Concerning the Influence of Housing Dimensions on the Response and Design of Microstrip Filters with Parallel-Line Couplings

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***Abstract*—In this paper, measured results obtained from a narrow-band microstrip filter are compared with computed responses obtained using two different classes of software for various assumed housing conditions. (Some results from filters with wider bandwidths are also cited.) Housing modes are found to have a potentially significant affect on the bandwidth of microwave filers that involve microstrip coupled lines, even though the housing resonant frequencies are much above the passband of the filter. When analyzing such filters using full-wave three-dimensional (3-D) or 3-D planar field solver programs, it is found to be necessary to accurately model the housing if high accuracy in the computed response is needed. For carrying out filter design by optimization, it is usually convenient to use faster programs that utilize a full-wave two-dimensional (2-D) field solver to obtain line parameters and then use transmission-line analysis for the third dimension. Such programs can introduce housing-mode errors since 2-D full-wave analysis implies an infinitely long housing. Ways for getting around this problem are suggested. Physical explanations for the various effects observed are presented and are supported by computer studies of the natural frequencies of a coupled-line microstrip filter structure in the presence of various housing perturbations.**

***Index Terms*—Bandpass filters, circuit modeling, high-temperature superconductors, microstrip filters, thin-film circuit packaging.**

I. INTRODUCTION

FOR MANY narrow-band high-temperature superconductor (HTS) applications, very accurate analysis of the circuit structures is required in the design process. However, at Superconductor Technologies Inc. (STI), Santa Barbara, CA, for both of the two widely used classes of commercial software programs, we frequently experienced unacceptable error between measured and computed responses of microstrip filter structures that involved parallel-line couplings. Further, we found that the two classes of software regularly disagreed with each other in a predictable pattern. (This phenomenon has also been observed by others [1].) This research was undertaken to find the source for these errors and, if possible, eliminate them. As will be shown, the errors turned out to be due to unexpectedly strong influence of housing modes, even though all the housing modes were well below resonance in the operating range of interest. Thus, it became evident that high accuracy in modeling the housing can be much more important than we previously thought. The difference between computed responses obtained by the two classes of software turned out to be due to the fact that one of the two classes inherently views the housing inaccurately. Unfortunately, the class of software that has this problem is by far the fastest. A way around this difficulty will be described in this paper. In the course of this study, useful physical insights were obtained as to why housing modes affect the bandwidth of frequency responses in the way they do. These insights will be discussed in connection with examples of measured and computed responses.

One of the classes of software involved in this study includes a two-dimensional (2-D) field solver to get line parameters, which are then used to rapidly compute a filter response using one-dimensional transmission-line analysis. Herein, we will refer to this as "2D + 1 CAD." In this paper, the 2D + 1 CAD results were obtained with "Super Compact,"[1] which does a *full-wave* 2-D analysis. Models for the discontinuities in the filter structure for inclusion in the 2D+1 CAD analysis were obtained using the full-wave three-dimensional (3-D) planar solver "Sonnet."[2] The 3-D planar solver was also used for computing complete frequency responses, and we shall refer to these calculations as "3D planar CAD." In order to pin down possible sources of error, a test HTS filter was designed and fabricated, which used an array of four parallel half-wavelength microstrip resonators, with quarter-wave parallel coupling lines at the ends, as shown in Fig. 1. This form of filter was used because of its simplicity and because it emphasizes the involvement of coupling between parallel lines.

II. RESPONSE EFFECTS DUE TO MOVING THE SIDEWALLS

Let us now review some of the experimental and computer experiments that were made and their results. At this point, we use 3D planar CAD analysis since it can accurately model the housing.

For the filter in Fig. 1, the substrate was MgO, 0.507-mm thick with $\varepsilon_r = 9.7$. The filter structure used TBCCO-2212 HTS, and the resonators were 14.35-mm long, 0.5-mm wide, with spacings of 1, 1.4, and 1 mm between, while the coupling lines at the ends were 0.2-mm wide, 7.20-mm long, and

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[1]Super Compact is a registered trademark of the Ansoft Corporation, Elmwood Park, NJ.

[2]Sonnet is a registered trademark of Sonnet Software Inc., Liverpool, NY.

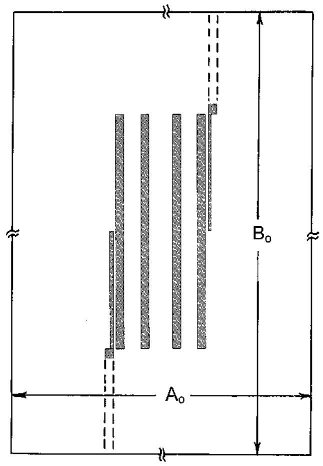

Fig. 1. Four-resonator-array test filter.

