to the center frequency \( f_r \) of a ceramic bandpass filter contributes almost 40-dB stopband rejection at the frequency \( f = 2.5 f_r \) of the first radial overtone. In conjunction with the filter shown in Fig. 27, for example, this simple \( LC \) circuit would be sufficient to suppress all spurious responses to about 60 dB below the passband level.

Spurious responses within the individual resonator may be suppressed by optimizing both the processing and the geometry of the resonator. However, some of these approaches are not compatible with economy. For instance, one known method completely eliminates all radial overtones, but results in a more costly resonator and a bulkier, less rugged package. On the other hand, complete elimination of one radial overtone, or partial suppression of two neighboring radial overtones, may be achieved by controlling the resonator geometry without sacrificing economy, size, or ruggedness.

Whatever the method, not all spurious responses within the resonator can be eliminated. However, the remaining ones can be confined to frequencies \( f \gg f_r \) where both their amplitude and the amplifier gain are reduced.

Additional spurious suppression may be obtained by proper design of the filter network. As was mentioned previously, efficient spurious suppression within the resonator requires a specific resonator geometry, in particular, a limited range for the ratios \( D/T \) and \( D/D' \) of the device shown in Fig. 1, or a limited range for the resonator impedances. This condition can easily be achieved in lattice filters, but it excludes the previously described ladder shown in Fig. 16 with its extreme \( C_{o1}/C_{o2} \) ratio.

Note that this extreme capacitance ratio was needed to raise the stopband rejection. The same purpose can be served by combining resonators of equal or approximately equal impedance levels (i.e., \( C_{o1}/C_{o2} = 1 \)) with capacitors in a ladder network. Figures 28(A), (B), and (C) show a few examples of basic sections, which may also be cascaded to obtain more selective filters.

Networks illustrated by Fig. 28(B) and (C) are especially interesting for "economy" filters since they may be

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**FIGURE 26.** An expansion of the approximate equivalent circuit shown in Fig. 2. This circuit explains undesired, or spurious, responses.

**FIGURE 27.** Wide-spectrum response of a commercial ceramic-ladder filter. Spurious peak at 2.1 MHz is due to thickness resonance (\( T = 0.092 \) cm).

**FIGURE 28.** Three examples of basic resonator-capacitor circuits.
designed to employ identical resonators \( Z_1 = Z_2 = Z \) in both the series and shunt branches and allowed to cover a continuous range of bandwidths and stopband rejections by merely changing the capacitors. Each of the response curves in Fig. 29 corresponds to a different set of capacitors \( C_1 \) to \( C_4 \) and an identical resonator \( Z \). For a quantitative example assume a shunt capacitance of 400 pF for resonator \( Z \). Then one obtains the response shown in Fig. 29(B) by adjusting capacitors \( C_1 \) to \( C_4 \) to the values 220, 450, 1300, 800, and 750 pF, respectively. The resulting network has an input impedance of 2.1 kilohms and an output impedance of 0.36 kilohm.

Aside from the spurious suppression within the resonators, the networks shown in Fig. 28(B) and (C) contribute additional suppression due to capacitive voltage division. In conjunction with resonators having partially suppressed first and second radial overtones, this type of filter has no spurious passband up to the thickness-mode responses, which appear around 7 MHz and are partially suppressed. If necessary, their effect may be further reduced or eliminated by low-pass filtering in the associated active circuitry.

**Circuit applications**

So far, application work has lagged, in that it has either concentrated on customer applications with specific constraints (for example, direct exchangeability with conventional IF transformers) or lacked sufficient collaboration and communication between the circuit designer and the filter designer.

A few applications of single two-electrode resonators were discussed previously, and a number of modifications and refinements have been suggested. For example, Fig. 11 indicates that in order to make full use of the resonator selectivity, the resistor \( R \) in the emitter-bypass configuration shown in Fig. 8(B) should be as large as possible. This may be done by replacing \( R \) by the constant-current-source emitter-collector path of a second transistor to provide an adjustable, and potentially very large, resistance in parallel to the resonator. The use of three-electrode resonators for selective and impedance transforming intercoupling of IF stages has been described previously. The basic circuit is shown in Fig. 30 and is principally applicable to ceramic filters. As explained in conjunction with the circuits shown in Figs. 8 and 9, the compromise in choosing the value for \( R \) reduces the gain of the stage unless the collector direct current is supplied through a choke or IF transformer.

The terminal impedances of ceramic filters normally range from 0.5 to 5 kilohms. The filter output is well adaptable to the input of common-emitter stages. Since the transistor's collector impedance generally is rather high, proper match at the filter input may be obtained if \( R \) equals the filter input impedance.

