Narrow-Band, Fixed-Tuned, and Tunable Bandpass Filters With Zig–Zag Hairpin–Comb Resonators

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***Abstract*—"Hairpin–comb" filters have been previously shown to have special properties that are advantageous for the design of compact, narrow-band, and bandpass microstrip filters. Herein, a new "zig–zag" form of hairpin–comb filter is introduced, which is shown to have additional important advantages for designing compact narrow-band filters. Examples with computed responses and the measured results from high-temperature superconductor trial designs are presented. The considerable flexibility available in the design of bandpass filters of this sort is shown to be helpful in the design of tunable bandpass filters having nearly constant bandwidth and passband shape as they are tuned. Measured results for tuning over nearly an octave range are presented.**

***Index Terms*—Bandpass filters, comb filters, distributed parameter filters, hairpin resonators, microstrip, resonator filters, superconductor filters, zig–zag lines.**

I. Introduction and Goals

CURRENTLY there are numerous applications where microstrip narrow-band filters are desired, which are as of small size as possible. This is particularly true for wireless applications where high-temperature superconductor (HTS) technology is being used in order to obtain filters of small size with high resonator Q's. The filters required are often quite complex with perhaps 12 or more resonators along with some cross couplings. Yet the wafers available for HTS filters usually have a maximum size of only 2 or 3 in. Hence, means for achieving filters as small as possible while preserving high-quality performance are very desirable. Therefore, the goals of this study are to achieve the following

1) Very small and compact resonators.
2) Weak couplings between resonators (as are required for narrow-band filters) while maintaining relatively small spacings between resonators. This is important because in the case of many microstrip resonator structures, in order to achieve narrow bandwidth, quite large spacings between resonators are required.
3) Very low parasitic coupling to resonators beyond nearest neighbor resonators so that unwanted parasitic couplings can be ignored in the design process. This is important so that a field solver used in the design process will only have to analyze pairs of resonators at a time.
4) Means for maintaining nearly constant bandwidth and passband shape if it is desired that the filters be tunable over a very sizable range.

Manuscript received June 27, 2002; revised October 21, 2002. This work was supported by the Defense Advanced Research Projects Agency under Contract MDA972-00-C-0010. Paper approved for public release. Distribution unlimited.

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Digital Object Identifier 10.1109/TMTT.2003.809631

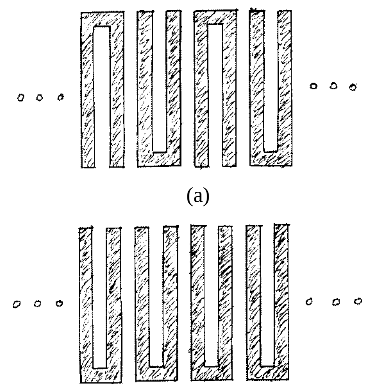

Fig. 1. (a) Conventional "hairpin-line" filter structure. (b) "Hairpin–comb" filter structure.

II. Properties of Hairpin–Comb Filter Structures

The filters discussed herein are closely related to the "hairpin–comb" filter configuration shown in Fig. 1(b). As is discussed in [1] and [2], this type of filter structure has properties that are quite useful for narrow-band filters. Fig. 1(a) shows what is commonly referred to as a "hairpin-line" filter (e.g., see [3]). Note that, in this structure, the orientations of the hairpin resonators alternate. This is done because it causes the electric and magnetic couplings to tend to add, thus resulting in maximum coupling for a given spacing between resonators. This is desirable for most applications, but is very poor for the case of narrow-band filters since then very large spacings between resonators will be required. In the case of the hairpin–comb configuration in Fig. 1(b), the resonators all have the same orientation, which causes the electric and magnetic couplings to tend to cancel. Using this structure, narrow-band microstrip filters can be realized with much smaller spacings between the resonators. Hence, this principle is incorporated in this study.

A subtle, but important phenomena that occurs in microstrip hairpin–comb structures is that a resonance effect occurs in the vicinity of the coupling region between resonators, and this creates a pole of attenuation adjacent to the passband. If the hairpin–comb structure is in a homogeneous dielectric, this pole will occur above the passband. However, in the case of conventional microstrip using a dielectric substrate, the even- and odd-mode wave velocities for pairs of coupled lines are different, and it turns out that the pole of attenuation typically

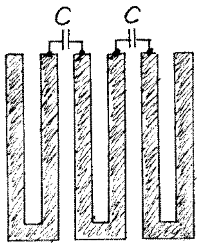

Fig. 2. Hairpin–comb structure with capacitances C added between the resonators.