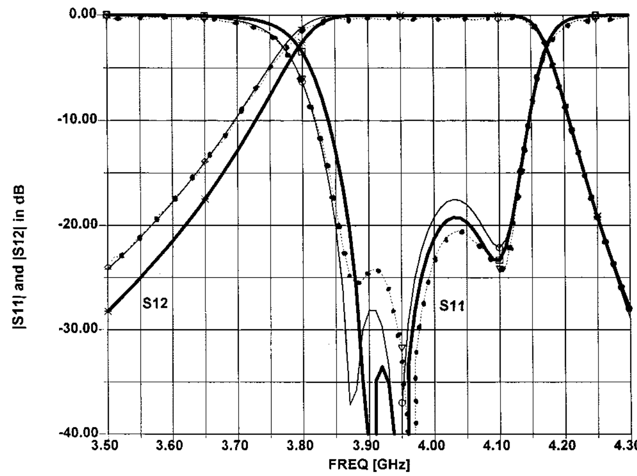

Fig. 2. Dots show the measured responses of the filter in Fig. 1 while the narrow solid lines show the 3D planar CAD results with the actual $A_o = 25.6$, $B_o = 51.2$ mm housing dimensions. The heavy solid lines show 3D planar CAD results if $A_o = 12.8$, $B_o = 25.6$ mm are assumed instead.

spaced 0.15 mm from the end resonators. The housing cover was 3.81 mm above the substrate, and the housing dimensions were approximately $A_o = 25.6$ and $B_o = 51.2$ mm. However, the initial 3D planar CAD responses were done assuming housing dimensions of $A_o = 12.8$ and $B_o = 25.6$ mm in order to speed the analysis. Using these dimensions gave a clearance of 3.7 mm between the resonators and the sidewalls at the left- and right-hand sides in Fig. 1 and 5.6 mm between the resonators and the sidewalls at the top and bottom of the drawing. It was thought, at first, that any further increase in spacing to the housing sidewalls should have little effect in the filter analysis, but this proved to be wrong.

The heavy solid lines in Fig. 2 show the $S11$ and $S12$ frequency responses computed using 3D planar CAD assuming the reduced-size 12.8 × 25.6 mm housing discussed above, while the actual measured responses are given by dotted lines. It is seen that the computed response is appreciably narrower and that all of the error occurs on the low side of the response. The narrow solid lines in the figure show the computed response using 3D planar CAD assuming the actual 25.6 × 51.2 mm housing dimensions. It is seen that now the agreement is excellent. For the $S12$ response the dots fall virtually on top of the narrow solid lines. There is some difference between the measured and computed $S11$ responses, but these differences occur at return loss levels of around 20 dB or more for which even a small difference in impedance makes a sizable difference in decibel return loss.

It was surprising to the authors that the frequency response was as sensitive to housing size as the results in Fig. 2 indicate. One might assume that the change in response is due to changes in the quasi-static (i.e., quasi-TEM) fields of the filter structure resulting from moving the sidewalls. However, even for the assumed smaller 12.8 × 25.6 mm housing the distance from the filter structure to the sidewalls is quite sizable making interaction with these fields unlikely, and tests discussed in Section VI herein confirm that moving the sidewalls as described above has negligible affect on the quasi-static fields of the filter structure. Hence, the observed effects must be due to modes that involve the entire housing.

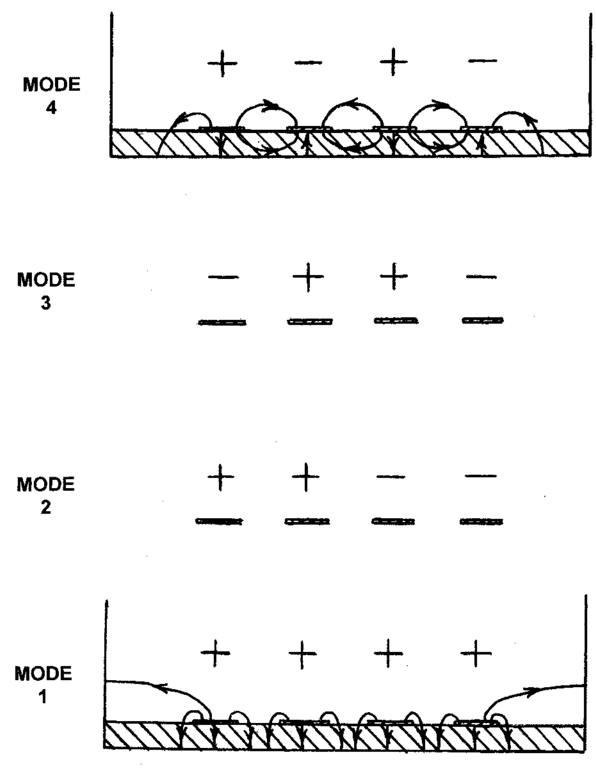

Fig. 3. Strip polarities for the four-resonator-array modes and some E-field sketches.

