Quasi-Planar Filters for Millimeter-Wave Applications

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(Invited Paper)

Abstract —This paper reviews a variety of quasi-planar low-pass, bandpass, and bandstop filters suitable for millimeter-wave applications. The emphasis is on ladder-shaped E-plane bandpass filters to highlight their advantages as well as limitations in terms of design, performance, and manufacturing. To extend their range of application it is suggested that E-plane filters be cascaded for better passband separation. A modified finline filter is presented to improve the performance and manufacturing of filters in waveguides below cutoff. Finally, it is shown that plated through holes can simplify filter housing fabrication and that surface-metallized composite housings are a lightweight and low-cost alternative to metal housings.

I. Introduction

FILTERS at millimeter-wave frequencies require accurate theoretical design and precision fabrication. This is necessary to avoid the difficult task of physical fine tuning at those frequencies. To meet these criteria, quasi-planar circuits have been widely used in recent years. Originally, the term *quasi-planar* was used to indicate that a microstrip or slotline circuit (planar circuit) is suspended in the E or H plane of a rectangular waveguide. In this paper we extend the definition of a quasi-planar circuit to structures with all-metal inserts, provided that the thickness of the metal sheet allows the application of photoetching techniques to realize the insert pattern. Examples of quasi-planar filter structures based on this wider definition are shown in Fig. 1. There are two classes of filters: those containing dielectric material to support a thin metallization pattern (5–35 μm thickness) on one or both sides of the substrate and those with relatively thick metallization (50–200 μm thickness) which do not require a supporting substrate as long as the insert is clamped in both broad walls of the filter housing. The first group of filters is suitable for large-scale integration of several components on the same substrate material. The latter embodies a commonly used solution for high-power signals and for stand-alone filter components, which at the same time provide lower losses due to the absence of a lossy substrate material.

In general a common feature of both filter classes is that their performance is essentially determined by the metallization pattern of the insert. Most important, however, as mentioned before, the inserts can be fabricated very accurately by photolithographic techniques, which reduces fabrication tolerances to ± 10 μm and, at the same time, allows low-cost mass fabrication. Finally, both types of inserts are clamped between the two halves of a low-cost split-block housing, which simplifies assembly and reduces machining costs to a minimum.

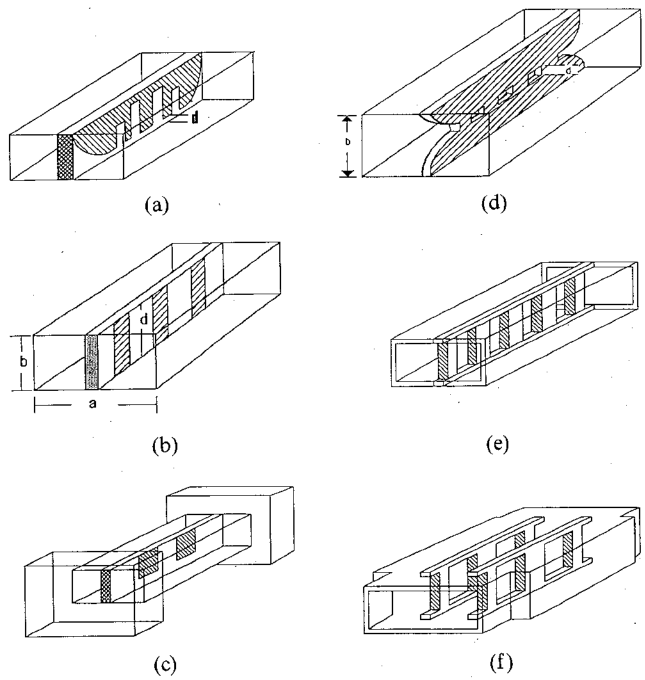

Fig. 1. Quasi-planar filter structures suitable for millimeter-wave applications.

Even though quasi-planar filters can be fabricated very accurately, the final filter realization often requires some physical fine-tuning since most of the design procedures are based on equivalent network theory. In this procedure, the filter discontinuities are represented as symmetrical T or Π networks. Several papers have been published in which the network parameters for finline discontinuities have been calculated using field theoretical methods (e.g. [19], [20], [26]–[29]). Based on the low-pass prototype design method described in [2], these networks are then represented as impedance or admittance inverters, respectively. The $ABCD$ matrices of equivalent networks of

Manuscript received January 25, 1988; revised July 28, 1988. This work was supported in part by the National Research Council of Canada under Contract (IRAP) CA 910-5-0928.
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IEEE Log Number 8824991.

0018-9480/89/0200-0324$01.00 ©1989 IEEE

subsequent discontinuities may then be cascaded to calculate the overall response of the circuit, but this procedure has certain disadvantages. First, it does not account for higher order mode (evanescent mode) interaction between subsequent discontinuities. Second, for most quasi-planar discontinuities lumped element parameters available in the open literature are calculated only for certain frequencies (for a small variety of ϵ_r, substrate thicknesses, slot and line widths, etc.) and it is impossible to derive closed-form expressions to describe their frequency-dependent variation accurately over a wide range of parameter variations. It is furthermore worth mentioning that most of the network element values which characterize a certain discontinuity are not very accurate since they are calculated with simplified field theoretical methods. Considering also that finline circuits, for example, are highly dispersive, this design method always requires circuit fine-tuning. For better design accuracy, rigorous field theoretical methods must be used to calculate the overall response of a quasi-planar filter in terms of generalized scattering parameters. This approach accounts automatically for the frequency-dependent field characteristic as well as changes in ϵ_r, substrate thickness, slot width, etc.

However, for finline circuits with reduced gap width ($d/b<1$, Fig. 1(a)) such a procedure is very complicated and requires mainframe computer power. In contrast, large-gap E-plane filters (Fig. 1(b), (e), and (f)), which are suitable for satisfying a wide range of filter requirements, may be analyzed and designed accurately using an accelerated mode-matching method, since the discontinuities involved are relatively simple and may be described rigorously by TE_{n0} modes (e.g. [6]).

The first part of this paper begins with a review of some commonly used filter structures in quasi-planar technology. In the second part the emphasis is placed on large-gap finline and metal insert filters to show not only their versatility but also their limitations in terms of out-of-band isolation, spurious passbands, and fabrication tolerances.