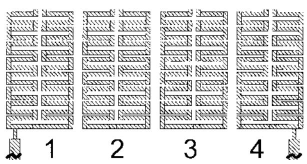

Fig. 3. Trial four-resonator "zig–zag hairpin–comb" filter.

occurs *below* the passband. However, the position of this pole of attenuation can be controlled to some extent by the addition of capacitive coupling between the open ends of adjacent resonators, as is shown in Fig. 2. It is found that as small amounts of capacitance C are *added* as shown in Fig. 2, the pole of attenuation moves *upwards* in frequency and causes the passband to be further narrowed [1], [2]. At some point, the pole will move into the passband and kill the passband completely. Adding still more capacitance, C will cause the pole to move up above the passband. This control of the pole position is a potentially useful feature of hairpin–comb structures. In the case of the zig–zag hairpin–comb structures about to be discussed, there are more degrees of freedom present, and the pole position for the case of $C = 0$ can be on either side of the passband depending on the design of the zig–zag structure.

III. Use of Zig–Zag Hairpin–Comb Resonators

The use of microstrip hairpin–comb filter structures is seen to be helpful in obtaining relatively small bandpass filters with resonators that lend themselves to quite high unloaded Q's. However, for applications where large numbers of resonators must be used on substrates of very limited size, or for filters on such substrates with a modest number of resonators, but with their passband at relatively low frequencies (say, in the 100-MHz range or lower), even more compact structures are needed. In order to meet this need, we have investigated hairpin–comb structures in which the hairpin-line structures are zig–zagged (or meandered) in order to reduce the size of the hairpin structure while at the same time presenting very limited coupling to adjacent structures.

Fig. 3 shows a four-resonator trial HTS microstrip zig–zag hairpin–comb filter structure that we have designed with the aid of Sonnet,[1] fabricated, and tested. The center frequency of the filter is roughly 2 GHz, and it uses a 0.508-mm-thick MgO $(\text{er} = 9.7)$ substrate along with TBCCO superconductor. The resonators are 3.49-mm wide and 4.8-mm high with a space of 0.45 mm between resonators 1 and 2 and between resonators 3 and 4, while 0.500 mm was used between resonators 2 and 3. The resonators could have been proportioned differently if that was desirable. Note that the large line sections in the resonators are oriented horizontally. The segments with this orientation in one resonator couple very little to the corresponding ones in the adjacent resonators. Most of the magnetic coupling between resonators comes from the short vertical sections adjacent to the gap between the resonators. Similarly, most of the electric coupling also comes from the vicinity of the vertical sections adjacent to the coupling gap. The degree of coupling between resonators is strongly influenced by the length chosen for these vertical sections, as well as by the spacing between the resonators. In this trial design, we used inductive-tap couplings at the input and output of the filter, though we could have, instead, easily used series-capacitance coupling at the upper left and upper right of the filter for coupling to the terminations.

The complex geometry of the filter in Fig. 3 would be very difficult to accurately analyze all at once using presently available computing power. Thus, it was analyzed by first analyzing resonators two at a time to get the coupling coefficients between adjacent resonators, while additional calculations were made to get the external Q's (i.e., Qe's) for the end resonators. The principles used were similar to those discussed in [4]. The expected overall passband response was then computed from the coupling coefficients and Qe's using a simplified model consisting of shunt half-wavelength open-circuited resonators coupled by ideal admittance inverters. Such simple models ignore any stray couplings beyond nearest neighbor resonators, so whether such couplings can be ignored and still give a good representation of the actual circuit response is an important question. To, at least, get some feel for the answer to this question, we tried computing the coupling between resonators 1 and 3 in Fig. 3 with resonators 2 and 4 removed. The computed coupling coefficient was $k13 = 0.0001696$, as compared to $k12 = 0.009483$ for coupling between resonators 1 and 2. We see that $k13/k12 \sim 1/56$. Thus, $k13$ appears to be sufficiently small compared to $k12$ so that it can probably be neglected. Of course, with resonator 2 in place, the value for $k13$ may be somewhat different. Similar calculations between resonators 1 and 4 with 2 and 3 removed give $k14/k12 \sim 1/285$.