The filter response is relatively insensitive to impedance mismatch, as described earlier for ladder filters. However, if large circuit impedance variations—due to strong AVC, for example—are anticipated, they may have to be counteracted by special means, such as resistive divider networks.

Figures 31 and 32 illustrate methods of applying the ladder filter shown in Fig. 18, or lattice filters, to common-emitter circuits. The presence of the \( LC \) circuit at the filter input facilitates direct replacement in conventional IF transformer networks.

Ceramic filters offer some features that are not easily

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**FIGURE 29.** A set of response curves derived for the network shown in Fig. 28(C).

**FIGURE 30.** An example of applying ceramic filters to transistor IF circuits.

**FIGURE 31.** Two examples of applying ladder filters to IF circuits.

**FIGURE 32.** Application of a ceramic lattice filter to an IF circuit.
obtained in the conventional (LC-coupled stages) manner. For example, the response of ceramic ladders suggests a way to eliminate adjacent-channel interference by placing the insertion-loss peaks at the adjacent-channel frequencies. Some high-fidelity applications specify 60-dB adjacent-channel (at 455 ± 10 kHz) rejection and a 6-dB bandwidth of 8 kHz. Four double-tuned circuits would be needed to match these requirements, which could also be satisfied with a four-resonator ladder filter of the type shown in Fig. 18.

Typical United States automobile-radio IF filters have two double-tuned transformers, and their selectivity may be duplicated by a number of ceramic filter configurations. Most circuits devised so far maintain a single-tuned LC circuit following the converter. The remaining selectivity can be provided by a filter of the type illustrated in Fig. 18 by a T or pi section of two-electrode resonators, or by three two-electrode resonators, combining an L section similar to that shown in Fig. 16 with one of the approaches shown in Figs. 8 and 9. Incidentally, ceramic IF filters for auto radios are preferably centered around f_c = 445 kHz instead of the traditional 262.5 kHz because of the considerable reduction in resonator size. One may recall that the historic reason for the lower IF in automobile radios was the limit for the unloaded Q of LC components—a reason that is no longer valid for high-Q ceramics.

A similar variety of choices is available for the relaxed selectivity specifications, but stringent cost requirements, of portable-radio IF amplifiers. For example, the circuit of Fig. 31(A) can produce an IF response of the form shown in Fig. 19(A) with a minimum stopband rejection of 46 dB, and bandwidths of 8, 20, and 31 kHz at the respective attenuation levels of 6, 23, and 40 dB. Other alternatives are filters like those shown in Figs. 15, 16, or 28, used in conjunction with a single-tuned circuit. Intermediate-frequency amplifiers without LC circuits are feasible when used with spurious-free ceramic networks similar to those described earlier.

Table II lists some selectivity specifications of various economical 455-kHz filters. B_e and B_o are the bandwidths at the 6- and 40-dB level, respectively. R_w is the rejection at the adjacent channel frequencies ± 10 kHz off the center frequency, and R_s is the minimum stopband rejection. The table refers to nominal design values, which do not take into account parameter tolerances, and the filter types correspond to configurations such as (A) identical-resonator structure illustrated by Fig. 28(B); (B) identical-resonator structure obtained by cascading two sections [Fig. 28(C)]; (C) ladder structure shown in Fig. 18; (D) ladder structure similar to that shown in Fig. 18, but with four resonators; and (E) identical-resonator structure illustrated in Fig. 28(B), combined with an LC transformer at the filter input.

Ceramic filters in integrated circuits

The ultimate goal of IC designers presumably is to combine all circuit functions, including the IF section, into one monolithic structure. As was mentioned previously, there is at present no economically feasible means to this end.

Ceramic filters, in their present form, cannot be integrated monolithically. However, hybrid integrated circuits have been made using thickness-mode ceramic IF filters. Also, radial-mode resonators become small enough at higher frequencies (> 1 MHz) to be included in IC packages. They are not economically competitive with conventional IF transformers at present; however, some experimental circuits are being developed using this technique, and pilot quantities are being evaluated for high-volume commercial systems.

The reason for distinguishing between discrete-component and integrated-circuit applications is the assumption that ICs will sooner or later require inductorless filters. The requirements for this type of filter were discussed earlier, and some inductorless ladder filters were mentioned. This section concerns preliminary results for inductorless lattice filters, where differential transformers are replaced by differential amplifiers—structures that are common in ICs but would not be economical for discrete-component IF circuits.

The hybrid lattice filter shown in Fig. 20(B) is equivalent to the full lattice shown in Fig. 20(A) only if the differential transformer is ideal. The lattice shown in Fig. 20 has the transfer function

$$\frac{r(s)}{e_{in}} = \frac{R(Z_o - Z_s)}{2R^2 + R(Z_o + Z_s) + Z_oZ_s}$$

FIGURE 33. An ideal differential current amplifier.