II. Q Factor

Appropriately shaped finline or microstrip circuits suspended parallel to the E or H field in rectangular waveguides can provide low-pass, bandpass, or bandstop performance. A number of papers have been published dealing with those structures (e.g. [3]–[9], [11]–[16], and [18]–[24]). A selection of commonly used filter inserts is given in Figs. 2, 5, and 10. Before discussing these structures in more detail, we must consider that the Q factor of finline filters, in comparison to that of waveguide cavity filters, is significantly reduced. This is particularly important for narrow-band applications of bandpass filters. Meier [1] showed in 1974 that the unloaded Q factor of a bilateral finline depends essentially on the gap width, d. The Q factor increases when d approaches the waveguide height. This effect is mainly due to the large electromagnetic field concentration in the slot region, which increases with smaller gap width. As a result, the longitudinal current per unit area of metallization surface in the slot region also increases. Therefore, losses in the gap region are higher when the gap width becomes smaller. In addition, because of the skin effect, these losses become more pronounced at higher frequencies. For all-metal insert filters the Q factor is approximately twice that of a dielectric-supported finline structure. Consequently, an all-metal insert bandpass filter can be designed for smaller bandwidths than its dielectric-supported finline counterpart.

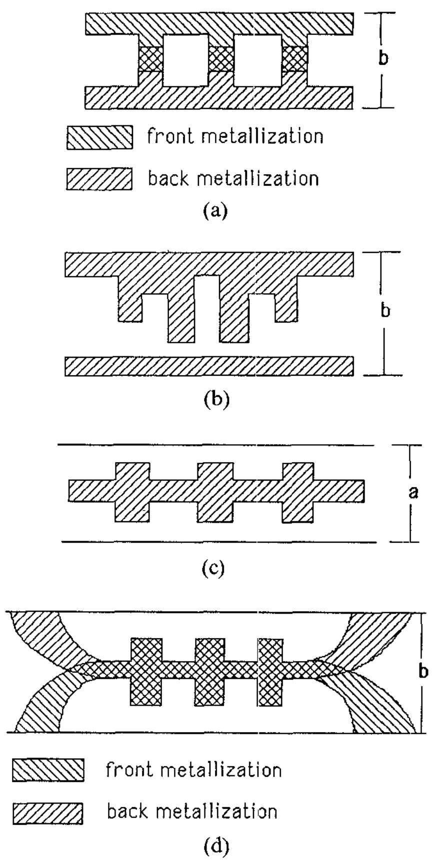

Fig. 2. Typical insert structures for low-pass filters. (a), (b), (d) E-plane circuits. (c) Microstrip on suspended substrate.

III. Review of Commonly Used Filters

A. Low-Pass Filters

In principle, low-pass filters can be designed using any structure with alternating impedance steps. However, filters in finline technology are only quasi-low-pass since any finline circuit will have a dominant-mode cutoff frequency, which occurs below that of the enclosing waveguide with no insert present. Two possible finline low-pass filters are shown in Fig. 2(a) and (b). The structure in (a) shows large impedance steps between alternating sections of overlapping (low-impedance) and nonoverlapping (high-impedance) antipodal finlines. An alternative design described in [22] uses the bilateral finline shown in Fig. 2(b). This design employs a series of approximately quarter-wavelength (actually $<\lambda_g/4$) coupled notches. The corresponding measured filter response is shown in Fig. 3. In the original work tapers were used at each end of the filter. As described in [22] the notches are regarded as

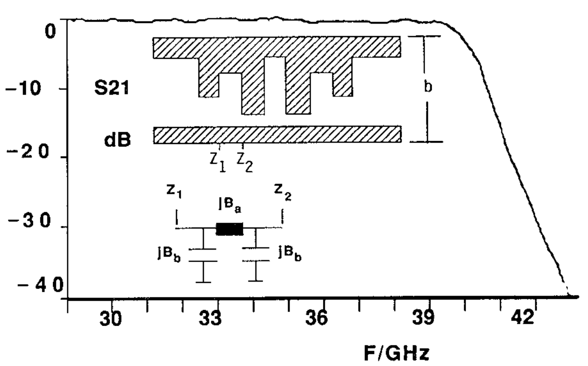

Fig. 3. Insertion loss of a Q-band E-plane bilateral finline low-pass filter (after [22]).

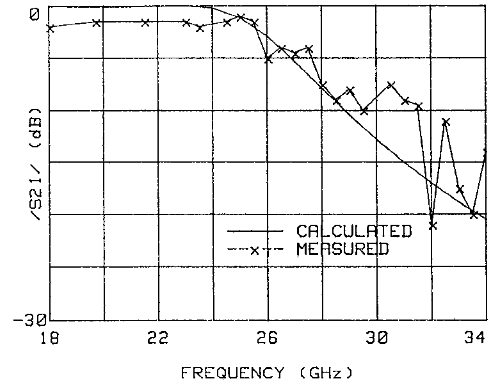

Fig. 4. Predicted and measured insertion loss of a broadside coupled K-band E-plane low-pass filter (after [24]).

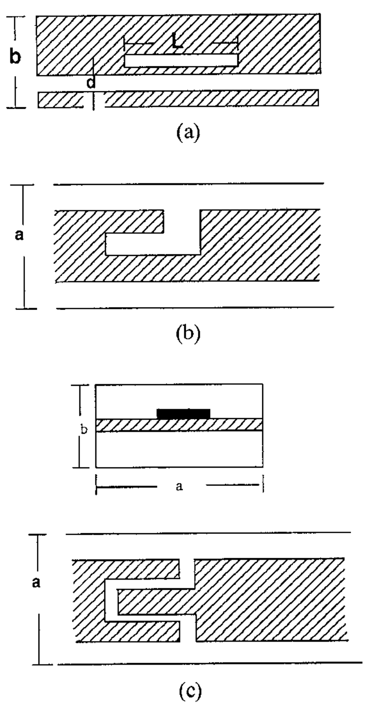

Fig. 5. Typical insert structures for quasi-planar bandstop filters. (a) E-plane structure; (b) and (c) suspended microstrip structures.

equivalent Π networks which in turn can be represented as admittance inverters. Subsequent notches were combined by an $ABCD$ matrix assuming that no higher order mode interaction occurs between the discontinuities. The accuracy of this design approach depends very much on the accuracy of the method for determining the susceptance parameters of each notch. Also, it is important to know whether higher order mode interaction occurs between the discontinuities, which would directly influence the electrical length between two notches. To reduce interaction between evanescent modes, the bilateral finline is preferred over the unilateral finline because in the latter case higher order modes may propagate at much lower frequencies. Moreover the cutoff frequency of higher order modes in unilateral finlines is significantly reduced if the mounting grooves to fix the insert exceed approximately 1/3 of the waveguide height [30]. In contrast, the symmetrical bilateral finline is virtually insensitive to this parameter and the monomode bandwidth is almost twice that of the unilateral finline.

