Optimized Waveguide *E*-Plane Metal Insert Filters for Millimeter-Wave Applications

RÜDIGER VAHLDIECK, JENS BORNEMANN, FRITZ ARNDT, AND DIETRICH GRAUERHOLZ

Abstract —**A design theory is described for rectangular waveguide metal insert filters that includes both higher order mode interaction and finite thickness of the inserts. Optimized design data for three- to five-resonator type filters with several insert thicknesses suitable for metal stamping and etching techniques are given for midband frequencies of about 15, 33, 63, and 75 GHz. Measured passband insertion losses of prototypes for midband frequencies of 15, 33, and 76 GHz are 0.2, 0.6, and 0.7 dB, respectively.**

I. Introduction

IN *E*-PLANE circuits supporting dielectrics cause additional losses. It may therefore often be advantageous to restrict the design for high-Q millimeter-wave circuits with a certain degree of integration, such as converters [1] and diplexers [2], to pure metal inserts (Fig. 1) placed in the *E*-plane of rectangular waveguides without any substrates [3]–[6]. This paper describes a numerical synthesis procedure for such metal insert filters. The given design data allow metal etching techniques appropriate for low-cost mass production.

The two- and three-section X-band inductive strip filters described in [4], [5] are calculated by an equivalent-circuit approach. However, the immediate higher order mode coupling between the discontinuities, which reduces the stopband attenuation for higher frequencies, is not taken into account. This effect is considered in [6], where three- and five-section Ku-band filter curves are calculated by an equivalent waveguide method. But the influence of the finite thickness of the metal inserts, which influences passband ripple behavior and midband frequency, is neglected.

In this paper, similar to the fin-line filter calculation in [7], the design of optimized metal insert filters is based on field expansion directly into incident and scattered waves of interest. This allows direct inclusion of both higher order mode coupling and finite strip thickness. Moreover, matching the fields at common interfaces yields the corresponding scattering matrix.

A simple computer program varies the filter parameters until the insertion loss within the passband yields a minimum and the stopband attenuation an optimum. The evolution strategy method [8] is applied where no differentiation step in the optimization process is necessary, which reduces the involved computing time. Data for optimized Ku-band, Ka-band, V-band, and E-band filters are given. The measured frequency response, of the metal insert filters that have been designed and operated at 15, 33, and 76 GHz, shows good agreement between theory and experimental results.

Manuscript received May 28, 1982; revised July 27, 1982.
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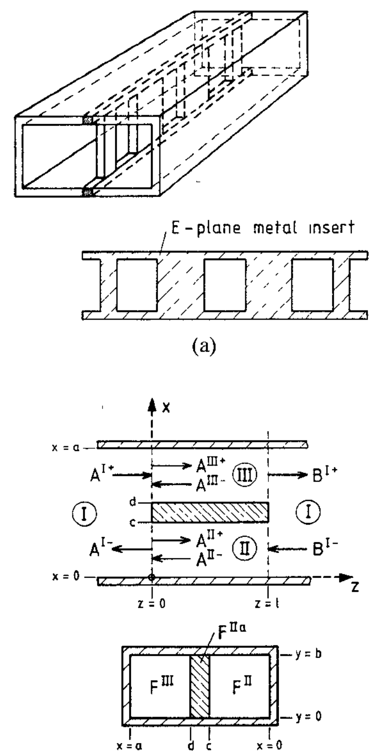


Fig. 1. *E*-plane metal insert filter without supporting dielectrics. (a) Waveguide and metal insert. (b) Transition waveguide to metal bar and cross section with common interfaces.

II. Theory

As in [7] for each subregion $\nu = \mathrm{I, II, III}$ (Fig. 1(b)), the fields [9]

$$\boldsymbol{E}^{(\nu)} = -j\omega\mu\nabla \times \Pi_{hx}^{(\nu)} \quad \boldsymbol{H}^{(\nu)} = \nabla \times \nabla \times \Pi_{hx}^{(\nu)} \tag{1}$$

are derived from the x-component of the magnetic Hertzian potential Π_h which is assumed to be a sum of suitable eigenmodes [9] satisfying the vector Helmholtz equation

$$\nabla^2 \Pi_h + k^2 \Pi_h = 0 \tag{2}$$

and the boundary conditions at the metallic surfaces at

$x = 0, c, d, a$ (Fig. 1(b))

$$\Pi_{hx}^{(\nu)} = \sum_{m=1}^{\infty} A_m^{(\nu)\pm} \cdot T_m^{(\nu)} \cdot \sin\left[\frac{m\pi}{p^{(\nu)}} \cdot f^{(\nu)}\right] \cdot e^{\mp jk_{zm}^{(\nu)} \cdot z}. \quad (3)$$