The solid lines in Fig. 4 show the measured response for the HTS trial filter in Fig. 3 measured at 77 K, while the dashed lines show the response computed from the aforementioned simplified model using Qe and coupling coefficient values obtained using Sonnet. For easy comparison of responses, the computed response was transposed somewhat so as to be centered on the middle of the measured response. The measured passband ripples are larger than are the computed ripples. This indicates that the actual Qe's are higher than the design $Qe = 110.7$ value. Increasing Qe to 144 in the computer model gives a computed passband response that is virtually identical in shape to the measured response. We believe that the difference between computed and measured Qe's was, at least, largely due to asymmetric positioning of the available dielectric tuners that had metal mounts that were a little too large to be located symmetrically. Asymmetric tuning of the end resonators can throw off the effective position of the coupling taps and thus affect Qe. If it had been important to have smaller passband ripples this

[1]Sonnet Software Inc., Liverpool, NY.

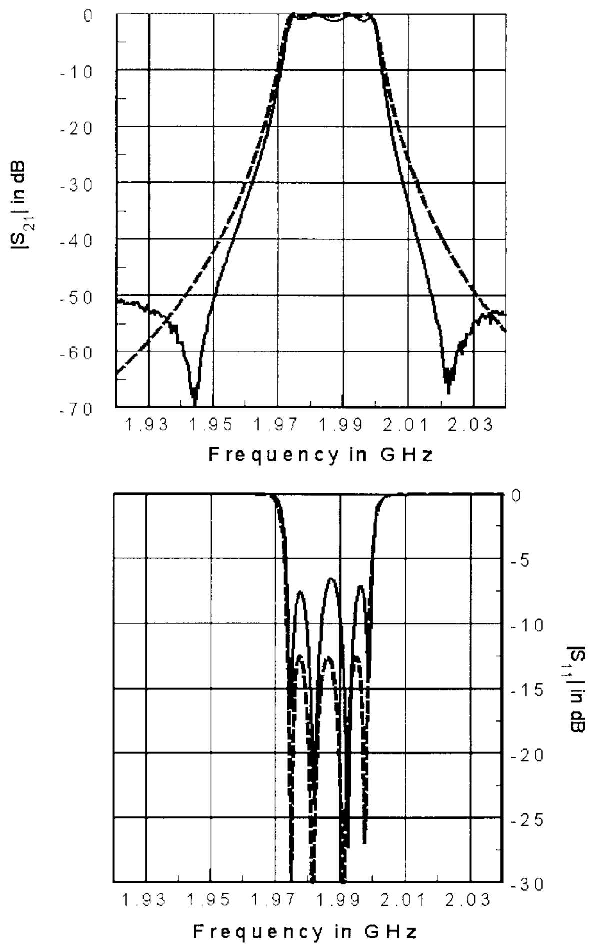

Fig. 4. Solid lines show the measured results for the filter in Fig. 3, while the dashed lines show results computed from a simple model along with coupling coefficients and external Q's obtained using Sonnet.

could easily have been achieved by simply moving the tap positions outward a bit to increase their coupling to the terminations. Additional measurements made at 77 K and roughly 2 GHz indicated that the resonators in Fig. 3 had an unloaded Q of somewhat over 40 000.

Note that the measured response in Fig. 4 shows poles of attenuation on both sides of the passband. These result from quarter-wave resonances in the two sides of the end resonators that short out the taps at frequencies somewhat above and below the resonator center frequency. These poles are attractive for many applications since they enhance the filter rates of cutoff. Note that the simplified computer model does not produce these poles. Since the bandwidth of a filter is determined predominantly by the couplings between resonators, bandwidth should be a relatively sensitive indicator of stray coupling between resonators. In Fig. 4, the measured 3-dB bandwidth is 27.3 MHz, while the 3-dB bandwidth computed from the simplified model was 29.0 MHz. The slightly smaller measured 3-dB bandwidth as compared to the computed response is to be expected as a result of the poles of attenuation in the measured response, which will tend to compress the passband somewhat. Thus, the measured response appears to be consistent with negligible coupling beyond nearest neighbor resonators in this type of structure, as was expected.