A true low-pass structure is shown in Fig. 2(c). The suspended substrate circuit is enclosed by a waveguide with reduced dimensions, so that only a quasi-TEM (no cutoff frequency) mode can propagate. An interesting variation of this structure, published in [24], is given in Fig. 2(d). A broadside-coupled microstrip circuit is suspended in the E plane of a rectangular waveguide and is therefore compatible with other E-plane circuits. The circuit configuration is symmetrical with respect to the x and y directions. Due to the antipodal tapers from the waveguide to the broadside-coupled microstrip lines, the field is concentrated mainly between the two conductors, so that only the odd-mode impedances need to be considered. The low and high odd-mode impedances of the low-pass filter were chosen to be 14 Ω and 70 Ω, respectively. The corresponding filter dimensions were calculated by using a spectral-domain program. Measured and predicted filter responses are given in Fig. 4.

B. Bandstop Filters

The variety of bandstop filters in E-plane technology is fairly limited. This is so because most of the finline discontinuities used to design filters represent equivalent circuits consisting of T or Π networks with inductive series reactance. The shunt reactance is purely capacitive for notches or posts and purely inductive for strips. Hence, resonant circuits in series with the transmission line cannot be realized [20]. However, placing a resonant slot parallel to the input finline creates a reaction-type resonator [20]. The resulting slot pattern is shown in Fig. 5(a). The equivalent network used to analyze the filter shows a series connection of an inductive and a capacitive element in shunt with the transmission line. A filter prototype was designed by first determining the normalized network parameters with a simplified field theoretical method and then using the filter design method described in [2]. The predicted and measured filter responses (Fig. 6) show an offset of approximately 500 MHz. This kind of filter may be fine-tuned by etching a post into the resonator section. A larger variety of bandstop filters may be obtained if other than

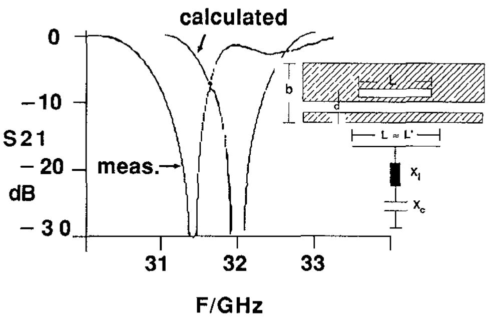

Fig. 6. Predicted and measured filter response of the E-plane bandstop structure (after [20]).

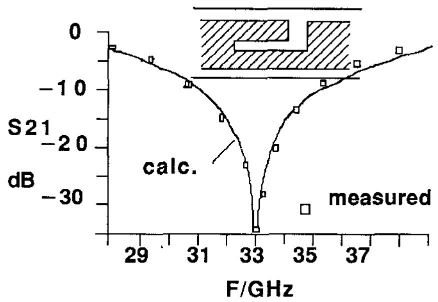

Fig. 7. Predicted and measured filter response of the suspended microstrip bandpass filter (after [21]).

E-plane circuits are taken into consideration. For example, microstrip circuits on suspended substrate (Fig. 5(b) and (c)) have been used in [21] to design spurline bandstop filters. Based on work performed by Schiffman and Matthaei [34] and by Bates [35], the authors in [21] developed a network model which accounts for the different even and odd quasi-TEM mode propagation velocities in a coupled suspended microstrip line. To suppress highly dispersive waveguide modes, the waveguide dimensions in the filter section were drastically reduced. Knowing the characteristic impedances of even and odd modes, well-known filter synthesis procedures [2] can be applied. A comparison between theoretical prediction and measurements, shown in Fig. 7, confirms the theory described in [21]. To ensure low losses outside the bandstop region a printed probe transition, similar to that shown in Fig. 8, was used. As can be seen from the measurement, this transition is fairly broad-band and can be utilized in the design of other filter structures in suspended substrate technique.

C. Bandpass Filters

Quasi-planar millimeter-wave bandpass filters using parallel or end coupled lines on suspended substrate are also feasible using the kind of transition (Fig. 8) discussed before. In the reduced size waveguide section, which accommodates the filter structure (see inset in Fig. 9), the quasi-TEM mode shows very little dispersion, and waveguide modes are below cutoff. Design techniques have been published in [15], [16], and [18] and very low insertion losses, wide passbands, and a large rejection bandwidth have been reported. As an example, the response of a seven-element parallel coupled filter for Ka-band is shown in Fig. 9 [18].

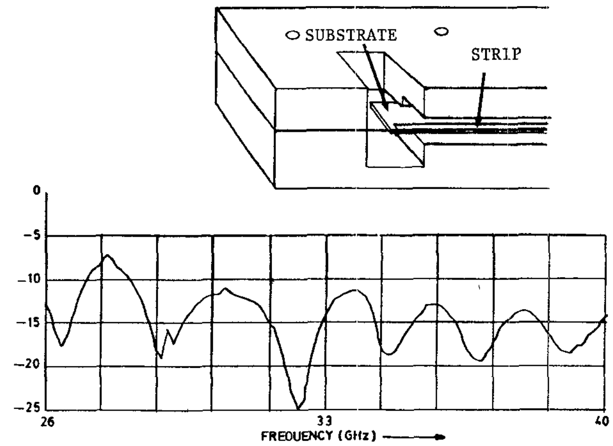

Fig. 8. Measured return loss of a printed probe transition between rectangular waveguide and suspended microstrip.

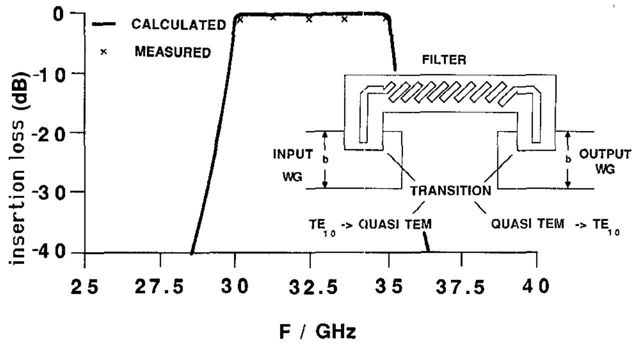

Fig. 9. Measured insertion and return loss of a Ka-band, seven-element, coupled line suspended substrate bandpass filter (after [18]).