III. Explanation for the Effects of Moving the Sidewalls

Note from the 3D planar CAD calculations for the circuit in Fig. 1 that doubling the size of the housing dimensions has virtually no affect on the high side of the passband, but causes the lower side of the passband to be extended downward to give a larger bandwidth. We believe that this can be explained in terms of housing modes coupling to the quasi-static "array modes" depicted in Fig. 3. The filter consists primarily of an array of four coupled lines, and the figure suggests the nature of the four quasi-static natural modes of this four-line structure. Mode 1 is the mode with the lowest velocity of propagation (and, hence, the lowest resonant frequency). For this mode, the polarities of the potentials on all of the strips are the same and nearly all of the fringing electric field goes from the strips through the substrate to ground. Thus, the high-ε_r substrate has a relatively strong slowing affect on this mode. Mode 2 has a reversal of the potentials on the strips. This causes part of the E-field to fringe *through the air* from strip to strip, which results in a lower effective dielectric constant for the mode and a higher wave velocity and resonant frequency. As suggested in this figure, Mode 3 has two reversals in potential (which give a still higher resonant frequency), while Mode 4 has three reversals in potential, the most electric field in the air, and the highest resonant frequency. Now the lowest order housing modes are what Harrington refers to as TMx modes [2]. These lowest order housing modes have their electric fields almost entirely in the direction normal to the substrate. From the sketch in Fig. 3, it is easily seen that Mode 1 of the array of strips potentially can couple very well to these housing modes because most of its E-field is also in the direction normal to the surface of the substrate. However, array Modes 2–4 will couple very little if at all to these housing modes because the reversals in polarities cause the coupling effects to cancel out (note the E-field sketch for Mode 4).

The vertical component of the E-field in Mode 1 as sketched in Fig. 3 is even symmetric, which tends to couple well to the vertical E-field of the lowest housing mode. This mode can be designated as a $\mathrm{TMx}_{\delta_{11}}$ mode, where the δ subscript implies that there is only a small variation in the field strength in the x-direction (vertical). However, the vertical E-field as observed versus the lengthwise direction on each resonator is odd symmetric since the resonators are $\lambda/2$ long. This would make the resonators try to couple to the second housing mode rather than the lowest mode. This mode can be designated as the $\mathrm{TMx}_{\delta_{12}}$ mode.

From the preceding, it appears that the lowest frequency array mode couples to an "evanescent" (i.e., below resonance) housing mode, which, in turn, further lowers the frequency of that array mode. This results in the lower edge of the passband being moved downward. The passband frequency of this filter is below the lowest resonant frequency of the housing, but the closer the operating frequency is to a housing-mode resonance that the array can excite, the more the housing-mode evanescent fields should tend to couple to the array. The $\mathrm{TMx}_{\delta_{12}}$ resonant frequency for the 12.8 $\times$ 25.6 mm housing assumed when computing the heavy solid line in Fig. 2 was 15.55 GHz, well above the roughly 3.95-GHz passband center for the filter. Meanwhile, the corresponding resonant frequency for the actual 25.6 $\times$ 51.2 mm housing is 7.82 GHz, much closer to the filter passband center. This can be expected to result in tighter coupling between the lowest array mode and the $\mathrm{TMx}_{\delta_{12}}$ housing mode, which results in a lowering of the frequency of the lowest array mode and an extension of the lower edge of the passband downward as was observed. Since the housing does not couple to the higher order array modes, the higher order modes influence the upper end of the passband just as they did before the sidewalls were moved, and it can be seen from Fig. 2 that the upper side of the passband is largely unaffected by the change in housing size.

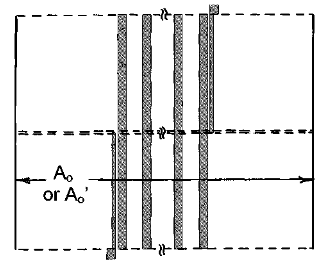

Fig. 4. Reinterpretation of the structure in Fig. 1 for use in an analysis program using 2D $+$ 1 CAD. (Fringing capacitances, etc., must also be included.)

IV. Considerations When Using 2D+1 CAD

When using 2D $+$ 1 CAD for modeling the filter in Fig. 1, it becomes necessary to break the structure up, as shown in Fig. 4, in order to account for the couplings between lines. Also, then fringing capacitances and junction inductances must be added. Their values were independently determined using 3D planar CAD. Since the input and output coupling lines are slightly longer than a quarter-wavelength, it is necessary to view the structure as three separate, arrays and then interconnect them. In Fig. 4, the region between the bottom horizontal dashed line and the lower of the two dashed lines near the center is represented by one array. In this region, the left-hand-side input line is included, but not the right-hand side. The region between the two closely spaced dashed lines in the center is analyzed separately because, in this region, the input lines at both ends need to be included, and then for the remaining top region, only the right-hand-side input line is included.

As previously mentioned, the 2D $+$ 1 CAD herein used Super Compact, which provides a full-wave 2-D solution. Being a 2-D solver it assumes that the lines are infinitely long. In Fig. 4, the dashed lines represent boundaries between regions, each of which is a section of an array of infinitely long lines. The array sections in Fig. 4 also have TMx housing modes, but their lowest resonance is for a housing as in Fig. 1 with B_o equal to infinity and all fields constant in the line-length direction. A_o alone is then the dimension that fixes the housing resonant frequency. The resonant frequency for this "2-D housing" is found to be 5.54 GHz when $A_o = 25.6$ mm and B_o is infinite (as compared to 7.82 GHz for the $\mathrm{TMx}_{\delta_{12}}$ mode with $B_o = 51.2$ mm). Since the housing resonance of concern for the modified filter model in Fig. 4 is lower than for the actual structure, as in Fig. 1, one might expect that the housing mode for the structure in Fig. 4 will couple more strongly to the lowest array mode and, thus, move the lower edge of the passband down below the lower edge for the actual structure in Fig. 1. The wide solid lines in Fig. 5 show the results obtained by 2D $+$ 1 CAD including a full-wave 2-D simulator, thus, housing modes can be observed. The results do indeed show a lowering of the lower passband edge as compared to the measured results (shown by the dotted lines).