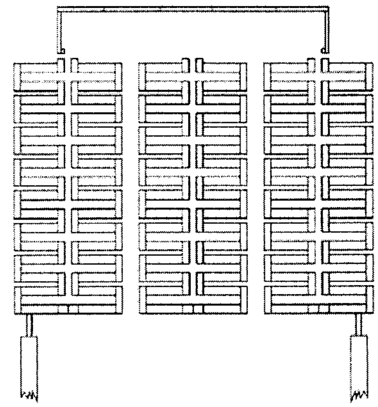

Fig. 5. Three-resonator zig–zag hairpin–comb filter with coupling added between resonators 1 and 3.

The measured response had a center frequency of $fo = 1986.6$ MHz, which is 15.3 MHz higher than was the center frequency computed for the pairs of resonators using Sonnet. This is consistent with results obtained for a previous two-resonator design [5]. In that case, cutting the 0.025-mm square cells to 0.0125-mm square reduced the error in computed fo as compared to the measured value by approximately half. This suggests that one may be able to compute a greatly improved value for fo by computing fo for a given cell size, then cutting the cell size in half and then seeing how much the center frequency is shifted. Let us call this shift Δ. Finally, the improved value for fo is obtained by shifting the value of fo obtained from the larger cell size by $2^*\Delta$. This principle appears to apply at least in this example, and has seemed to apply in some other cases where we have used Sonnet. A reviewer of this paper has suggested that this principle can be best understood by use of the "space mapping" point-of-view [6].

The above-mentioned two-resonator example [5] used inductive-tap connections to the terminations, as was done in Fig. 3. Surprisingly enough, the responses computed using Sonnet and the measured responses for that two-resonator example did not exhibit the poles of attenuation on both sides of the passband, which are to be expected from this type of connection. We conjecture that this may have been due to stray coupling between the input and output transmission lines as they were relatively closely spaced in that example. It will be noted that the two-resonator example discussed in Section IV does exhibit poles on both sides of the passband.

In many bandpass filter designs, it is desired to add carefully designed couplings beyond nearest neighbor resonators in order to introduce additional poles of attenuation beside the passband of the filter, or to alter the time-delay characteristics of the filter. We have found that such couplings are unusually simple to introduce in microstrip hairpin–comb filters. Fig. 5 shows a three-resonator zig–zag hairpin–comb filter with coupling between resonators 1 and 3 added. The coupling is introduced by a line connected to the resonators by capacitive gaps. The sign (or phase) of the coupling must be chosen correctly because one phase may have the effect of introducing poles of attenuation adjacent to the passband while the other phase may introduce poles at "complex frequencies" instead, which primarily affect

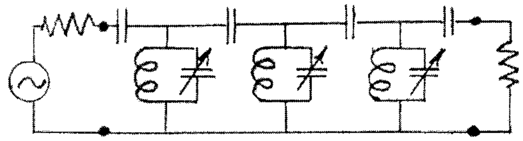

Fig. 6. Simple bandpass filter tuned by lumped variable capacitances.

the delay characteristics of the filter. In some types of filters, it can be awkward to get the desired signs for the couplings. However, in the case of the filter in Fig. 5, one can easily obtain either positive or negative coupling simply by the choice of the sides of the resonators at which the coupling connections are made.

IV. TUNABLE CONSTANT-BANDWIDTH ZIG–ZAG HAIRPIN–COMB FILTERS

Many electronically tunable filters employ electronically variable capacitors. Fig. 6 shows a filter configuration that is a convenient way to realize a filter with such tuning capacitors. Note that it uses fixed inductors and fixed coupling capacitors. In most practical applications, it is desired to maintain a constant bandwidth Δf as the filter is tuned. Unfortunately, for the filter structure in Fig. 6, this is far from the case. As can be shown using information in [4, Sec. 8.02], as the filter is tuned, the bandwidth will increase with center frequency f_o as f_o^3 instead of being constant as f_o varies. Further, in order to preserve the shape of the filter passband, the external Q's of the end resonators should *increase linearly* with f_o. However, for the filter structure in Fig. 6, the external Q's will, instead, *decrease* with f_o as $1/f_o^3$. Thus, the tunable filter structure in Fig. 6 will have very strong variations in the passband width and shape as the filter is tuned.

