Hairpin-Comb Filters for HTS and Other Narrow-Band Applications

George L. Matthaei, *Life Fellow, IEEE,* Neal O. Fenzi, *Member, IEEE,*
Roger J. Forse, *Member, IEEE,* and Stephan M. Rohlfing

***Abstract*— The folded half-wavelength resonators in hairpin-comb filters all have the same orientation, whereas the orientations of the resonators in conventional hairpin-filters alternates. Hairpin-comb filters are shown to have attractive properties for design of compact, narrow-band filters such as are often desired for high-temperature-superconductivity (HTS) and other applications. The results obtained from two- and four-resonator trial HTS-filter designs are discussed. The filters are seen to have very strong stopbands, and their computed and measured performance are found to be in very good agreement.**

***Index Terms*— Bandpass filters, microstrip filters, superconducting filters.**

I. Introduction

IN MANY applications, keeping filter structures to a minimum size is very important. This is particularly true of high-temperature superconductor (HTS) filters where the available size of usable substrates is quite limited. In the case of narrow-band microstrip filters (say, with bandwidths of the order of 1% or less), this problem can become quite severe because the substantial difference between the even- and odd-mode wave velocities, when the substrate dielectric constant is large, can create relatively large forward coupling between the resonators. This presents a need for large spacings between the resonators in order to obtain the required narrow bandwidth [1]. This may make the overall structure unattractively large, or totally impractical for some HTS situations. Hairpin-comb filters provide a possible way around this problem. Of course, the use of hairpin resonators also reduces the size of a filter since the folded half-wavelength resonators are somewhat less than a quarter-wavelength long. Hairpin-comb filters have some properties similar to those of comb-line filters, but have an advantage in that they do not require any ground connections. This is particularly important for circuits such as microstrip HTS filters for which making ground connections would be very difficult.

Manuscript received November 21, 1996; revised April 25, 1997. This work was supported in part by the Naval Research Laboratory under Contract N00014-95-C-2126.

G. L. Matthaei and N. O. Fenzi are with Superconductor Technologies, Inc., Santa Barbara, CA 93111-2310 USA.

R. J. Forse was with Superconductor Technologies, Inc., Santa Barbara, CA 93111-2310 USA. He is now with Motorola, Schaumburg, IL, 60196 USA.

S. M. Rohlfing was with Superconductor Technologies, Inc., Santa Barbara, CA 93111-2310 USA. He is now with Rockwell International, Newbury Park, CA 91320 USA.

Publisher Item Identifier S 0018-9480(97)05378-7.

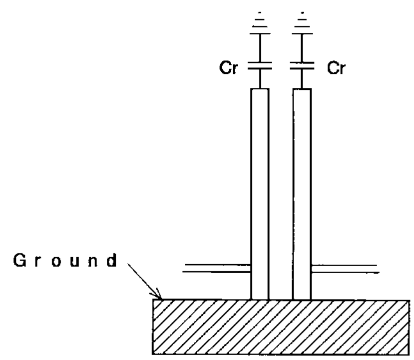


Fig. 1. A stripline two-resonator comb-line filter with tapped-line couplings at the input and output.

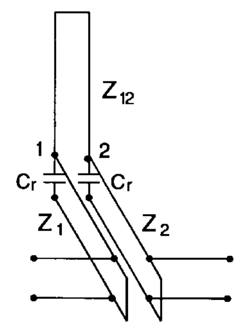


Fig. 2. An equivalent circuit for the stripline structure in Fig. 1 using a homogeneous dielectric.

II. Some Properties of Comb-Line Filters and Hairpin-Comb Filters

Fig. 1 shows a two-resonator comb-line filter [2]. For the present, it will be assumed to be realized in stripline with a homogeneous dielectric so the even- and odd-mode velocities on the coupled lines will be equal, thus preventing forward coupling. The two resonators are grounded at the crosshatched sidewall, and in this example, the input and output couplings are provided by tapped-line connections. It turns out that this structure would have no passband at all if it were not for the *loading* capacitors C_r. From the equivalent circuit for this combline filter shown in Fig. 2 it can be seen why this happens. Since the resonators in Fig. 1 are shorted at one end, with $C_r = 0$, they are resonant when they are a quarter-wavelength long. As seen from their nodes 1 and 2 in Fig. 2, they look like shunt-connected parallel-type resonators, which would yield a passband at this frequency. However, there is a series-connected shorted stub of impedance Z_{12}