$A_m^{(\nu)\pm}$ are the still unknown eigenmode amplitudes of the forward and backward waves which are suitably normalized by $T_m^{(\nu)}$, so that the power carried by a given wave is proportional to the square of the wave-amplitude coefficients. This will lead directly to the desired scattering parameters. The notations $p^{(\nu)}$, $f^{(\nu)}$, $k_{zm}^{(\nu)}$, and $T_m^{(\nu)}$ are explained in the Appendix.

By matching the tangential field components at the common interfaces F^{II}, F^{III}, $F^{\mathrm{II}a}$ (Fig. 1(b)) across the step discontinuity at $z = 0$

$$E_y^{\mathrm{I}} = \begin{Bmatrix} E_y^{\mathrm{II}} & (x, y) \in F^{\mathrm{II}} \\ E_y^{\mathrm{III}} & (x, y) \in F^{\mathrm{III}} \\ 0 & (x, y) \in F^{\mathrm{II}a} \end{Bmatrix}$$

$$H_x^{\mathrm{I}} = \begin{Bmatrix} H_x^{\mathrm{II}} & (x, y) \in F^{\mathrm{II}} \\ H_x^{\mathrm{III}} & (x, y) \in F^{\mathrm{III}} \end{Bmatrix} \quad (4)$$

the coefficients $A_m^{(\nu)\pm}$ in (3) can be related to each other after multiplication with the appropriate orthogonal function, which leads to the corresponding coupling integrals given in the Appendix. This yields the three-port scattering matrix $(S)_{z=0}$ at the step discontinuity $z = 0$ (Fig. 1(b))

$$\begin{pmatrix} A^{\mathrm{I}-} \\ A^{\mathrm{II}+} \\ A^{\mathrm{III}+} \end{pmatrix} = (S)_{z=0} \cdot \begin{pmatrix} A^{\mathrm{I}+} \\ A^{\mathrm{II}-} \\ A^{\mathrm{III}-} \end{pmatrix}. \quad (5)$$

The step discontinuity at $z = l$ (Fig. 1(b)) can be treated in a similar manner. The overall two-port scattering matrix (S) of the discontinuity waveguide to section of metal E-plane bar and back to waveguide is given by

$$\begin{pmatrix} A^{\mathrm{I}-} \\ B^{\mathrm{I}+} \end{pmatrix} = (S) \begin{pmatrix} A^{\mathrm{I}+} \\ B^{\mathrm{I}-} \end{pmatrix} \quad (6)$$

where the coefficients of the scattering matrix are explained in the Appendix.

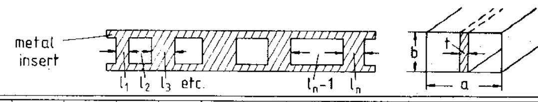
The scattering matrix of the total metal insert filter is then calculated by directly combining the single scattering matrices like in [10]. Compared with the commonly used multiplication of transmission matrices, this procedure preserves numerical accuracy, since the direct combination of scattering matrix parameters contains exponential functions with only negative argument.

For computer optimization, the expansion into nine eigenmodes at each discontinuity has turned out to be sufficient. The final design data are proved by expansion into 15 eigenmodes.

III. Design

E-plane metal insert filters with three, four, and five resonators for midband frequencies of about 15, 33, 63, and 75 GHz are chosen for design examples (Table I). The corresponding waveguide housings are WR 62 (Ku-band), WR 28 (Ka-band), WR 15 (V-band), and WR 12 (E-band). The metal insert thicknesses are chosen to be $t = 1$, 0.9 mm, 0.5 mm (for filing or metal stamping techniques), and 100 μm (for metal etching techniques).