It is of interest to note that if one could realize a practical filter consisting of capacitively tuned *series* L–C resonators along with *inductance* couplings, the bandwidth variation would not be as severe. It can be shown that the bandwidth would vary linearly with f_o while the external Q's of the end resonators would vary as $1/f_o$ (instead of the linear variation desired for the external Q's). Thus, the bandwidth and passband shape errors incurred in this type of filter would still be objectionable, but not be as bad as are those for the filter in Fig. 6. For the case of filters having capacitive tuning and a combination of capacitive and inductive coupling between the resonators, the changes in the passband as the filter is tuned would probably lie somewhere between the two extremes discussed above. However, it is clear that, in any case, corrective measures will be required in order to design filters to maintain nearly constant bandwidth and passband shape as the filter is tuned.

Reference [4, Secs. 17.02–17.04] discuss the problem of obtaining nearly constant bandwidth in tunable coaxial and waveguide filters, while [7] treats this problem for a type of stripline combline filter. Reference [8] surveys the literature on tunable filters up to 1991 and has some discussion of the problem of obtaining constant bandwidth. However, few of the references cited actually explicitly consider that problem. Reference [9] shows a two-resonator filter structure with capacitive tuning and magnetic couplings. As discussed above, this combination of capacitive tuning and *magnetic* coupling gives a much weaker variation of bandwidth with frequency since the bandwidth tends to vary directly with the center frequency f_o instead as f_o^3 (as for the case of capacitive tuning and *capacitive* coupling). However, [9] does not suggest any means for obtaining constant bandwidth. References [10] and [11] deal with microstrip versions of the stripline filters treated in [7]. The filter treated in [12] is also closely related to those in [7], [10], and [11]. It is a microstrip comb-line filter structure with a single hairpin–comb resonator inserted in it to generate a pole of attenuation on the low side of the passband. However, in [12], no measures are taken to achieve constant bandwidth. All of the aforementioned tunable filter structures are much larger for a given operating frequency range than the structure about to be discussed herein.

The design flexibility available in zig–zag hairpin–comb filters makes them ideal candidates for achieving nearly constant bandwidth as a filter is tuned by variable capacitances. Recall that for filters tuned by simple variable capacitances, unless special measures are introduced, the bandwidth always increases as the center frequency increases (instead of remaining constant, as is usually desired). The approach used herein in order to force nearly constant bandwidth is to introduce a pole of attenuation at an appropriate location above the tuning range of the passband. As the passband is tuned up toward this pole, its influence then tends to "push down" the upper edge of the passband, thus limiting the width of the passband. When developing a constant-bandwidth tunable zig–zag hairpin–comb filter, we first started out with a two-resonator structure with resonators like those in Fig. 3. In this structure, the pole of attenuation associated with the coupling region between the resonators was above the passband and too high in frequency to give adequate limiting of the width of the passband as the center frequency was moved upwards. Thus, we wanted to move this pole to a lower frequency. As was discussed in connection with Fig. 2, adding capacitive coupling between the resonators in a microstrip hairpin–comb structure tends to move the pole of attenuation upwards. Conversely, if we add a *negative* capacitance, we can expect the frequency of the pole to move downward, which is what we want in this case. An effect similar to adding a negative capacitance can be achieved by increasing the spacing between the adjacent faces of the zigs at the top of the two resonators (which reduces the mutual capacitance between the upper regions of the two resonators) while keeping their spacings at the bottom fixed. After computer experimentation, we came up with the microstrip resonator configuration shown in Fig. 7. This worked quite well.

The shaping of the resonator zig–zags shown in Fig. 7 was effective for getting the coupling coefficient between the resonators to vary with frequency so as to give nearly constant bandwidth. However, it was still necessary to provide for forcing the external Q's of the resonators to increase approximately linearly with frequency in order for the filter passband to have an acceptable shape as the filter is tuned. In order to control the external Q of the end resonators of the filter, special circuits were added in series with the input and output of the filter, as shown in Fig. 7. These added circuits consisted of an interdigital capacitor in parallel with an inductor that was in the form of a meander line. This inductance and capacitance in parallel were added in series at the input and output of the filter so as to create a series

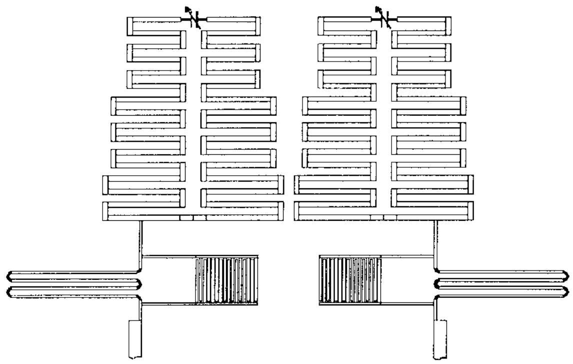

Fig. 7. Tunable two-resonator zig–zag hairpin–comb filter designed so as to maintain nearly constant bandwidth and passband shape as the filter is tuned.