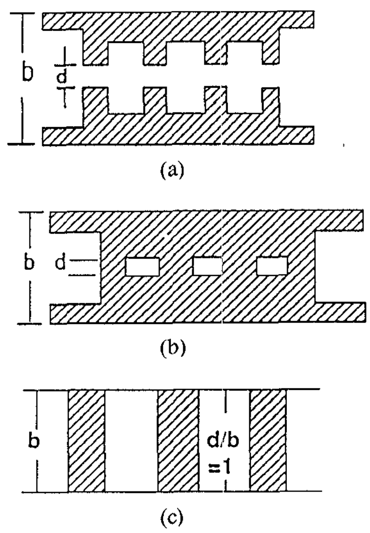

Fig. 10. Typical E-plane bandpass filter inserts.

E-plane bandpass filters are easily integrated with other E-plane structures. They are usually designed by cascading half-wave resonators. Possible slot patterns are shown in Fig. 10(a)–(c). The most commonly used are structures related to Fig. 10(b) and (c), where the half-wave resonators are realized by full-height or reduced-gap finline sections directly coupled via short conductive septa. The length of the resonators for a given center frequency depends on the gap width, dielectric constant, and substrate thickness. The equivalent network for the coupling section can be described as a T network consisting of inductive elements. Alternatively, in Fig. 10(a) the coupling section is realized by a notch which can be represented as a Π network containing an inductive series element and capacitive shunt elements.

The filter structure in Fig. 10(b), as well as the special case for $d/b=1$ (Fig. 10(c)), belongs to the class of inductive direct-coupled cavity filters for which a well-developed design theory is given by Levy [31] using impedance inverters and low-pass prototypes. Based on this design procedure, a narrow-gap unilateral finline filter was designed by Verver [23] for a 28.0 GHz center frequency. The septum lengths corresponding to the calculated T parameters were taken from [27]. Due to the thin metallic septum, the electromagnetic field penetrates into the coupling region left and right from the septum and increases the electrical resonator length. The mechanical resonator length is therefore significantly shorter than $\lambda_g/2$. To meet the required center frequency it is necessary to account for this effect. Correction values for the septum length may therefore be taken from the paper by Pic and Hoefer [25]. The measured filter response with corrected septum length is given in Fig. 11. The insertion loss averages 1.7 dB. An equivalent design using a bilateral finline with a gap width $d/b=1$ is shown for comparison. In this case the insertion loss is less than 0.6 dB with an improved filter slope. This result is mainly due to the improved Q factor of the large-gap finline filter and is partly due to the choice of a bilateral rather than a unilateral structure. The large-gap finline filter was designed using EPLANFIL, which is a field-theory-based CAD program specially designed for this kind of filter. The software calculates the filter response in terms of the generalized scattering matrix parameters and optimizes the filter dimensions. Alternatively, those filters may be designed using the equivalent network theory described by Bui *et al.* [13]. This procedure uses an explicit formula derived by Rhodes [32] to obtain the element values of a distributed low-pass prototype element. The crucial step in this procedure is determining the normalized reactances from the scattering parameters of the waveguide bifurcation of finite length. From [32] one obtains (Fig. 12)

$$jx_L = \frac{1-S_{12}+S_{11}}{1-S_{11}+S_{12}} \tag{1}$$

$$jx_p = \frac{2S_{12}}{(1-S_{11})^2-S_{12}^2} \tag{2}$$

where S_{11}, etc., are the scattering coefficients of the dominant mode. The S parameters can be calculated very accurately by using a mode-matching technique which includes the interaction effects between fundamental and higher order modes at subsequent discontinuities as well as the finite metallization thickness of the septum. Since x_L and x_p are a function of the septum length, both parameters are plotted over L in Fig. 12(a) for different metallization thicknesses and frequencies. The method described in [13] was reported to provide good practical results and in the case of a D-band finline filter the measured response deviated only 3.5 GHz from the theoretical prediction [13]. According to the authors, similar results can be obtained for all-metal insert filters. Therefore, to enlarge the data base for the practicing engineer, Fig. 12(b) provides data for the normalized reactances of a single metal insert and two different metallization thicknesses (127 μm and 50 μm).

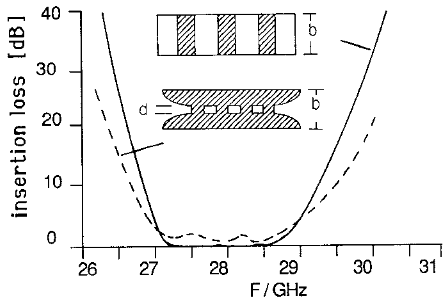

Fig. 11. Comparison between large-gap ($d/b=1$, solid line) bilateral and reduced-gap ($d/b<1$, dotted line) unilateral three-resonator finline filters in *Ka*-band. (Data for the reduced gap finline filter courtesy of Communication Research Center (K. Verver), Ottawa, Canada.)

Filter fine-tuning can be avoided and a higher degree of freedom in the design can be achieved if the filter response is calculated entirely in terms of generalized scattering parameters. The advantage over equivalent network theory is that this method accounts automatically for propagating as well as nonpropagating modes and their mutual interaction. Furthermore, the accuracy of the method is not affected by changing the frequency or the structural dimensions (i.e., metallization thickness) from standard to nonstandard values. It is true that the computation time of this method is considerable compared to network analysis methods. On the other hand, fine-tuning a filter at W- or D-band is a time-consuming and expensive task. If this can be avoided by using an accurate design program which instead occupies inexpensive CPU time on a personal computer (a typical filter design using EPLANFIL requires less than 50 minutes on a PC), the cost for a filter design can be reduced significantly. For large-gap finline and metal insert filters, rigorous design methods have been reported in [5]–[9], [11], and [12]. The author has modified the theory published in [5]–[9]. This led to a significantly accelerated algorithm which is now implemented in the CAD filter software EPLANFIL. The following analysis and designs are performed by using this software.