TABLE I
Computer Optimized Design Data for E-Plane Metal Insert Filters
($n = 2n_{\mathrm{res}} + 1$, n_{res} = Number of resonators)



Frequency-band waveguide housing	Number of resonators	Insert thickness t (mm)	$l_1=l_n$ (mm)	$l_2=l_{n-1}$ (mm)	$l_3=l_{n-2}$ (mm)	$l_4=l_{n-3}$ (mm)	$l_5=l_{n-4}$ (mm)	$l_6=l_{n-5}$ (mm)	Midband frequency (GHz)	3 dB-band-width (MHz)	Max. passband ripple (dB)	Measured results see
Ku - Band	3	1.0	2.995	10.112	9.991	10.136			14.88	160	0.02	
a = 15.799 mm	3	0.9	2.238	10.504	8.915	10.548			14.57	193	0.001	Fig. 4
b = 7.899 mm	4	1.0	2.67	10.032	8.66	10.061	9.404		14.93	240	0.2	
(WR 62)	5	1.0	2.679	10.036	8.59	10.064	9.357	10.064	14.92	245	0.29	
Ka - Band	3	0.5	1.352	4.656	4.754	4.666			32.87	255	0.001	
a = 7.112 mm	3	0.5	1.337	4.658	4.507	4.668			32.87	310	0.018	
b = 3.556 mm	3	0.51	1.009	4.778	3.87	4.796			32.52	428	0.001	Fig. 5
(WR 28)	4	0.5	1.351	4.658	4.247	4.669	4.565		32.87	370	0.23	
	5	0.5	1.351	4.652	4.348	4.663	4.755	4.663	32.88	330	0.15	
	5	0.5	1.346	4.655	4.388	4.665	4.796	4.665	32.88	325	0.1	
V - Band	3	0.1	0.713	2.24	2.263	2.247			62.47	1190	0.035	
a = 3.76 mm	4	0.1	0.695	2.238	2.105	2.744	2.298		62.5	1400	0.13	
b = 1.88 mm	5	0.1	0.704	2.226	2.102	2.231	2.304	2.229	62.65	1420	0.33	
(WR 15)												
E - Band	3	0.1	0.613	1.911	1.978	1.917			75.14	1070	0.022	Fig. 6
a = 3.1 mm	4	0.1	0.617	1.923	1.915	1.928	2.1		75.0	1100	0.059	
b = 1.55 mm	5	0.1	0.617	1.92	1.906	1.927	2.103	1.928	75.0	1100	0.27	
(WR 12)												

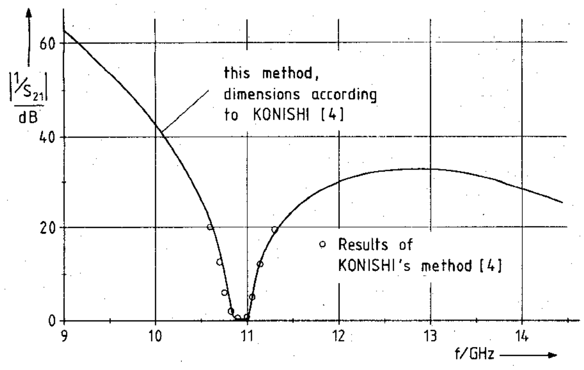
An optimizing computer program is used based on the evolution strategy method [8], which requires no differentiation step in the optimization process. The parameters l_1 to l_n (resonator dimensions, Table I) were varied for roughly fixed midband frequencies and 3-dB bandwidths, as well as for given waveguide housing dimensions, number of resonators, and metal insert thicknesses, until the insertion loss within passband yielded a minimum and the stopband attenuation an optimum. The total time for the optimization of one set of filter parameters was about 10–30 min with a Siemens-7880 computer.