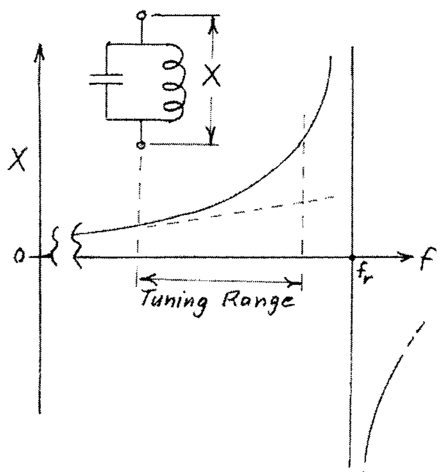

Fig. 8. Approximate reactance characteristic of the semi-lumped LC decoupling networks at the input and output of the circuit in Fig. 7.

reactance as sketched in Fig. 8. Note that as the tuning frequency is moved toward the upper end of the tuning range the reactance increases quite rapidly. This tends to increasingly decouple the end resonators from the terminations and increases the external Q of the end resonators as the frequency increases. After considerable computer experimentation, the microstrip structure in Fig. 7 was arrived at for tuning over roughly the 500–950-MHz range. In this example, it was convenient to realize the desired L and C by use of HTS circuitry. However, having a high Q is not very important for these elements, and using non-HTS lumped L's and C's external to the substrate would probably not have increased the loss very much. Of course, the filter techniques illustrated in Fig. 7 can also be implemented entirely in non-HTS form, but the loss would be considerably higher in that case because of the increased loss in highly reactive parts of the circuit.

The filter structure in Fig. 7 was fabricated using TBCCO superconductor on 0.508-mm-thick MgO. In order to tune the filter, for the present purposes, the variable capacitors shown in Fig. 7 were replaced by fairly lengthy HTS interdigital capacitors that were photoetched on the substrate along with the rest of the circuit. It was planned to successively scribe away portions of the interdigital capacitors and to thus gradually reduce the tuning capacitances in order to tune the passband to higher frequencies. With the complete interdigital capacitances

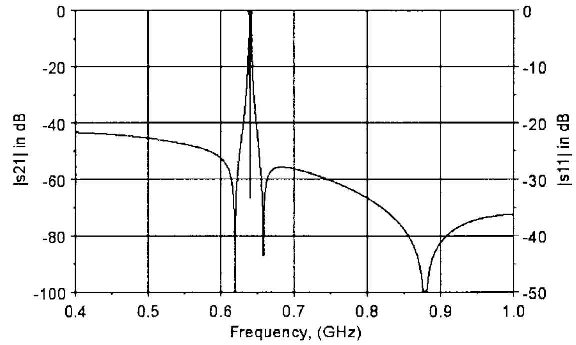

Fig. 9. Computed broad-band response for the filter in Fig. 7 when tuned to $f_o = 0.640$ GHz.

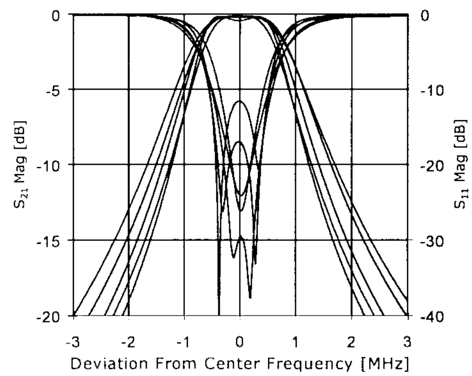

Fig. 10. Superposition of the measured passband responses for the filter in Fig. 7 when tuned to $f_o = 0.498, 0.555, 0.634, 0.754,$ and 0.948 GHz.