IV. Large-Gap Finline Filters

A conventional bilateral finline bandpass filter ($d/b=1$) designed for *Ku*-band is shown in Fig. 13. Although a comparison with unilateral large-gap finline filters shows equivalent performance in theory, measurements have

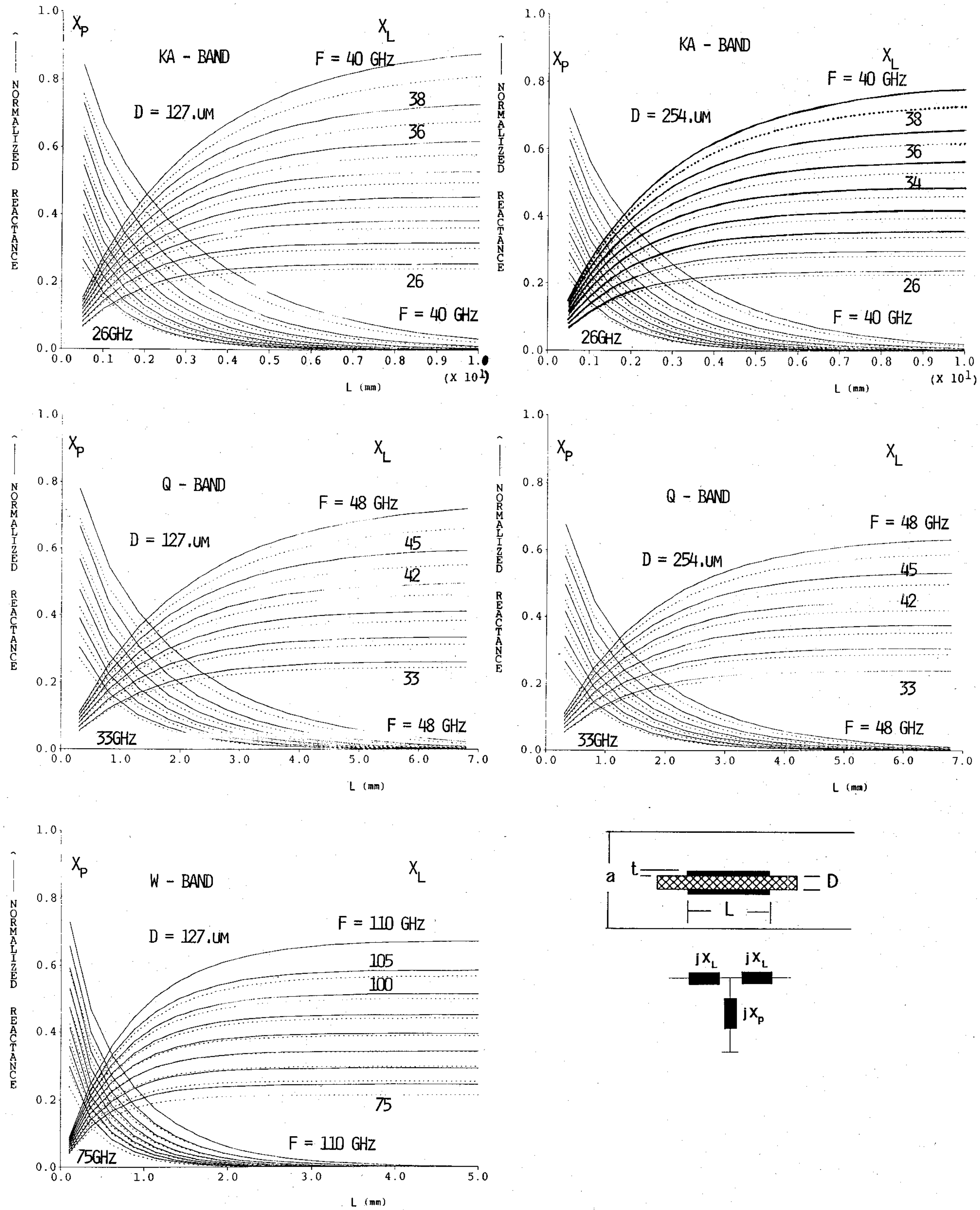

Fig. 12. (a) Equivalent T network of the coupling section in a bilateral finline filter and the normalized reactances versus septum length L and frequency in *Ka*-band, *Q*-band, and *W*-band. $\epsilon_r = 2.22$, (———) $t = 17.5\ \mu$m, (– – – –) $t = 35\ \mu$m. *(Continued on next page.)*