FIGURE 34. An active filter that is equivalent to a full-lattice filter.
where the generator and load resistance is $R$.

An ideal differential current amplifier is shown in Fig. 33. Pins 1 and 2 are both assumed to be at ground potential, and the bandwidth is assumed to be infinite. The output voltage $e_o$ is given by

$$e_o = K(i_1 - i_2)$$  \hspace{1cm} (15)

If resistors and reactive elements (later to be ceramic resonators) are combined with the ideal differential current amplifier as illustrated in Fig. 34, and it is analyzed using Eq. (15) to find the transfer function, the result is

$$\frac{e_o}{e_{in}} = \frac{K(Z_a - Z_b)}{R^2 + R(Z_a + Z_b) + Z_aZ_b}$$  \hspace{1cm} (16)

Note that except for the gain constant, Eqs. (16) and (14) are exactly the same. This shows that an ideal differential current amplifier can replace the ideal differential transformer of a hybrid lattice circuit.

The ideal differential current amplifier can be approximated using the differential-input, differential-output operational amplifier shown in Fig. 35. If the operational amplifier is ideal (has infinite gain, infinite bandwidth, infinite input impedance, and zero output impedance), the circuit shown in Fig. 35 is exactly equivalent to Fig. 34. Real operational amplifiers, with high input impedance and low output impedance, will be nearly ideal if the ratio $R_i/R$ is small compared with the open-loop voltage gain.

The equivalence between the symmetrical lattice loaded at both input and output with resistance $R$ and the circuit shown in Fig. 35 holds only if the voltage source in Fig. 35 has zero internal impedance. The assumption of a zero source impedance may not be practical. In addition, the circuit shown in Fig. 35 has the disadvantage of relatively small gain, since the ratio $R_i/R$ must be small in comparison with the open-circuit voltage gain of the amplifier. The circuit illustrated in Fig. 36 does not suffer from this disadvantage. The ratio $R_i/R_{in}$ sets the voltage gain, which must be less than, or equal to, the open-circuit voltage gain.

The voltage at pin 3 is $-R_i/R_{in}e_{in}$, and that at pin 4 is $(R_i/R_{in})e_{in}$.

Analysis of the circuit of Fig. 36, assuming $R_i = R/2$, gives the transfer function

$$\frac{e_o}{e_{in}} = \frac{-R_f}{R_{in}} \frac{Z_a - Z_b}{\frac{1}{4}(R^2 + RZ_a + Z_aZ_b)}$$  \hspace{1cm} (17)

Except for gain constants, the only difference between (14) and (17) is the factor $\frac{1}{4}$ for $R^2$.

Figure 37 shows the response of an experimental filter of the type illustrated by Fig. 36, designed for a 3-dB bandwidth of 8 kHz. It corresponds approximately to the selectivity of a critically coupled double-tuned LC circuit.

Equation (17) is used in a procedure developed to design ceramic active bandpass filters with either Chebyshev or Butterworth polynomial characteristics or filters with return-loss peaks at finite frequencies. Each basic sec-

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**Figure 35.** A practical utilization of Fig. 34, using a differential-input, differential-output operational amplifier.

**Figure 36.** Operational amplifier with possibility for higher gain due to the fact that ratio $R_i/R_{in}$ sets voltage gain.

**Figure 37.** Response curve of an experimental filter similar to that shown in Fig. 36.

**Figure 38.** Filters with greater selectivity can be designed by cascading sections.
tion of Fig. 36 corresponds to a second-order response, and more selective filters are designed by cascading the sections as in Fig. 38. Further details of this design method will be made available at a later date.\footnote{11}

The previous circuits are by no means the only configurations for applying ceramic resonators in conjunction with linear integrated circuits. We have thus far only shown lattice structures with integrated circuits. Figure 39 shows several ways of using ceramic ladder filters alone and with lattice filters.

Combining a ladder and lattice with a linear IC combines the desirable properties of both. That is, the ladder can supply stable insertion-loss peaks near the passband edges, and the lattice can give increasing rejection far away from the passband. The measured response of a circuit with the structure represented by Fig. 39(D) is shown in Fig. 40.

The active lattice of Fig. 35 or of Fig. 36 may also be combined with equal-resonator ladder filters (discussed previously). Such a structure has the advantage that all resonators may be types with reduced spurious responses. The effects of spurious responses could also be reduced by using active RC filtering in the feedback loop of the differential amplifier.

REFERENCES

20. This type of filter and its design will be described in Electronic Design in the near future.
23. Ph.D. thesis to be submitted to Case Western Reserve University.