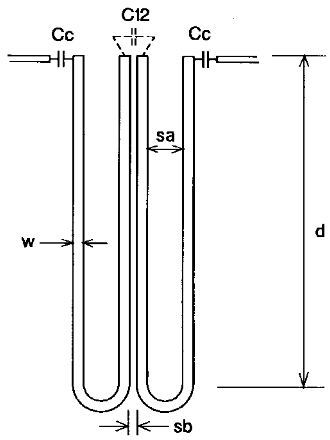


Fig. 3. A two-resonator hairpin-comb filter with capacitance couplings at the input and output.

between nodes 1 and 2. (The shunt stubs in Fig. 2 can be thought of as being associated with propagation having E fields extending from the resonators to the ground planes, while the series stub is associated with propagation having the E field extending from one resonator to the other in the slot between the resonator lines.) If the dielectric is homogeneous, the series stub is also resonant at this frequency and acts like a series-connected *bandstop* resonator between the two shunt *bandpass* resonators. This creates a pole of attenuation at the same frequency that a passband would otherwise occur. Thus, the potential passband is totally blocked. However, if loading capacitors C_r are added at the ends of the resonators in Fig. 1, the resonator lines must be shortened in order to maintain the same passband frequency. This shortens all three lines in Fig. 2 the same amount, and causes the pole of attenuation to move up in frequency, away from the passband. In general, the more capacitive loading that is used, the further the pole of attenuation would be above the passband and the wider the passband of the filter can be. If only small loading capacitors C_r are used, a very narrow passband can be achieved even though the resonators are physically quite close together. Similar operation also occurs if more resonators are present.

It is of interest to note in Fig. 2, $Z_1 = Z_2 = 1/(vC_1)$ while $Z_{12} = 1/(vC_{12})$, where v is the wave velocity in the structure, C_1 is the capacitance per-unit length between either line in Fig. 1 and ground, while C_{12} is the capacitance per-unit length between the lines. If the structure in Fig. 1 is realized in the microstrip, the even and odd modes have different velocities. Then the corresponding equivalent circuit is too complicated to be very useful for purposes of physical insight such as was discussed in connection with Fig. 2 for the homogeneous case. However, some of the same kinds of circuit properties exist, though in modified form.

Fig. 3 shows a *hairpin-comb* filter of the type treated in this paper. This filter is somewhat analogous to the combline filter in Fig. 1. (In this case, series-capacitance input and output couplings are shown, though tapped-line couplings (as in Fig. 1) could be used in this filter also.) The resonator lines are roughly a half-wavelength long, but are folded back on themselves so the height of the resonators is a little less than

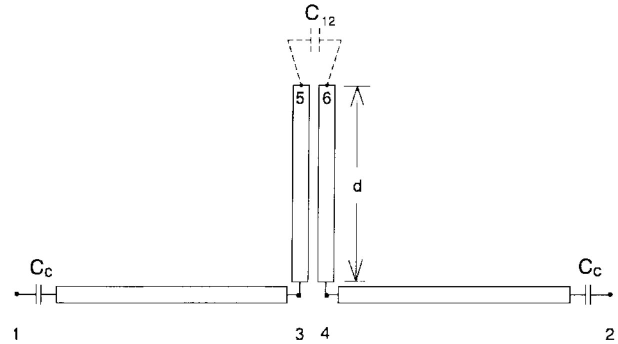


Fig. 4. A simplified model for the circuit in Fig. 3. In this model the weaker couplings between lines are ignored.

a quarter wavelength. Note that, unlike the combline filter in Fig. 1, the structure in Fig. 3 has no ground connections. However, since the opposite sides of a hairpin resonator have opposite potentials, there is a virtual ground running down through the center line of symmetry of the resonator. Thus, one might expect the filter in Fig. 3 to have properties similar to those of a combline filter. It turns out that the hairpin-comb filter does have similarities to a combline filter, but the hairpin-comb behavior is more complex. Capacitors such as C_{12} in the figure are optional but can be included to provide more flexibility in a design.