IV. Results

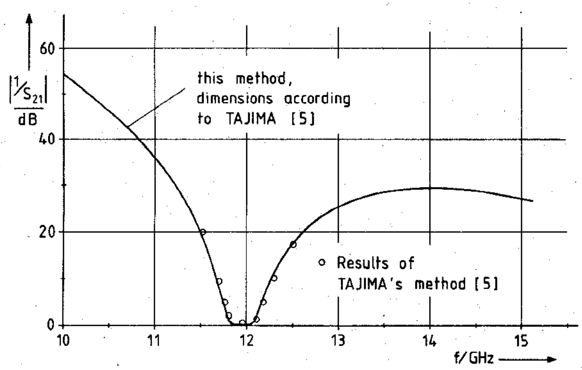
Fig. 2 shows insertion loss curves of the filters according to Konishi *et al.* [4] and Tajima *et al.* [5] calculated by our method (solid line) compared with available results of the equivalent-circuit approach of [4] and [5]. As expected, the given near-midband frequency results of [4], [5] agree well with our results in this frequency range. It is assumed, however, that for other frequencies the difference would be more obvious since the higher order mode interaction of the inductive strips, included in our method, influences above all the stopband behavior of the filter.

The influence of the finite thickness of the strips is demonstrated in Fig. 3 at the filter example given by Saad *et al.* [6]. The results of our method and of Saad's method, where the thickness influence is neglected, agree to some extent for an assumed thickness of $t = 0.3$ mm. For $t = 1$ mm, however, the difference between the two results is quite evident.

Fig. 4 shows the calculated and measured insertion loss $(1/|S_{21}|)$ in decibels as a function of frequency for a three-resonator Ku-band E-plane metal insert filter with an insert thickness of 0.9 mm (Table I). The calculated mini-



(a)



(b)

Fig. 2. Comparison of insertion-loss results calculated by this method and by the equivalent-circuit approach [4], [5]. (a) Two-resonator metal insert filter of Konishi *et al.* [4], data according to [4]. ——: this method; ○○○○: Konishi's method (graphical reproduction from [4]). (b) Three-resonator metal insert filter of Tajima *et al.* [5], data according to [5]. ——: this method; ○○○○: Tajima's method (graphical reproduction from [5]).

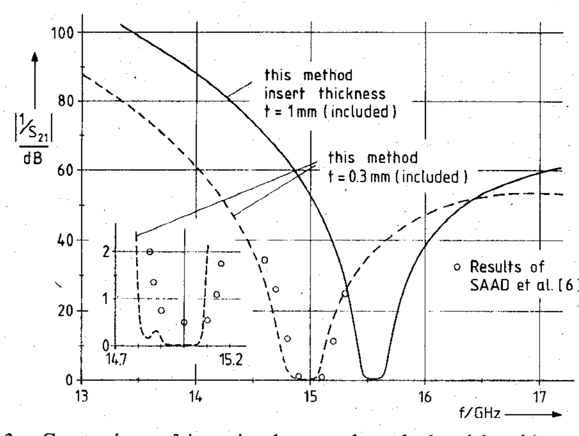


Fig. 3. Comparison of insertion-loss results calculated by this method (which includes the finite insert thickness t) and by Saad's method [6] (which neglects that influence). ——: $t = 1$ mm, this method, ---------: $t = 0.3$ mm, this method, and ○○○○: Saad's method (graphical reproduction from [6]).

mum insertion loss in passband is 0.001 dB, the measured value is about 0.2 dB.

The corresponding insertion loss values of a *Ka*-band filter with an insert thickness of 0.51 mm (Table I) are shown in Fig. 5. The calculated minimum insertion loss in passband is 0.001 dB, the measured value is about 0.6 dB.

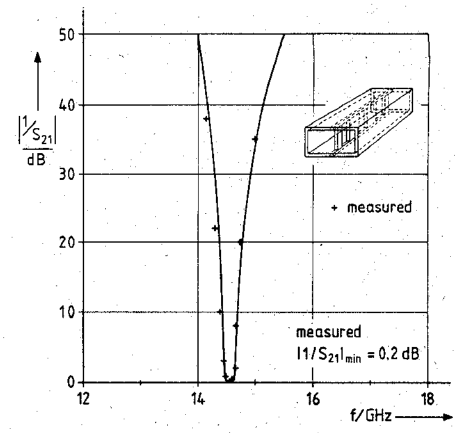


Fig. 4. Calculated and measured insertion loss of a *Ku*-band metal insert filter (Table I).