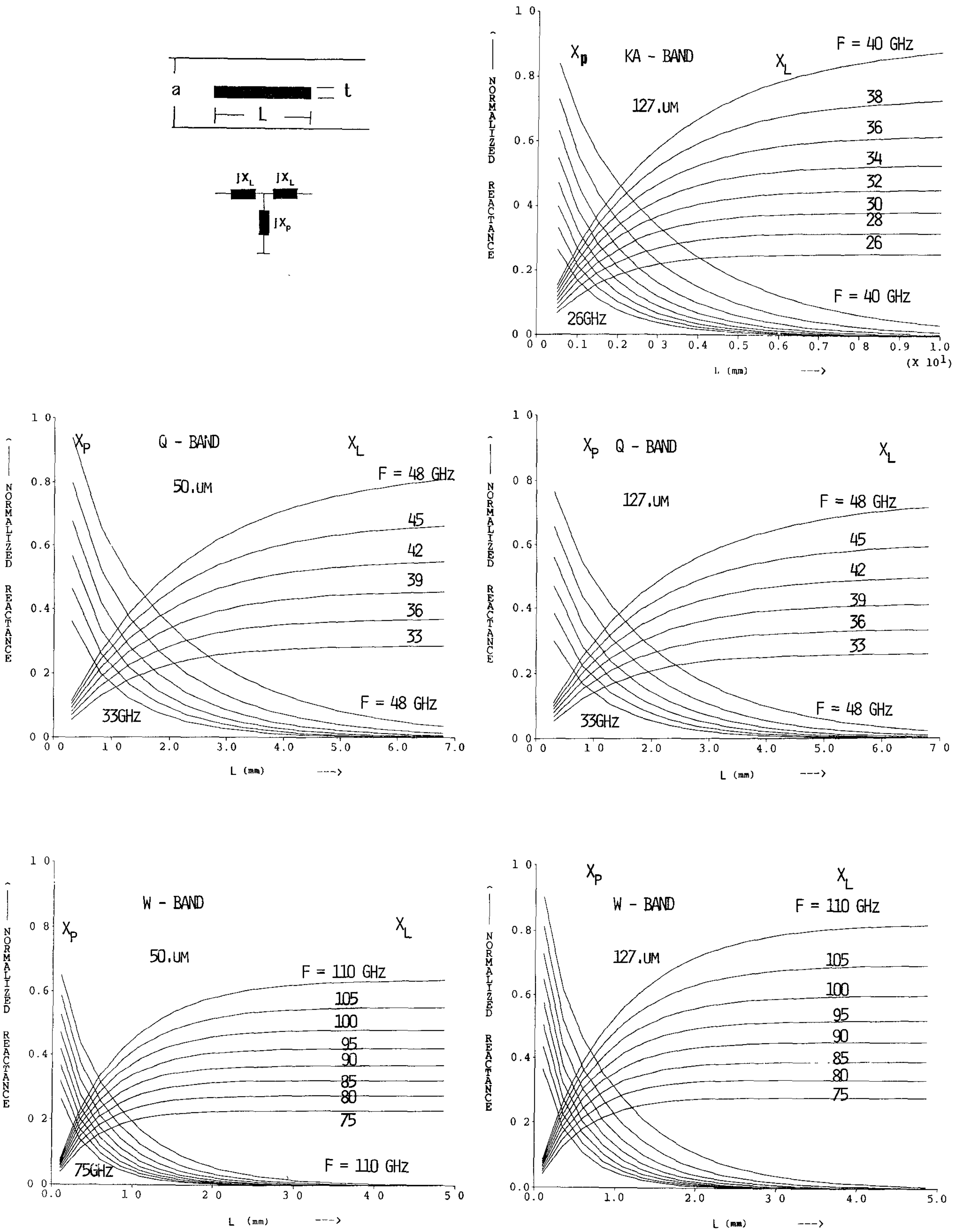

Fig. 12. (*Continued*) (b) Equivalent T network for the coupling section in single metal insert filter. Normalized reactances versus septum length L and frequency in *Ka*-band, *Q*-band, and *W*-band.

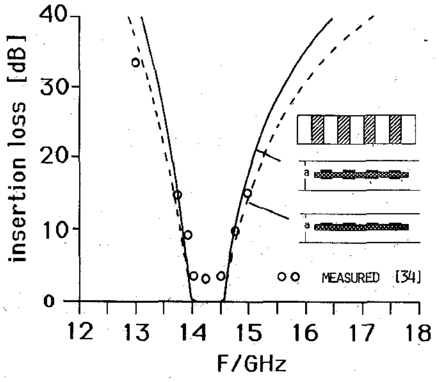

Fig. 13. Comparison between a three-resonator bilateral and unilateral large-gap finline filter in *Ku*-band ($\epsilon_r = 2.22$, $d/b = 1$).

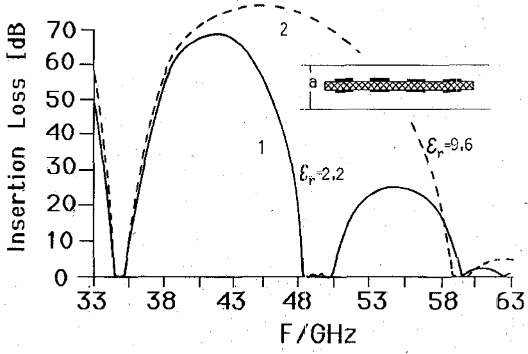

Fig. 15. Three-resonator *Q*-band finline filter designed for the lower end of the waveguide band, $d/b = 1$, $a = 5.69$ mm, solid line $\epsilon_r = 2.22$, dotted line $\epsilon_r = 9.6$.

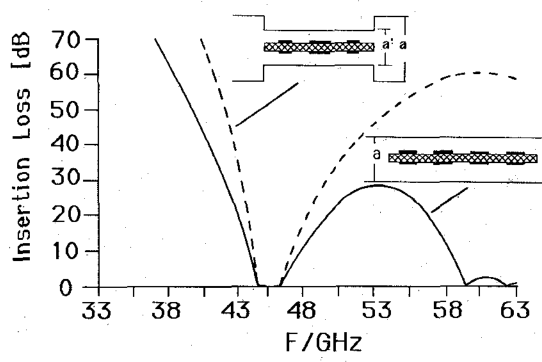

Fig. 14. Three-resonator *Q*-band finline filter designed for the band end, $\epsilon_r = 2.22$, $d/b = 1$ (solid line $a = 5.69$ mm, dotted line $a' = 4.3$ mm).

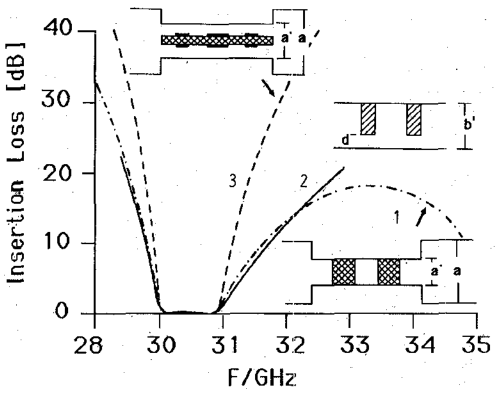

Fig. 16. Comparison between a two-resonator conventional dielectric resonator filter ($\epsilon_r = 10.5$, $-\cdot-\cdot-$), a filter with nontouching fins ($\epsilon_r = 2.2$, $d/b < 1$, solid line) after [36], and a modified ladder-shaped finline filter ($\epsilon_r = 10.5$, $d/b = 1$, dotted line).

shown that the response of the unilateral filter shows 3 dB insertion loss over 0.3 dB for the bilateral filter. One explanation for this response may be that unilateral structures are more sensitive to the groove depth for fixing the insert in the waveguide mount [30]. These effects are not included in the filter design theory and are negligible for bilateral filters. Furthermore, due to a higher energy concentration in the substrate region in one of the coupling subsections, losses imposed by the substrate material have a greater influence on the overall losses. It is therefore advisable to use bilateral structure for finline filters whenever possible.