Insight into the operation of hairpin-comb filters can be obtained using the simplified equivalent circuit in Fig. 4 to model the two-resonator filter in Fig. 3. In this equivalent circuit, the closely coupled lines of length d in the resonators are retained in their original form but the remainder of each resonator is straightened out and positioned in dog-leg form. The net effect of this approximation is that the weaker couplings between the more separated lines are ignored as well as the changes in propagation due to the curvature in the lines at the bottom of the resonators. Comparison was made between a more exact analysis of the structure in Fig. 3 (which included all of the couplings and quite accurate modeling of the curvature at the bottoms of the resonators) and the approximate structure in Fig. 4. This comparison showed excellent qualitative agreement between the two circuit models though the frequency scale for the response of the circuit in Fig. 4 was somewhat altered due to the approximations involved.

The operation of the simplified circuit in Fig. 4 can be viewed as being more-or-less dual to that of the circuit in Fig. 2. Assume for the moment that the circuit in Fig. 4 has a homogeneous dielectric so that the odd and even modes have the same velocities, the vertical coupled stubs are $\lambda/4$ long when the resonators are resonant, and $C_{12} = 0$. Since nodes 5 and 6 of the coupled stubs are open-circuited, the propagation having E fields extending between the two lines has a $\lambda/4$-resonance, which reflects a short circuit *between* nodes 3 and 4, and as a result, provides a direct connection between the two resonators. If this were the only effect present, there would be very strong transmission between nodes 1 and 2. However, at this same frequency, the propagation on the coupled lines, which has E fields extending from the lines to ground, acts like $\lambda/4$ open-circuited stubs connected in shunt at nodes 3 and 4. The resonance associated with this propagation effectively shorts both nodes 3 and 4 directly to ground. This shorts out all transmission and creates a pole of attenuation at the frequency at which the resonators are resonant. Because of this pole of attenuation in the middle of what would otherwise be the passband of the structure, the structure has no passband at all, no matter how closely spaced the coupled lines are— an effect similar to that which occurs in conventional comb lines if no loading is used on the resonators.

Now, if the coupled lines in Fig. 4 have a length d, which is less than $\lambda_o/4$ at the frequency for which the resonators have their $\lambda_o/2$ resonance, the pole of attenuation created by the coupled lines will occur at a frequency above the frequency of resonance of the resonators, and the structure will have a passband. If d is not greatly less than $\lambda_o/4$, the pole will be relatively close to the passband and the passband will be very narrow, even though the resonators may be closely spaced. This ability to use closely spaced resonators, plus the fact that the folded resonators are somewhat less than $\lambda_o/4$ in height, makes it possible to design very compact narrow-band hairpin-comb filters.

When a hairpin-comb filter is realized in microstrip so that the even- and odd-mode velocities are unequal, similar performance occurs but with significant differences that are not easily explained by physical intuition. For example, we analyzed several two-resonator designs having proportions as in Fig. 3 using substrate dielectric constants ranging from 2.6 to 24.1 are analyzed. In all cases, even though the coupling length d was less than $\lambda_o/4$ (the resonators in Fig. 3 are nominally $\lambda_o/2$ long if straightened out), the adjacent pole of attenuation occurred *below* the passband rather than above (as occurs in the case of homogeneous dielectric). The larger the dielectric constant of the substrate, the lower in frequency this pole was (relative to the passband frequency). An interesting and useful phenomena here is that if a capacitance C_{12} is included in the gap, as is suggested in Fig. 3, the pole of attenuation moves *up* in frequency. Thus, by addition of the capacitance C_{12} the pole can be moved up close to the passband so as to give a very narrow passband, or, by making C_{12} still larger, the pole can be moved up through the passband to provide a pole on the upper side of the passband. This resonance phenomena, which occurs in the coupling regions between resonators, can be a very useful property.

III. Differences Between Hairpin-Comb and Conventional Hairpin-Resonator Filters

Fig. 5(a) shows a well-known form of a hairpin-resonator bandpass filter [3]. It can be thought of as an alternative version of the parallel-coupled-resonator filter first introduced by Cohn [4], except that here, the resonators are folded back on themselves. Note that the orientations of the hairpin-resonators alternate. This results in quite strong coupling, and the structure is capable of considerable bandwidth. However, in the case of narrow-band filters, particularly for microstrip filters on a high-dielectric substrate, this structure is undesirable as it may require quite large spacings between the resonators in order to achieve the desired narrow bandwidth.