in place the filter tuned to a center frequency of 498 MHz. Fig. 9 shows the computed responses versus frequency when the filter is tuned to 640 MHz. Note the poles of attenuation immediately on both sides of the passband. These are due to the tap connections on the end resonators as was discussed in Section III. These poles move along with the passband as it is tuned. The pole at approximately 880 MHz is the one that is used to limit the passband width as the filter is tuned to higher frequencies. As the passband is tuned upwards from the lower end of the tuning range, this pole also moves upwards some, but not as rapidly as does the passband. Thus the pole is closer to the passband as the tuning frequency increases. At the 948-MHz upper end of our tuning range (where the interdigital tuning capacitors were entirely scribed away), the pole was still above the passband, though relatively close to it. Fig. 10 shows a superposition of the passband responses obtained at various frequencies as portions of the tuning capacitors were scribed away. Note that for practical engineering purposes, the passband shape and width remained remarkably constant over this 498–948-MHz range (a frequency ratio of 1.9). Resonator unloaded Q measurements were made at 77 K, and the Qu's were determined to be in the 85 000–90 000 range at roughly 1 GHz.

For many practical applications, one would probably like to use microelectromechanical system (MEMS) capacitors or variable capacitance diodes with filters of this sort so the filters could be tuned electronically. Filters with interdigital tuning capacitors such as we used in our initial experiments may also have practical applications where filters having a certain bandwidth are needed for a number of different center frequencies.

Many filters could be fabricated at the same time with interdigital tuning capacitors. Each circuit could then have its capacitors scribed to give a desired passband center frequency. Alternatively, instead of interdigital capacitors etched on the surface of the substrate, one could simply attach to the resonators small parallel-plate capacitors made from thin slabs of dielectric with conducting material deposited on the top and bottom surfaces.

V. Conclusions

It has been seen that the use of microstrip zig–zag hairpin–comb filter structures can give unusually compact filters for applications where very narrow bandwidths are required. The resonators in such filters can have an exceptionally small amount of coupling for a given spacing between resonators. This is advantageous for the design of narrow-band filters and, at least in many cases, can simplify the analysis of a filter since unwanted stray couplings can be neglected and still obtain an accurate design. The use of inductive tap couplings between the loads and end resonators is found to be convenient. Such taps also have the attractive property that they produce a pole of attenuation immediately above and also immediately below the passband, thus enhancing the filter attenuation characteristics. The inclusion of prescribed couplings beyond nearest neighbor resonators in hairpin–comb filters is seen to be particularly simple since the sign of the coupling is fixed by a simple choice of connections. Zig–zag hairpin–comb filters provide a relatively large degree of design flexibility in proportioning the physical structure and in placing poles beside the passband. This flexibility (along with semi-lumped LC circuits adjacent to the terminations) has been shown to be useful in the design of tunable filters so as to maintain a nearly constant bandwidth and passband shape as the filter is tuned. The trial design was seen to maintain an unusually good approximation to a constant bandwidth and passband shape when tuned over nearly an octave range. The example presented used only two resonators, but the same principles should be applicable to filters with more resonators.

As this paper is about to go to press, [13] has come to the author's attention. This excellent book discusses some "compact" and "miniaturized" filter structures in Chapter 11 that may be of interest to the reader.

Acknowledgment

The author wishes to thank B. Willemsen and E. Prophet, both of Superconductor Technologies Inc. (STI), Santa Barbara, CA, nd K. Raihn (formerly of STI), Agile Materials and Technologies Inc., Santa Barbara, CA, for valuable discussions, and also for assistance with laboratory tests and the preparation of this paper.

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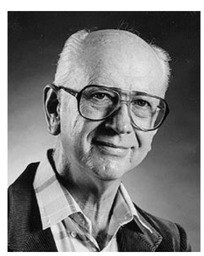

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From 1951 to 1955, he was on the faculty of the University of California at Berkeley, where he was an Assistant Professor (his specialty was network synthesis). From 1955 to 1958, he was engaged in system analysis and microwave component research with the Ramo-Wooldridge Corporation. From 1958 to 1964, he was with the Stanford Research Institute, where he was engaged in microwave device research and became Manager of the Electromagnetic Techniques Laboratory in 1962. In July 1964, he joined the Department of Electrical Engineering University of California at Santa Barbara, where he was a Professor. In July 1991, he became a Professor Emeritus. At that time, he joined the staff of Superconductor Technologies Inc., Santa Barbara, CA, on a part-time basis. He has authored or coauthored numerous papers, coauthored *Microwave Filters, Impedance-Matching Networks, and Coupling Structures* (New York: McGraw-Hill, 1964), and contributed to several other books. His current interests are in the area of design techniques for passive and active microwave HTS circuits.

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