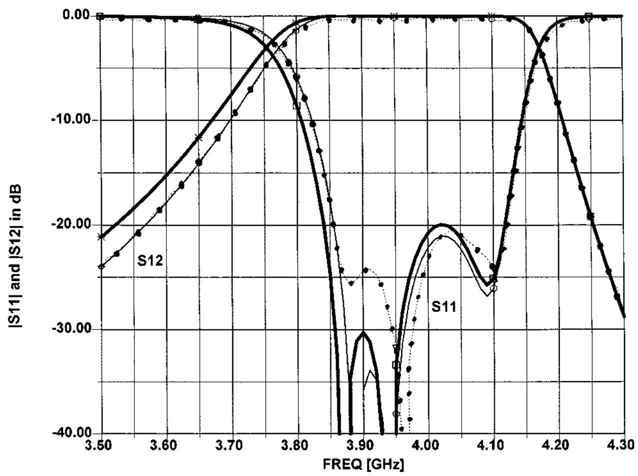

Fig. 5. Dots above show the measured response for the filter while the heavy solid lines show the response obtained utilizing 2D + 1 CAD with $A_o = 25.6$ mm in Fig. 4 with $f_{\rm calc} = 3.95$ GHz (i.e., line parameters are calculated at the center of the passband). The narrow solid lines show the 2D + 1 CAD results if we replace $A_o = 25.6$ with $A_o' = 13.8$ mm in Fig. 4.

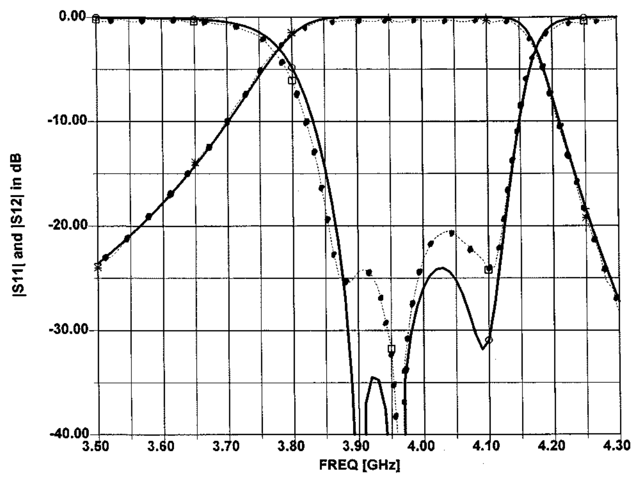

Fig. 6. Dots above show the measured response while the solid lines show the response computed using 2D + 1 CAD with $A_o = 25.6$ mm as in the actual filter, but with the line parameter computed using $f_{\rm calc} = 2.8$ GHz instead of 3.95 GHz.

In order to raise the frequency of the housing-mode resonance so as to reduce its coupling to the lowest array mode, the width of the housing was adjusted experimentally to give the best agreement with the measured results. Reducing the width of the housing from $A_o = 25.6$ mm to $A_o' = 13.8$ mm gave the results shown by the narrow solid line in Fig. 5. With this adjustment, the agreement between the 2D + 1 CAD and the measured results is seen to be excellent. With this A_o' value, the housing as defined in Fig. 4 has a 2-D resonance at 10.25 GHz, as compared to the $\mathrm{TMx}_{\delta_{12}}$ resonance of 7.82 GHz for the actual housing.

In the above example, the coupling to the housing modes is decreased in the 2D + 1 CAD calculations by reducing the housing width from $A_o = 25.6$ mm to $A_o' = 13.8$ mm. An alternative approach is to keep A_o at 25.6 mm and experimentally lower the frequency at which the 2D + 1 CAD program computes the line parameters from $f_{\rm calc} = 3.95$ GHz (used in all our previous calculations) to a lower frequency at which the coupling to the evanescent housing modes is reduced an appropriate amount. (Super Compact permits this.) A 2D + 1 CAD experiment was tried using $A_o = 25.6$ mm as for the actual structure, but with $f_{\rm calc}$ adjusted experimentally to the value $f_{\rm calc} = 2.80$ GHz, which caused the computed attenuation on the low side of the passband to agree very well with the measured values, as is seen in Fig. 6. There again, the dotted lines show the measured response while the solid lines show the 2D + 1 CAD computed results. The agreement between measured and computed results are, in general, quite good, though the agreement for the passband return loss is not quite as good as that shown in Fig. 5, where the agreement between the dotted measured data and the narrow-line 2D + 1 CAD data is remarkable. We believe the reason that here the lowering of $f_{\rm calc}$ to reduce the coupling to housing modes in the 2D + 1 CAD calculation does not work quite as well as does altering the housing width while keeping $f_{\rm calc}$ at the passband center is due to the influence of dispersion. Lowering $f_{\rm calc}$ not only affects coupling to evanescent housing modes in the line-parameter calculations, but also introduces some error in the calculation of dispersion since the $f_{\rm calc}$ used is not the actual passband center frequency. However, the error seen in Fig. 6 due to computing dispersion at a frequency a factor of $2.8/3.95 = 0.71$ below the actual passband center is seen to be quite small. It appears that for filters such as that in Fig. 1 (where the 20-mil substrate thickness is only about 1/47th of a wavelength in the medium of the substrate at the passband center) dispersion has very weak effects as compared to those caused by coupling to evanescent housing modes.