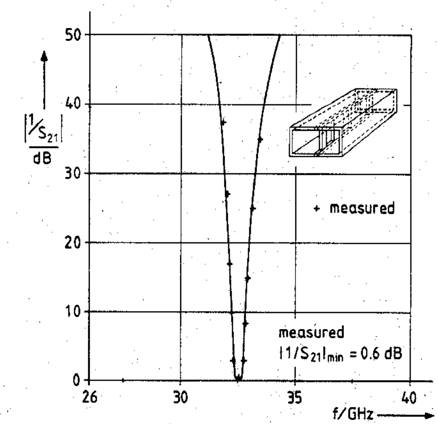


Fig. 5. Calculated and measured insertion loss of a *Ka*-band metal insert filter (Table I).

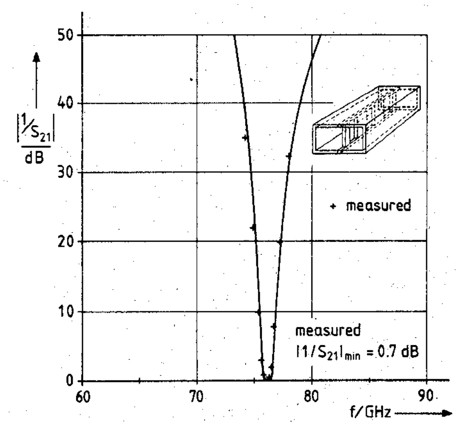


Fig. 6. Calculated and measured[1] insertion loss of an *E*-band metal insert filter (Table I).

Fig. 6 shows the calculated and measured[1] insertion loss for an *E*-band filter. The insert thickness is 0.1 mm, the resonator structure has been produced by metal etching. The calculated minimum insertion loss is about 0.02 dB,

[1] The waveguide housing dimensions of the experimental *E*-plane filter deviated from the WR 12 dimensions by about -55 μm as has been checked by a measuring microscope. This leads to a frequency shift from 75 to 76 GHz taken into account for the theoretical curve in Fig. 6.

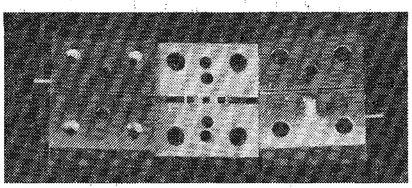


Fig. 7. Photograph of the metal-etched E-band filter structure (Fig. 6).

versus the measured value of about 0.7 dB. The photo-etched filter structure of the E-band filter is shown in Fig. 7.

V. Conclusion

A design theory has been described for optimum low-insertion loss E-plane (inductive) metal insert filters. The theory includes higher order mode interaction and the finite thickness of the insert.

Three- to five-resonator filters are chosen for design examples for Ku-, Ka-, V-, and E-band. The filter structures can be produced by metal etching techniques. The low-insertion loss design leads to measured passband insertion losses of about 0.2, 0.6, and 0.7 dB for midband frequencies of about 15, 33, and 76 GHz, respectively. The measured results verify the theory.

Appendix

Abbreviations in (3)

Variable $f^{(\nu)}$ and constant $p^{(\nu)}$ according to subregion $\nu = \mathrm{I, II, III}$

$$\begin{bmatrix} f^{(\mathrm{I})} \\ f^{(\mathrm{II})} \\ f^{(\mathrm{III})} \end{bmatrix} = \begin{bmatrix} x \\ x \\ a-x \end{bmatrix}$$

$$\begin{bmatrix} p^{(\mathrm{I})} \\ p^{(\mathrm{II})} \\ p^{(\mathrm{III})} \end{bmatrix} = \begin{bmatrix} a \\ c \\ a-d \end{bmatrix}.$$

Propagation factor k_z

$$k_{zm}^{(\nu)2} = k^{(\nu)2} - \left(\frac{m\pi}{p^{(\nu)}}\right)^2$$

with

$$k^{(\nu)2} = \omega^2 \mu \epsilon_\nu.$$

Normalization function due to the related power

$$T_m^{(\nu)} = \frac{1}{k_{zm}^{(\nu)}\sqrt{\omega\mu k_{zm}^{(\nu)}}} \cdot \begin{cases} \sqrt{\dfrac{2}{ab}}, & \nu = \mathrm{I} \\ \sqrt{\dfrac{2}{bc}}, & \nu = \mathrm{II} \\ \dfrac{1}{\sqrt{b\dfrac{a-d}{2}}}, & \nu = \mathrm{III} \end{cases}$$