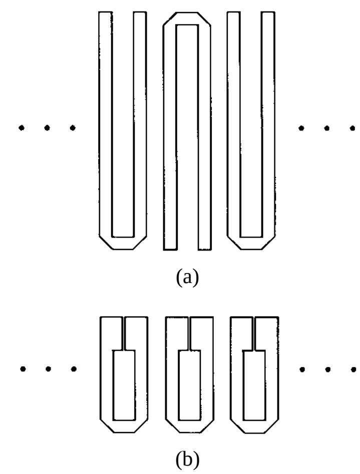


Fig. 5. (a) Common form of a hairpin-resonator filter structure. Note the alternating orientations of the resonators. (b) Another common form of a hairpin-resonator filter which uses resonators that have reduced height by virtue of capacitance loading. In this case the orientations of the resonators makes little difference.

The hairpin-comb type of filter (as in Fig. 3) differs from the hairpin filter in Fig. 5(a), primarily in that the orientation of the resonators in a hairpin-comb filter is always the same. This difference is of major importance. Resonances that occur in the coupling regions between resonators, greatly reduce the coupling between resonators, and for the case of microstrip structures, the addition of a small capacitance C_{12} between resonators (as is shown in Fig. 3) permits the elimination of the passband entirely, even though the resonators are quite closely spaced.

Fig. 5(b) shows another common form of a hairpin-resonator filter [5] which might at first be thought to be fundamentally the same as the hairpin-comb filter in Fig. 3, whereas it is actually quite different. In this case the open-circuited ends of the resonator have been considerably foreshortened, and then a strongly capacitive gap is added to bring the remaining structure to resonance. The resonators are then semilumped, the lower part being inductive and the upper part being capacitive. The coupling between resonators is almost entirely inductive, and it makes little difference whether adjacent resonators are inverted with respect to each other or not. Hence, they are usually made to have the same orientation. This structure may permit reasonably small spacings between the resonators because the reduced size of the inductive part of the resonators reduces the mutual inductance between resonators. However, the resonators in Fig. 5(b) tend to have relatively low Q's. This occurs because the electric and magnetic stored energy in a resonator must be equal at resonance. Since the inductive part of the resonators in Fig. 5(b) has been considerably reduced in size, a relatively high current density will be required in order to produce the needed stored magnetic energy. The resonator loss increases considerably since the conductor loss is proportional to the square of the current density. In the case of normal metal circuits, usually the conductor loss is the dominant loss, though in HTS circuits, this may not be the case. (For example, in some microstrip HTS circuits, radiation losses may dominate.) However, even in HTS circuits, the relatively high current density in resonators of the type in Fig. 5(b) is undesirable since high current densities can lead to nonlinear effects.

From the above examples it can be seen that the hairpin-comb type of filter differs from the hairpin-filter structures in Fig. 5(a) and (b), in that the hairpin resonators all have the same orientation, while the coupling regions between resonators are sufficiently long so as to have resonance effects which can greatly reduce the coupling between resonators at frequencies in the range of the desired passband. It will also be noted that the hairpin-comb structure in Fig. 3 uses rounded sections at the bottoms of the resonators, rather than rectangular sections as in Fig. 5(a) and (b). This is not fundamental to this type of filter, but these round sections have been used to prevent regions with unnecessarily high current density—which may cause nonlinear effects in the superconductor.

IV. Some Design Examples

Herein, the focus is on examples of narrow-band microstrip hairpin-comb HTS filters. In such cases, the couplings beyond nearest-neighbor resonators is much more important than it would be in relatively wide-band hairpin-filter structures, as in Fig. 5(a). This is because for a hairpin-comb filter the direct coupling between adjacent resonators is relatively small so that the stray couplings beyond nearest-neighbor line sections becomes much more important. In order to obtain accurate designs it is important to include couplings beyond nearest neighbors. This makes the use of the more common design procedures based on network synthesis techniques impractical. As a result, what might be called *educated cut and try* was used to obtain the desired responses. An in-house CAD program was used which handles multiple lines utilizing the method of lines (MoL) [6], and which also treats single- or multiple-curved line sections using the methods in [7]. This program obtains the quasi-static capacitance and inductance matrices for multiple lines, and uses these data for computing frequency responses. Structures like the semilumped capacitors were designed with the aid of the planar full-wave analysis program EM.[1]