It should be noted that the housing for the structure in Fig. 1 provided considerable clearance between the filter structure and housing so that reducing the housing size did not cause any significant alteration of the quasi-TEM fields of the structure. If the housing had been smaller so that reducing the housing further would cause significant alteration of the quasi-TEM fields, then leaving the housing size unchanged and adjusting the coupling to the housing modes by varying $f_{\rm calc}$ would have been a much better approach. In Section VII, we suggest a procedure for fixing $f_{\rm calc}$ in 2D + 1 CAD optimization of filters so as to correctly account for the coupling of housing modes to parallel coupled lines in a filter.

V. Effects on Response Due to Varying the Lid Height

The heavy solid line in Fig. 7 shows the measured response for the filter in Fig. 1 when the cover height above the substrate is 3.81 mm, as in the previously discussed examples, while the dots show the measured response when the cover height is raised to 6.325 mm. The thin solid lines show the response computed using 3D planar CAD to model this latter situation. The measured and computed results are seen to be in very good agreement. It is interesting to note that just as for the cases where the sidewalls of the housing were moved, raising the lid of the housing affects only the lower side of the passband.

Results in Section VI indicate that the effect described above is not due to housing modes and can be explained in terms of the electric field alone. Consider the case of the E-field of Mode 4

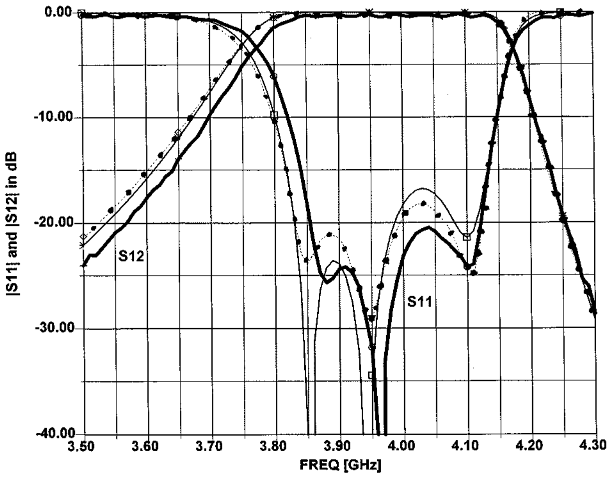

Fig. 7. Heavy solid lines show the measured response when for the filter in Fig. 1 the cover height is $h_u = 3.810$ mm, while the dots show the measured response when the cover height is raised to 6.325 mm. The narrow solid lines shows the 3D planar CAD computed response for $h_u = 6.325$ mm.

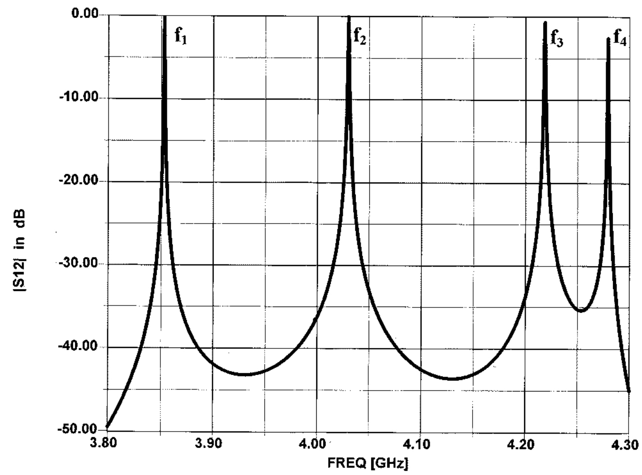

Fig. 8. 2D + 1 CAD response for the structure in Fig. 4 with the end coupling lines removed and 100 000-Ω terminations attached at the lower left-hand side and upper right-hand side.

sketched in Fig. 3. Since the polarities of the potentials on the strips alternate, the electric flux from the tops of the strips tends to largely fringe to their near neighbors and the height of the cover will make little difference since very little flux goes there. However, for the case of Mode 1, if in the sketch a cover is added, there will be flux extending from the tops of the strips (in which all have the same polarity) to the cover and this will reduce the amount of flux flowing through the substrate to the ground plane. Thus, the higher the cover plate, the less flux will go to the cover and the more electric flux will flow through the dielectric to the ground plane. This means that as the cover is raised, the velocity of Mode 1 will be lowered, and this will cause the lower side of the passband to be extended downward, as in Fig. 7. It is easily seen that because of the polarity reversal in Mode 2, it will be less sensitive to the position of the cover, and progressively more so for modes 3 and 4. Thus, when raising the lid height, Mode 1 will be slowed the most (as compared to its original velocity), Mode 2 will be slowed less, Mode 3 still less, and Mode 4 (the fastest mode) very little. These phenomena will be exhibited more explicitly in Section VI. Because of these relationships between lid height and array mode velocities, raising or lowering the lid height primarily affects the lower side of the passband.