At the end of a waveguide band the second stopband attenuation of finline filters is significantly reduced due to the onset of parasitic modes propagating in the coupling subsections (Fig. 14). This effect has been discussed recently by Vahldieck and Hoefer [7]. On the other hand, filters for the lower end of a waveguide band have rather long resonators because of the rapid increase of the guide wavelength toward cutoff. Therefore, harmonic resonances can occur within the single mode range of the waveguide band (Fig. 15, curve 1). To alleviate the first problem it has been suggested in [7] to reduce the width of the housing in the filter section. This measure increases the cutoff frequency of the fundamental and higher order modes in the coupling subsections and, at the same time, increases the passband separation and second stopband attenuation significantly (Fig. 14). To increase the passband separation of finline filters for the lower end of a waveguide band, the solution is quite different. In this case we need to shorten the resonator length, which can be done by increasing the relative permittivity of the supporting substrate material or by widening the waveguide section of the filter. The effect of the latter measure will be shown in conjunction with metal insert filters. Fig. 15 (curve 2) shows the resulting filter response using a high-dielectric-constant material. The spurious harmonic passband is pushed up in frequency and the second stopband attenuation is improved significantly.

V. Modified Finline Filters

Dielectric resonator filters in waveguide sections operating below cutoff are an attractive solution when extremely compact filter components are required (Fig. 16, curve 1) [17]. To avoid the expensive machining of the resonator blocks required for the conventional design, finline filters with nontouching fin sections have been suggested [36] to take advantage of high-precision photolithographic fabrication techniques. In [36] the authors have utilized the spectral-domain technique to calculate the response of a two-resonator filter in *Ka*-band (Fig. 16, curve 2). In this configuration the nontouching fin sections are the resonators which support the fundamental mode above cutoff. Evanescent-mode coupling is obtained via the waveguide sections with large-gap fins ($d/b=1$). The disadvantage of this structure is that the Q factor is significantly reduced. To alleviate this problem a modified, ladder-shaped finline filter is used in a waveguide section below cutoff (Fig. 16, curve 3). In contrast to conventional finline filters a thick, high-permittivity substrate material is used to lower the cutoff frequency of the fundamental TE_{10} mode in the resonator section. These filters show excellent stopband attenuation and increased passband separation and are much shorter than conventional dielectric resonator filters as well as conventional finline filters in waveguides operating above cutoff.

VI. All-Metal Insert Filters

For high-power applications the metallization thickness of the filter inserts must be increased. On the other hand, the metallization thickness on commercially available substrate material is typically 17.5 μm. For thicker metallization, pure metal sheets (e.g. 99.9 percent copper) must be used, which are available in the range between 50 and 508 μm or more. As long as the metallization thickness remains below approximately 200 μm, the insert is still amenable to photolithographic techniques. Most of the limitations discussed in conjunction with finline bandpass filters will remain the same for metal insert filters and need not be discussed again. However, overcoming some of the problems discussed for finline filters requires different solutions. As for finline filters, all-metal insert filters show improved performance toward the end of a waveguide band by placing the ladder-shaped insert in a narrower waveguide section and connecting both ends to the standard waveguide by a simple inductive step transition without complicating the mechanical design of the component [7]. Filters for the lower end of the waveguide band show a harmonic passband within the single mode range of the waveguide band. For metal insert filters the only choice in shortening the resonator length is to widen the filter waveguide section such that the first passband is now centered in this single-mode range of that section [7]. By inserting the filter section between the two standard waveguide flanges, we are able to push the spurious passband up in frequency without complicating the mechanical design of the filter component. To suppress parasitic modes in the wider coupling subsections we must now use twin metal inserts [7]. For even better performance, triple metal inserts may be used [8]. Fig. 17 shows a comparison between a conventional single metal insert filter in a standard waveguide housing and twin and triple metal insert filters in a wider waveguide section. E-plane filters may be cascaded [9] to provide continuous attenuation between f_0 and $2f_0$ by manipulating the inevitable spurious responses of the individual filters such that they overlap and cancel each other as shown in Fig. 18, curve 3. For a constant first passband, one filter may be placed in a wider waveguide housing, which means a shorter resonator length and increased passband separation (Fig. 18, curve 1), whereas the second filter is placed in a narrower waveguide housing, which increases the resonator lengths and leads to a shrinking passband separation (Fig. 18, curve 2).

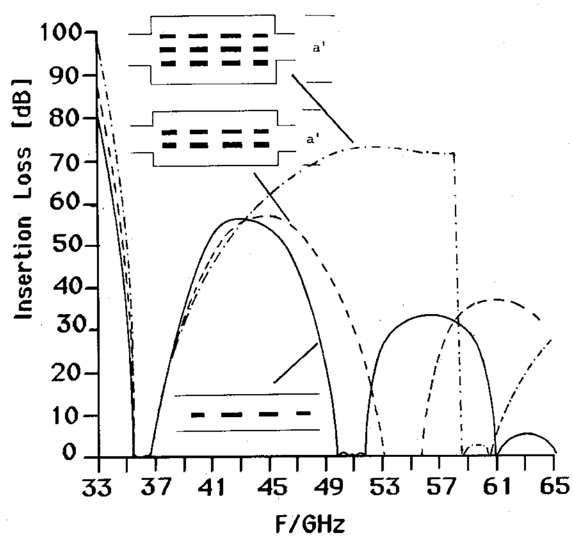

Fig. 17. Four-resonator all-metal insert filters designed for the lower end of the waveguide band (Q-band). (———): conventional E-plane ladder-shaped single metal insert filter. (– – – –): Twin metal insert filter in a wider waveguide housing, $a'=6.7$ mm. (–·–·–·–·–·): Triple metal insert filter in a wider waveguide housing, $a'=6.9$ mm, $t=127$ μm.

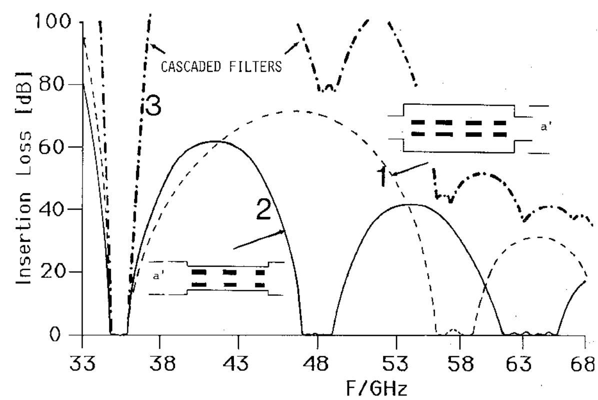

Fig. 18. Example of a Q-band eight-resonator cascaded twin metal insert filter prototype, curve 3. Curve 1 and curve 2 show the filter responses of the individual four-resonator filters in wider and narrower waveguide housings.