A two-resonator microstrip HTS filter (as in Fig. 3) was designed using a $LaAlO_3$ substrate $h = 0.267$ mm (10.5 mil) thick having $\varepsilon_r = 24.1$. The HTS was YBCO, and in the figure $d = 8.504$, $sa = 1.0$, $w = 0.30$, and $sb = 0.20$ (all in mm). The coupling capacitance C_c was about $C_c = 0.216$ pF, though a π-equivalent circuit for the coupling capacitor was actually used. Accurate analysis of the capacitor C_{12} as designed turned out to be a problem. This is because the two ports for the capacitor were adjacent and so close together as to interact, and the capacitor finger structure was not symmetrical as viewed from these ports. (If the finger structure had been symmetrical as seen from the ports, an accurate analysis could

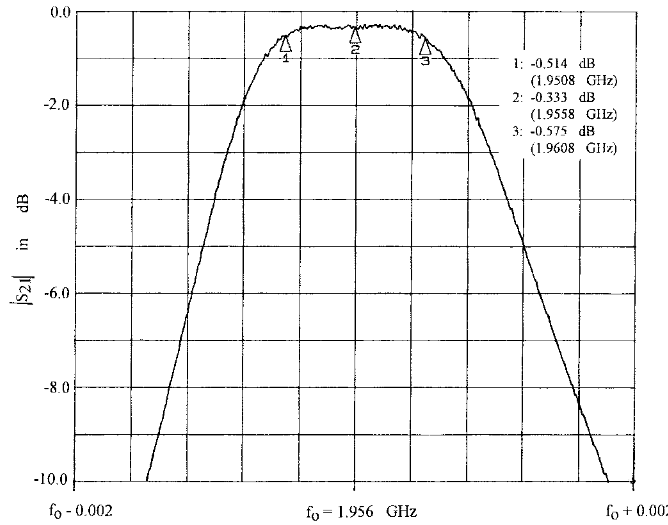


Fig. 6. A measured response for a trial microstrip HTS two-resonator hairpin-comb filter of the form in Fig. 3.

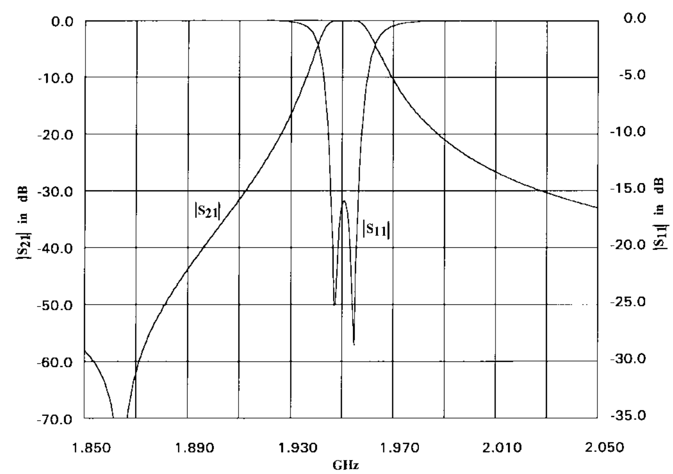


Fig. 7. A broad-range computed response for the two-resonator trial filter.

have been obtained using even- and odd-mode excitation.) A final value for C_{12} (0.076 pF) for use in computing the theoretical response was obtained by varying the value of C_{12} used in the program until the computed frequency of the pole of attenuation below the passband closely agreed with the measured frequency for that pole (1.865 GHz). Then the computed passband width at points 1 dB down from the minimum attenuation was $\Delta f = 14.8$ MHz, and the passband center frequency was computed to be $f_o = 1.97$ GHz. This compares with measured values of $\Delta f = 14.2$ MHz and $f_o = 1.955$ GHz. This is an approximately 0.73% bandwidth. Fig. 6 shows the measured passband response of this filter, while Fig. 7 is a computed response showing the nature of the response of this type of two-resonator filter on a more broad-range basis. Note that as previously mentioned, for microstrip cases such as this the pole of attenuation adjacent to the passband occurs below the passband even though the coupling region length d is less than $\lambda_o/4$. The measured response data was obtained with the circuit cooled to 77° K. With regard to the unloaded Q of the resonators, we have not made any systematic tests on single HTS hairpin

[1]EM is a full-wave field solver for planar circuits, produced by Sonnet Software, Liverpool, NY 13090.