VI. Verification of the Theory by Observation of the Filter Natural Frequencies

In order to study the various phenomena discussed above, it is useful to observe the "natural frequencies" of the filter structure with very high external Q's at the terminations. To make these observations, we used 2D + 1 CAD because of its speed. Further, when using Super Compact, we can make the transmission-line parameter calculations at the passband center frequency $f_{\rm calc} = 4$ GHz (which will cause the parameters to include effects of dispersion and coupling to housing modes), or, we can evaluate the line parameters at a very low frequency, say, $f_{\rm calc} = 0.01$ GHz, which effectively gives a quasi-static solution with no dispersion or housing mode effects. This then provides a powerful tool for identifying what may be due to dispersion and housing modes and what cannot be due to them.

TABLE I
Cases for Which Natural Frequencies Were Computed

Cases Evaluated			
Case	A_0 or A_0', mm	f_{calc}, GHz	h_u, mm
A	25.6	4.0	3.810
B	12.8	4.0	3.810
C	25.6	4.0	6.325
D	25.6	0.1	3.810
E	12.8	0.1	3.810
F	25.6	0.1	6.325

In order to compute the natural frequencies of the structure under various conditions, the coupling lines at the left- and right-hand-side ends of the filter in Fig. 4 were removed and 100 000-Ω terminations were connected at the lower left-hand-side and upper right-hand-side ends of the filter. In this situation, it was no longer necessary to break the structure into three arrays of lines, as was done in Section IV. Fig. 8 shows the computed response for this structure for Case A in Table I for which $A_o = 25.6$ mm in Fig. 4, $f_{\rm calc} = 4.0$ GHz, and the cover is $h_u = 3.810$ mm above the substrate. The natural frequencies for the four modes were then

$$\begin{aligned} f_1 &= 3.8530 \text{ GHz} \\ f_2 &= 4.0300 \text{ GHz} \\ f_3 &= 4.2183 \text{ GHz} \\ f_4 &= 4.2793 \text{ GHz} \end{aligned} \tag{1}$$

where the subscripts refer to the mode number. Similar calculations were made for all six cases defined in Table I. Herein, we are not interested so much in the frequencies of individual modes, but instead we are interested in the percentage changes in the frequencies when the structure parameters are changed. Such data is presented in Table II. There, for example, the row that starts with %BA gives the percentage change in each of the natural frequencies when we change to Case B from Case A,

TABLE II
PERCENTAGE DIFFERENCES IN NATURAL FREQUENCIES BETWEEN VARIOUS PAIRS OF CASES

Mode Frequency Shift in Percent				
	Δf_1	Δf_2	Δf_3	Δf_4
%BA	0.452	0.109	0.036	0.014
%AD	-2.027	-0.653	-0.274	-0.187
%ED	0.041	0.037	0.019	0.012
%CA	-0.208	-0.104	-0.012	-0.002
%FD	-0.839	-0.185	-0.024	-0.007

and the value of Δf_n in that row is the percentage shift in the resonant frequency of mode n. The other rows are interpreted analogously.

The %BA row gives the percent changes in the mode frequencies when the housing width is cut to 12.8 mm from 25.6 mm while $f_{\rm calc} = 4.0$ GHz. Note that the percent shift in the frequency of Mode 1 is considerably the largest, as was presumed to be the case in Section III. There we theorized strong coupling between the array Mode 1 and the $\mathrm{TMx}_{\delta_{12}}$ housing mode, and at most weak coupling between the other array modes and this housing mode. Also in line with our previous hypothesis, cutting the housing width in half reduces the coupling between the array Mode 1 and the $\mathrm{TMx}_{\delta_{12}}$ housing mode and, thus, causes f_1 to move up in frequency. Now the row for %ED in Table II gives percentage frequency shifts for the same change in housing width, but with $f_{\rm calc} = 0.01$ GHz so that both solutions are very nearly quasi-static. This virtually eliminates housing modes and dispersion from the calculation of the parameters for the coupled lines. Note that the frequency shifts indicated in row %ED are all very small. This shows that cutting the housing width in half had a negligible effect on the quasi-TEM fields of the array, and is consistent with the view that the large shift in the Mode 1 frequency observed in row %BA is due to a reduction in the coupling to a housing mode.

Row %AD in Table II shows the changes in resonant frequencies when the physical structure is held fixed with $A_o = 25.6$ mm while $f_{\rm calc}$ is changed to $f_{\rm calc} = 4.0$ GHz (dispersion and housing modes included) from 0.01 GHz (a quasi-static case). As expected, all of the natural frequencies are moved down somewhat since dispersion causes the energy to be somewhat more concentrated in the dielectric. Note that, again, Mode 1 is moved down decidedly the most due to its coupling to a housing mode.

Row %CA of Table II shows the shifts in the natural frequencies when $A_o = 25.6$ mm, and $f_{\rm calc} = 4.0$ GHz, while the lid height is raised to $h_u = 6.325$ mm from $h_u = 3.810$ mm. As discussed in Section V, raising the lid height causes less electric flux to go to the lid and more of the flux to pass through the substrate and terminate on the ground plane. The increased amount of electric flux in the substrate causes the waves to be slowed and the natural frequencies to be lowered, as is seen in Row %CA. Note that the frequency is lowered most for Mode 1, and the natural frequencies are progressively lowered less as the mode order is increased for reasons discussed in Section V.