VII. Practical Aspects

The typical bandwidth achievable with large-gap E-plane filters (metal insert as well as finline) ranges from 0.5 percent to 20 percent. However, for a bandwidth smaller than 1 percent the insertion loss will increase because the Q factor is not sufficiently high. Typical insertion loss values can be expected from 0.5 to 1 dB for frequencies below 60 GHz and from 1 to 2 dB for frequencies up to 140 GHz (depending on the bandwidth). The return loss is typically between 15 and 25 dB.

The influence of housing tolerances on the filter performance of ladder-shaped ($d/b=1$) finline or metal insert filters is more significant than in the case of reduced-gap ($d/b<1$) finline filters. This is so because in the latter

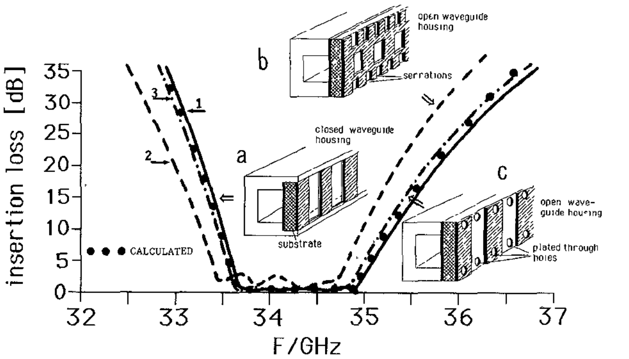

Fig. 19. Measured performance of a four-resonator ladder-shaped finline filter in a *Ka*-band waveguide mounted in three different ways. Curve 1: closed waveguide housing; curve 2: open waveguide housing with serrations; curve 3: open waveguide housing with plated though holes.

configuration most of the electromagnetic field in the resonator section is concentrated in the gap region, which reduces the required machining accuracy for the housing considerably. Therefore, machining tolerances required for ladder-shaped *E*-plane filter housings are tighter. For example, if the waveguide width of a *W*-band filter housing is reduced by only 10 μm the center frequency (95 GHz) of the filter may shift up to 2 GHz toward higher frequencies. Also, the thickness of the metal insert affects the filter performance significantly. Changing the thickness of the metal sheet from 127 μm to 50 μm may reduce the center frequency of a *W*-band filter by 3 GHz. These effects have been analyzed theoretically using EPLANFIL. To preserve the accuracy of the theoretical design when using EPLANFIL (which is typically within 0.3–0.5 percent of the measured response when the following suggestions are considered), it is necessary to fabricate the filter housing first to the appropriate length, measure the actual dimensions of the housing as well as the insert thickness, and then take these values as input for the design procedure. One fabrication parameter which is less predictable is the etching tolerances, which are typically ± 10 μm unless sputtering techniques are used. A worst-case analysis performed for a *W*-band metal insert filter has shown that by shortening all resonator lengths and increasing all coupling section lengths by 10 μm, the resultant frequency shift was 200 MHz at 95 GHz center frequency. Furthermore, the insertion loss increased from 1.1 dB to 2.0 dB.

For finline filters the choice between an open or a closed waveguide housing is very important. Experience has shown that an open waveguide housing, in which the insert is simply sandwiched between two waveguide halves, is not suitable for large-gap finline filters. Even with serrations, radiation losses have been found to be too high, resulting in higher insertion loss and a shift in midband frequency. The closed waveguide housing seemed to be the only alternative, and was used extensively in the past, providing excellent results. However, the machining effort for milling a small slit in the waveguide mount to fix the filter insert in the waveguide enclosure (Fig. 19(a)) and to provide at the same time good electrical contact with the housing without deforming the insert substrate is time consuming and expensive. To solve this problem, the following solution seems to be adequate because the required accuracy of the milling process is reduced to a minimum. In this solution we use the open waveguide housing to accommodate the filter insert. However, to prevent radiation effects, the serrations on the substrate (Fig. 19(b)) are replaced by plated through holes to establish a short circuit between both waveguide halves (Fig. 19(c)). If the distance between the plated through holes is smaller than approximately $\lambda_g/8$ (λ_g corresponds to the highest frequency of interest), the waveguide is virtually closed and no radiation will occur. A comparison between conventional ladder-shaped finline filters designed for 34.25 GHz, mounted in three different ways, is shown in Fig. 19. It can be seen that the filter response of the device in a closed waveguide housing as well as in the open waveguide housing with plated through holes comes closest to the predicted filter response (0.3 percent deviation) calculated with EPLANFIL. It is furthermore noticeable that the filter with the plated through holes was incorporated into a metallized plastic housing. The prototype plastic waveguide was machined to the required dimensions (less 40 μm) and then metallized up to a 20 μm copper thickness by using sputtering techniques. A comparison between the same filter in a metal housing shows identical performance. Since the plastic material used for the filter housing is suitable for injection molding, this approach offers tremendous potential for mass production of low-weight and low-cost filter components.

VIII. Conclusion

This paper presented an overview of quasi-planar filters suitable for millimeter-wave applications. The shortcomings of certain filter structures have been discussed and possible solutions to improve filter performance have been introduced. The emphasis of the discussion for bandpass filters was on ladder-shaped *E*-plane structures, and their versatile applications were demonstrated. To analyze and design large-gap finline and metal insert filters, the field-theory-based CAD software EPLANFIL was used, which is fully operational on personal computers. It was shown that cascading filters improves the passband separation of *E*-plane filters. Furthermore, enclosing ladder-shaped finline filters in waveguides below cutoff provides better performance than conventional dielectric resonator filters or filters with nontouching fins. Finally, finline filters in metallized plastic housings have been tested. Results have shown performance equivalent to that of metal housings. However, plastic housings are significantly lighter and they are suitable for low-cost injection molding. Thus metallized plastic housings offer a tremendous potential for future lightweight, low-cost, mass-producible microwave and millimeter-wave components.

Acknowledgment

The author gratefully acknowledges the support of M. Leitner and J. Ruxton, BTI Inc., Carleton Place, Ont.,

Canada, which enabled him to study certain manufacturing aspects of E-plane filters, especially in conjunction with composite wavegiude housings.

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