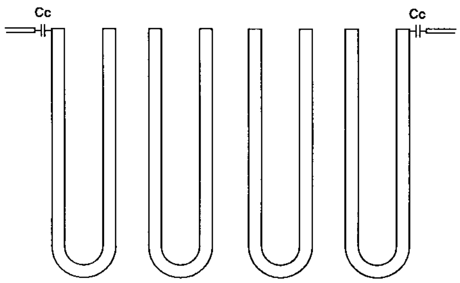


Fig. 8. A four-resonator hairpin-comb filter with capacitance couplings at its ends.

resonators in order to evaluate their Q's. However, tests made at Superconductor Technologies, Inc. (STI) on certain forms of microstrip bandstop resonators have yielded Q's in excess of 30 000 at frequencies around 2 GHz. One might be tempted to try to estimate the average unloaded Q of the resonators in the present filter from measurement of the minimum loss at midband. However, in this case, that does not work because the housing used to test the filter included normal-metal connectors, a 1.07-cm-long section of gold microstrip at each end of the housing (which provided a convenient transition between the connectors and the HTS circuit), and bond wires between these various parts. The minimum loss in the passband was 0.33 dB, most of which is believed due to loss in the nonHTS parts.

A four-resonator trial HTS microstrip hairpin-comb filter (as shown in Fig. 8) was also designed, fabricated, and tested. Using dimension definitions as in Fig. 3, in this case, the YBCO circuit used $d = 8.626$, $sa = 1.5$, $w = 0.5$, and the spacing between the resonators at the center of the filter was 1.45 (all in mm). The $LaAlO_3$ substrate had dimensions 16 $\times$ 34 $\times$ 0.283 mm (0.630 $\times$ 1.339 $\times$ 0.011 in) with $\varepsilon_r = 24.1$. Fig. 8 is somewhat oversimplified in as much as a section of the resonator line next to C_c on both ends of the filter was reduced in width. This was done in order to reduce the capacitance to ground in those regions to compensate for the capacitive detuning effect of the coupling capacitors C_c. The capacitors C_c were realized in interdigital form with seven fingers. Their effect was largely as a series capacitance, but they were modeled as a π-capacitor circuit in order to include their stray capacitance to ground. Slight tuning of the two inner resonators was accomplished by insertion of dielectric material near the resonators. Fig. 9(a) shows the measured and computed transmission response of the filter while Fig. 9(b) shows the measured and computed return loss. The passband width at points 1-dB down from the minimum loss point was 17.2 MHz, and the measured passband was centered at 1.8360 GHz. The percentage bandwidth was 0.94. The minimum passband loss was measured to be about 0.41 dB. Note that this is only about 0.08 dB more than the minimum loss that was observed for the two-resonator filter. (Of course, the somewhat wider bandwidth of the four-resonator filter reduced its midband loss.) The relatively small difference between the loss of the two- and four-resonator filters is consistent with the assumption that most of the minimum loss measured was due to the nonHTS parts associated with the filter test housings.