In Section V, it was stated that the effects we observe when the lid height is increased are believed to be due to electric field effects rather than coupling to a housing mode. If this view is correct, then we should still see lowering of the natural frequencies when the lid height is raised even when $f_{\rm calc} = 0.01$, thus, we have a quasi-static solution. Row %FD in Table II shows the results for this situation. Note that in this case, not only are the shifts in natural frequencies present, but they are considerably larger than they were when dispersion and housing modes were present. We believe that the explanation for this lies in that the presence of dispersion tends to cause the fields to be more concentrated in the substrate so that when $h_u = 3.810$ mm and dispersion was accounted for there was less flux going to the lid to begin with. Raising the lid height then gave less change in the electric flux patterns; hence, less change in the natural frequencies.

VII. REGARDING USE OF 2D + 1 CAD FOR DESIGN OF COMB-LINE, INTERDIGITAL, AND OTHER FILTERS

As an example, let us now contemplate how the principles discussed above might affect the analysis and design of a microstrip comb-line filter [3]. For comparison purposes, we will assume a center frequency of $f_o = 3.95$ GHz for the structure in Fig. 1 and that the housing width is also $A_o = 25.6$ mm, as in Fig. 1. The resonator lengths for the comb-line structure could be chosen to be, say, 3/16-λ long [3], which would be about 5.4 mm, assuming we are using the same substrate as in Fig. 1. Thus, instead of having a vertically oriented array of four 14.35-mm-long resonators, as in Fig. 1, we would have a vertically oriented array of four 5.4-mm-long resonators with all of the resonators shorted to the bottom sidewall and capacitively coupled to ground at the upper sidewall [3]. In this case, the housing dimensions could be of the order of $A_o \times B_o = 25.6$ mm $\times$ 5.4 mm (instead of 25.6 mm $\times$ 51.2 mm, as in Fig. 1). In this comb-line case, the lowest housing resonance (the $\mathrm{TMx}_{\delta_{11}}$ housing mode couples to the comb structure) will be at about 26 GHz, which is about 6.6 times the passband center of the filter. It appears safe to assume that for this comb-line filter, any coupling to evanescent housing modes will be negligible because the operating frequency is so far below the lowest housing resonance. However, unfortunately, a 2D + 1 CAD analysis views the structure as having strong coupling to housing modes. This is because the 2-D full-wave analysis implies that the 5.4-mm lengths of the resonators are stretched out to infinity. Thus, the apparent housing size for the 2-D analysis is $A_o \times B_o = 25.6$ mm by infinity, with a lowest resonance of 5.5 GHz. This is only 1.4 times the passband center, and strong coupling to evanescent housing modes will be included in the computed response. This apparent coupling to housing modes will cause the computed passband to be overly large as compared to the actual response since the lower edge of the passband will be moved downward, as discussed in Section III. It is useful to note that the 2D + 1 CAD analysis error due to housing modes would be considerably worse for our hypothetical comb-line filter example than it is for the filter in Fig. 1. This is because in the case of Fig. 1, there is a considerable amount of coupling to housing modes in the actual structure so the large amount of coupling included in 2D + 1 CAD analysis with full-wave 2D field solutions is not so far off from reality. However, in the comb-line case with very weak housing mode activity, a 2D + 1 CAD analysis with a full-wave 2-D field solver is much further off from reality. However, the presence of the loading capacitors may reduce the response error to some degree since the part of the resonator energy in them is probably about correct.

As indicated in connection with the footnote to Section I, Swanson has long been aware of the fact that 2D + 1 CAD predicts overly large passbands for microstrip comb-line or interdigital filters. He has also devised a procedure for getting around this problem. He uses a 2D + 1 CAD program for design of such filters because of the speed such programs have in carrying out optimization. After an initial optimized design has been obtained for the desired bandwidth using the 2D + 1 CAD program with $f_{\rm calc}$ set at the passband center, he then computes the response of this initial design using a 3D planar CAD program in order to determine how much smaller the actual bandwidth will be. (The 3D planar CAD program correctly treats all housing effects, but is relatively slow.) A bandwidth shrinkage factor $S < 1$ equal to the ratio of the bandwidth computed using 3-D planar CAD analysis to the bandwidth of the initial 2D + 1 CAD design can then be computed. Last he reoptimizes the 2D + 1 CAD design for an enlarged bandwidth, which is $1/S$ times the original bandwidth. The bandwidth of this revised 2D + 1 CAD design should then also shrink by the factor S when the filter is fabricated, thus yielding the final desired bandwidth.

It appears that Swanson's procedure provides a general and very practical way of getting around the bandwidth errors that may occur in filter designs obtained by 2D + 1 CAD optimization. A filter such as that in Fig. 1, which uses a relatively large housing, could also be designed using his approach. However, in the case of microstrip comb-line and perhaps some interdigital filters, which use relatively small housings, it may be possible to use a simpler approximate procedure with good results. Since here we assume the actual filters under consideration have negligible coupling to housing modes, we would like to be able to remove the effects of such modes from the 2D + 1 CAD computed response in order to obtain the correct bandwidth. In a 2D + 1 CAD program such as Super Compact, we can eliminate housing modes by calculating the line parameters at a very low frequency such as $f_{\rm calc} = 0.1$ GHz, which will yield an essentially quasi-static solution. (Some 2D + 1 CAD programs may not do *full-wave* 2-D field analysis and, thus, already give a quasi-static solution.) This will, of course, remove dispersion as well as the housing modes from the line parameter calculations. However, from an example discussed in Section IV, we concluded that at least when the substrate is as electrically thin at the passband center frequency as it is for the structure in Fig. 1, dispersion effects are very weak compared to the effects that can result from evanescent housing modes. Thus, our suggestion for use in the 2D + 1 CAD optimization of filters, which are expected to have negligible coupling between the evanescent housing modes and the filter structure, is to set $f_{\rm calc}$ at a small value so that the solution is nearly quasi-static. This should effectively eliminate housing modes from the analysis and yield improved results.