Coupling Integrals Due to the Orthogonal Expansion

$$I_{mq}^{\mathrm{II}} = \int_{x=0}^{c} \sin\frac{m\pi}{a}x \cdot \sin\frac{q\pi}{c}x\,dx$$

$$I_{mk}^{\mathrm{III}} = \int_{x=d}^{a} \sin\frac{m\pi}{a}x \cdot \sin\frac{k\pi}{(a-d)}(a-x)\,dx.$$

Scattering Coefficients in (6)

$$(S)_{11} = (S)_{22} = (W)^{-1}(Z)$$

$$(S)_{21} = (S)_{12} = (W)^{-1}(Y)$$

with the abbreviations

$$(W) = \left(U - (\Psi+U)^{-1}\varphi(\Psi+U)^{-1}\varphi\right)$$

$$(Z) = \left((\Psi+U)^{-1}\varphi(\Psi+U)^{-1}\varphi - (\Psi+U)^{-1}(\Psi-U)\right)$$

$$(Y) = \left((\Psi+U)^{-1}\varphi - (\Psi+U)^{-1}\varphi(\Psi+U)^{-1}(\Psi-U)\right)$$

where

U unit matrix

$\Psi = L^{\mathrm{III}} 2\partial^{\mathrm{III}} + L^{\mathrm{II}} 2\partial^{\mathrm{II}} - L^{\mathrm{III}}(N^{\mathrm{III}})^{-1} - L^{\mathrm{II}}(N^{\mathrm{II}})^{-1}$

$\varphi = L^{\mathrm{III}} 2\xi^{\mathrm{III}} + L^{\mathrm{II}} 2\xi^{\mathrm{II}}$

$\partial^\nu = (U - D^\nu D^\nu)^{-1}(N^\nu)^{-1}$

$\xi^\nu = (U - D^\nu D^\nu)^{-1} D^\nu (N^\nu)^{-1}$

$L^\nu = (R_{Emn}^{\mathrm{I}})^{-1} \cdot R_{Emq}^{\nu}$

$N^{\mathrm{II}} = (R_{Hmq}^{\mathrm{II}})^{-1} \cdot R_{Hkk}^{\mathrm{II}}$

$N^{\mathrm{III}} = (R_{Hmk}^{\mathrm{III}})^{-1} \cdot R_{Hkk}^{\mathrm{III}}.$

Diagonal matrix of the eigenmodes (also below their cutoff frequency) along the waveguide sections i between the step discontinuities

$$(D)^\nu = \begin{pmatrix} e^{-jk_{z1}^\nu \cdot l_i} & & 0 \\ & \ddots & \\ 0 & & e^{-jk_{zm}^\nu \cdot l_i} \end{pmatrix}.$$

Abbreviation R

$$R_{Emm}^{\mathrm{I}} = \frac{a}{2} k_{zm}^{(\mathrm{I})} T_m^{(\mathrm{I})}$$

$$R_{Emq}^{\mathrm{II}} = k_{zq}^{(\mathrm{II})} T_q^{(\mathrm{II})} I_{mq}^{\mathrm{II}}$$

$$R_{Emk}^{\mathrm{III}} = k_{zk}^{(\mathrm{III})} T_k^{(\mathrm{II})} I_{mk}^{\mathrm{III}}$$

$$R_{Hmq}^{\mathrm{II}} = k_{zm}^{(\mathrm{I})2} T_m^{(\mathrm{I})} I_{mq}^{\mathrm{II}}$$

$$R_{Hmk}^{\mathrm{III}} = k_{zm}^{(\mathrm{I})2} T_m^{(\mathrm{I})} I_{mk}^{\mathrm{III}}$$

$$R_{Hkk}^{\mathrm{III}} = \frac{a-d}{2} k_{zk}^{(\mathrm{III})2} T_k^{(\mathrm{III})}$$

$$R_{Hkk}^{\mathrm{II}} = \frac{c}{2} k_{zk}^{(\mathrm{II})2} T_k^{(\mathrm{II})}.$$

I coupling integrals (see above).

Acknowledgment

The E-band metal insert filter, as well as the V-band filter of [7], were measured in the microwave laboratory of AEG-Telefunken, Ulm, Germany. The authors take the opportunity to thank Dr. Rembold, the head of the laboratory and the members of his staff, especially Dr. Menzel, for this aid.

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