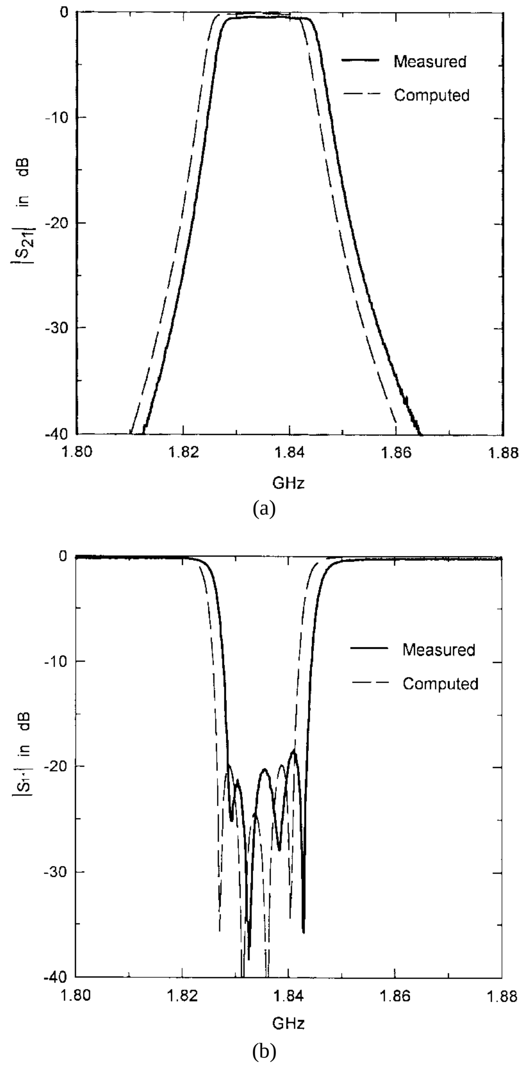


Fig. 9. (a) Computed and measured transmission responses for a trial HTS microstrip four-resonator filter of the form in Fig. 8. Some tuning of the center two resonators was required for the measured response. (b) Computed and measured return loss for the trial four-resonator filter.

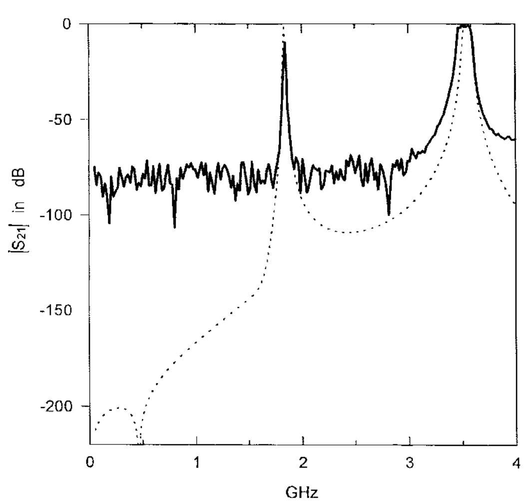


Fig. 10. Broad-range computed and measured responses for the trial four-resonator microstrip filter. The measured response is limited by the network analyzer dynamic range.

Fig. 10 presents a computed response which shows the predicted response for this filter over a wide range of frequencies and attenuation. Corresponding measured data is also shown, but it was impossible to measure the extremely high attenuations that these computations reveal. These results show that this type of filter has inherently very strong stopbands. Note that for the measured data the second passband is wider and somewhat lower in frequency than the computed second passband. This is believed to be due to the affects of dispersion, which are not included in this paper's computer analysis. It is of interest to note that the pole of attenuation at about 0.4 GHz in the computed response is also observed in the computed response of the center two resonators in this structure taken by themselves. Thus, this pole appears to be associated with the coupling gaps between resonators. However, note the knee in the attenuation characteristic at about 1.7 GHz. This is indicative of poles of attenuation nearby (somewhat off of the $j\omega$ axis of the complex frequency plane). These poles are believed to be due to coupling beyond nearest-neighbor resonators.

V. Conclusion

The hairpin-comb type of filter holds promise for the fabrication of compact narrow-band filters with no ground connections. This can be useful for planar filters designed using normal metal conductors but is particularly helpful for HTS filters. It can be shown that this general type of structure is potentially useful for either stripline or microstrip realizations, though the designs will come out rather different for given design specifications. The coupling mechanism between hairpin-comb resonators is seen to create a pole of attenuation which is usually near the passband. With the addition of capacitors between adjacent resonators it is found to be possible to adjust the position of the poles of attenuation which were generated in this manner. Computed examples indicate, that in the microstrip case, it is possible to place such poles on either side of the passband, which is a potentially very useful property. Also, in the microstrip case, additional poles of attenuation are observed due to couplings beyond nearest-neighbor resonators. The two HTS-filter examples demonstrated very good measured performance which was in good agreement with the computations. The stopband attenuation of this type of filter is seen to be exceptionally strong.

References

[1] G. L. Matthaei and G. L. Hey-Shipton, "Concerning the use of high-temperature superconductivity in planar microwave filters," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1287–1293, July 1994.

[2] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*. Norwood, MA: Artech House, 1980, pp. 497–506; 516–518.