The details of actual filter design are beyond the scope of this paper, and we have not tested the accuracy of the above suggested simplification on an actual filter design. However, we do believe it should be helpful in appropriate situations. This simplification would not be appropriate for the filter in Fig. 1 since that filter has a relatively large housing, thus housing modes have a significant effect, and use of $f_{\rm calc} = 0.10$ GHz would not apply. However, it is interesting to see how much the bandwidth increases in that structure in a 2D + 1 CAD analysis if we go from computing the line parameters at $f_{\rm calc} = 0.10$ GHz (no housing mode coupling) to $f_{\rm calc} = 3.95$ GHz (full inclusion of 2D+1 CAD analysis housing modes). For $f_{\rm calc} = 0.10$ GHz and $A_o = 25.6$ mm, the 10-dB down bandwidth is found to be 0.423 GHz and for $f_{\rm calc} = 3.95$ GHz , this bandwidth increases to 0.535 GHz, giving an expansion ratio of $0.535/0.423 = 1.265$. As can be seen, the housing modes implied in the 2D + 1 CAD analysis have a very substantial affect on the bandwidth.

For the actual design of filters such as in Fig. 1, which have considerable coupling to evanescent housing modes, at the outset, we do not know what value of $f_{\rm calc}$ will give the correct amount of mode coupling in order to enable us to accurately predict the bandwidth. In these situations, it appears to be necessary to use a 3-D planar CAD (or other 3-D CAD) analysis as in Swanson's original procedure in order to find out how to properly account for housing modes. It may also be useful to consider a modified form of Swanson's procedure, which seeks to produce the correct amount of coupling to housing modes in the final optimization process. Using this approach, as in Swanson's procedure, an initial design is obtained using 2D + 1 CAD optimization, and this design is then analyzed using 3-D planar CAD analysis. We then vary $f_{\rm calc}$ in the 2D + 1 CAD analysis of the structure until the passband width agrees with that computed using 3-D planar CAD analysis of the initial structure. (This is the same process as was used in the last example in Section IV, the results of which are plotted in Fig. 6.) This should give the right amount of coupling to the housing modes, but, of course, the bandwidth is narrower than desired. The final step is to reoptimize the structure for the actual desired bandwidth while using this value of $f_{\rm calc}$. This differs from Swanson's original procedure in which $f_{\rm calc}$ is set at the filter passband center, and the final optimization is for an oversized bandwidth, which is to shrink to the desired bandwidth when the filter is constructed. The suggested modifications may give increased accuracy because the final optimization to get the desired bandwidth is done while including very nearly the correct amount of coupling to the housing modes.

As this paper was in its final stages of preparation, [4] was called to our attention. Reference [4] includes two filter examples, one of which is a microstrip interdigital filter centered at about 4.9 GHz. The housing size is not discussed, but since the short circuits were generated by vias, the housing could have been considerably larger than the interdigital structure. The measured bandwidth appears to be about 7%. The measured response is presented along with responses computed using quasi-static analysis and using full-wave hybrid-mode analysis. The measured frequency bandwidth at the 10-dB down points appears to be about 18% larger than that predicted by the quasi-static analysis. This suggests that the housing was sizable, and coupling to evanescent housing modes was substantial. The hybrid-mode computed response falls between the quasi-static response and the measured response. (A discrepancy such as this could easily have resulted from the housing not being accurately modeled.)

VIII. Concluding Remarks

As can be seen from Section II and following sections, in the case of narrow-band microstrip filters involving coupled lines, housing modes have a significant affect on the passband of the filter even though the passband may be located at a frequency that is considerably below the lowest relevant housing-mode resonance. However, in the case of narrow-band filters with semilumped-element structures, housing modes may have a negligible affect on the response even if the housing is quite large. This is because the structure may have very little coupling to the housing-mode fields.

In the case of narrow-band filters with parallel-line couplings, the change in bandwidth due to coupling to evanescent housing modes may be very sizable compared to the overall passband width. However, it may be that when dealing with analogous wide-band filters, the change in the bandwidth may be small compared to the overall bandwidth, hence, not very important. In order to get a better feel as to how much housing modes might influence the $2D + 1$ CAD design of more wide-band filters, the simplified filter structure and terminations used in Section VI were modified to give a filter with a computed bandwidth of 17.8% at the 1-dB down points (with f_{calc} set at the passband center). Resetting f_{calc} to 0.01 GHz and then increasing it to 4.0 GHz (the passband center) was found to give an increase in frequency bandwidth of about 10%. (For this experiment, both bandwidths were computed at the 10-dB down points.) Thus, it appears that even for a filter of this more sizable bandwidth, design correction for housing modes can be of importance in $2D + 1$ CAD design if high accuracy is required.

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