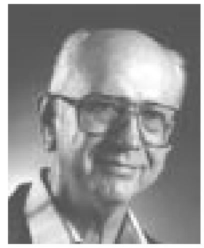
[3] E. G. Cristal and S. Frankel, "Hairpin-line and hybrid hairpin-line/half-wave parallel-coupled-line filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 719–728, Nov. 1972.

[4] S. B. Cohn, "Parallel-coupled transmission-line-resonator filters," *IRE Trans. Microwave Theory Tech.*, vol. MTT-6, pp. 223–231, Apr. 1958.

[5] M. Sagawa, K. Takahashi, and M. Makimoto, "Miniaturized hairpin resonator filters and their application to receiver front-end MIC's," *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 1991–1997, Dec. 1989.

[6] R. Pregla and W. Pascher, "The method of lines," in *Numerical Techniques for Microwave and Millimeter-Wave Passive Structures*, T. Itoh, Ed. New York: Wiley, 1989.

[7] H. Diestel, "A quasi-TEM analysis for curved and straight planar multiconductor systems," *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 748–753, Apr. 1989.



George L. Matthaei (S'49–A'52–M'57–F'65–LF'89) received the B.S. degree from the University of Washington, Seattle, in 1948, and the Ph.D. degree from Stanford University, Stanford, CA, in 1952.

From 1951 to 1955, he was on the faculty of the University of California, Berkeley, where he was an Assistant Professor, with a specialty in network synthesis. From 1955 to 1958, he was engaged in system analysis and microwave-component research at the Ramo-Wooldridge Corporation. From 1958 to 1964, he was at Stanford Research Institute, engaged in microwave-device research, where he became Manager of the Electromagnetic Techniques Laboratory in 1962. In 1964, he joined the Department of Electrical Engineering at the University of California, Santa Barbara, where he was a Professor. In 1991, he retired and became a Professor Emeritus. Following his retirement from teaching, he joined the staff of Superconductor Technologies, Inc., in Santa Barbara, CA, on a part-time basis. He is the author of numerous papers, co-author of the book *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, and a contributor to several other books. His current interests are in the area of the application of high-temperature superconductivity to passive and active microwave and millimeter-wave circuits.

Dr. Matthaei is a member of Tau Beta Pi, Sigma Xi, and Eta Kappa Nu. He was the winner of the 1961 Microwave Prize of the IEEE Microwave Theory and Techniques Group. In 1984, he received the IEEE Centennial Medal, and in 1986 the Microwave Career Award of the IEEE Microwave Theory and Techniques Society.

Neal O. Fenzi (M'93) received the B.S.E.E. degree from New Mexico State University, Las Cruces, in 1982.

From 1982 to 1984, he worked with the microwave filters division of Hughes Aircraft Co., Space and Communications Group, designing and building filters for satellite communications. From 1984 to 1991, he was with Amplica, Inc., designing microwave amplifiers and subassemblies for the telecommunications and military markets. Since 1991, he has been an Engineering Section Head with Superconductor Technologies, Inc., Santa Barbara, CA. He has authored many papers and holds one patent in the HTS field. His area of interest is applying HTS technology to the design of useful products.

Roger J. Forse (M'94) received the B.S.E.E. and M.S.E.E. degrees from the University of Illinois, Urbana, in 1970 and 1975, respectively.

From 1975 to 1990, he designed thin-film receiver subsystems at Watkins-Johnson and Avantek. In 1990, he joined Superconductor Technologies, Inc., Santa Barbara, CA, where he worked on HTS filter design. In 1995, he joined Motorola, Schaumburg, IL, where he is currently working on the modeling of RF printed circuits.

Stephan M. Rohlfing received the B.S.E.E. degree from Stevens Institute of Technology, Hoboken, NJ, in 1984.

From 1984 to 1988, he worked for the Microwave Semiconductor Corporation, Somerset, NJ, designing both hybrid and discrete amplifiers. He was with Veritech Microwave, South Plainfield, NJ, before joining Amplica, Inc., Newbury Park, CA, in 1988, where he designed hybrid amplifiers. In 1991, he joined Superconductor Technologies, Inc., Santa Barbara, CA, where he did research and development on HTS filters. In 1996, he joined Rockwell International, Wireless Communications Division, Newbury Park, CA, where he designs MMIC's.