LOW-INSERTION-LOSS FIN-LINE FILTERS FOR MILLIMETRE-WAVE APPLICATIONS

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ABSTRACT

A design theory is described for low-insertion-loss fin-line filters with fin-line gap widths over the total height of the waveguide housing. The theory includes both higher order mode propagation and the finite thicknesses of the dielectric and the metallic fins. An evolution strategy synthesis method yields optimum design data for three- and five resonator-type fin-line filters with several substrate thicknesses. The mid-band frequencies chosen are about 13, 34, 66, and 75 GHz. Measured minimum insertion-losses in the passband are about 0.2 dB (13 GHz), 0.5 dB (34 GHz), and 1.9 dB (65 GHz) for three-resonator-type filters as an example.

INTRODUCTION

Fin-line filters, [1] to [4], are very attractive circuit elements for integrated mm-wave circuits. Up to now, with exception of [1], mainly fin-line filters with narrow gap widths have been developed. Lower insertion-loss in the passband, however, can be obtained with larger gaps. The reason is that the measured quality factor of bilateral fin-lines increases with respect to gap width between the fins. The optimum value is reached if the gap width is equal to the total height of the waveguide housing [1]. Low-insertion-loss fin-line filters are therefore calculated, designed and described using this fin-line type (Fig. 1).

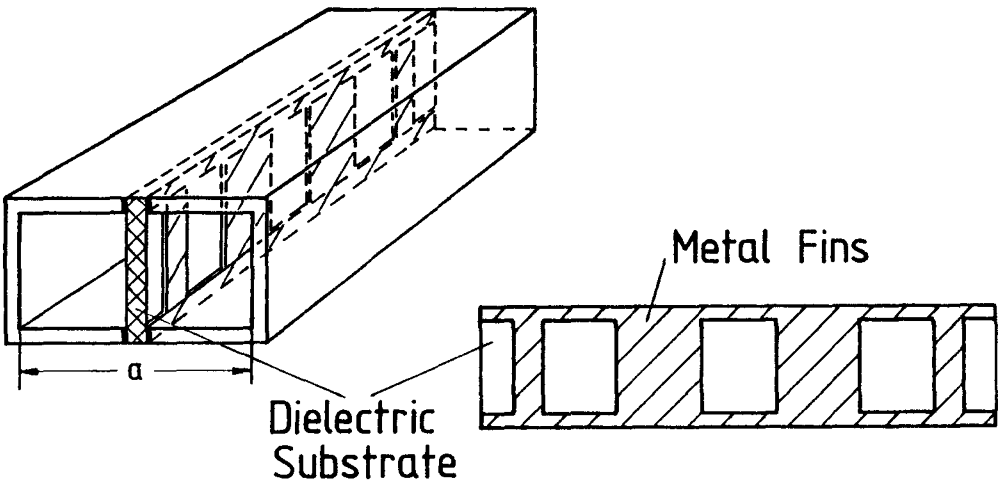

Fig. 1: Low-insertion-loss fin-line filter

A survey of available theories, [1] to [5], for three-dimensional fin-line structures shows that they contain restrictions which may limit the low-insertion-loss design. Therefore a design theory is introduced, based on the orthogonal expansion into eigenmodes, that includes the effects of both higher order mode propagation and the finite thickness of the dielectric and the metallic fins on mid-band insertion-loss, stop-band attenuation and mid-band frequency.

In order to be able to take advantage of the low-insertion-loss potential of this fin-line-filter type, a simple synthesis procedure is described which is based on the principle of evolution optimization [6], [7]. The filter dimensions are optimized for a given thickness of the dielectric and copper cladding, as well as for the roughly fixed mid-band frequency and 3db-bandwidth. Calculations and design are exemplarily verified by measurements.

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THEORY

The basic idea of the theory is to regard the fin-line-filter as consisting of alternating waveguide structures: waveguide with perpendicular a dielectric slab in the middle of the waveguide, and three parallel waveguides, the middle of which is filled with the same dielectric (cf. Fig. 2a). Therefore two scattering matrix types have to be calculated: the scattering matrix of the three-waveguides-structure of finite length within the waveguide (Fig. 2b), and the scattering matrix of the dielectric-slab-structure of finite length also within the same waveguide (Fig. 2c). The scattering matrix of the total fin-line structure is obtained by suitably combining the transitions, the length of waveguide I (cf. Figs. 2b and 2c) being reduced to zero if structures b or c (or inverse) are joined together directly.

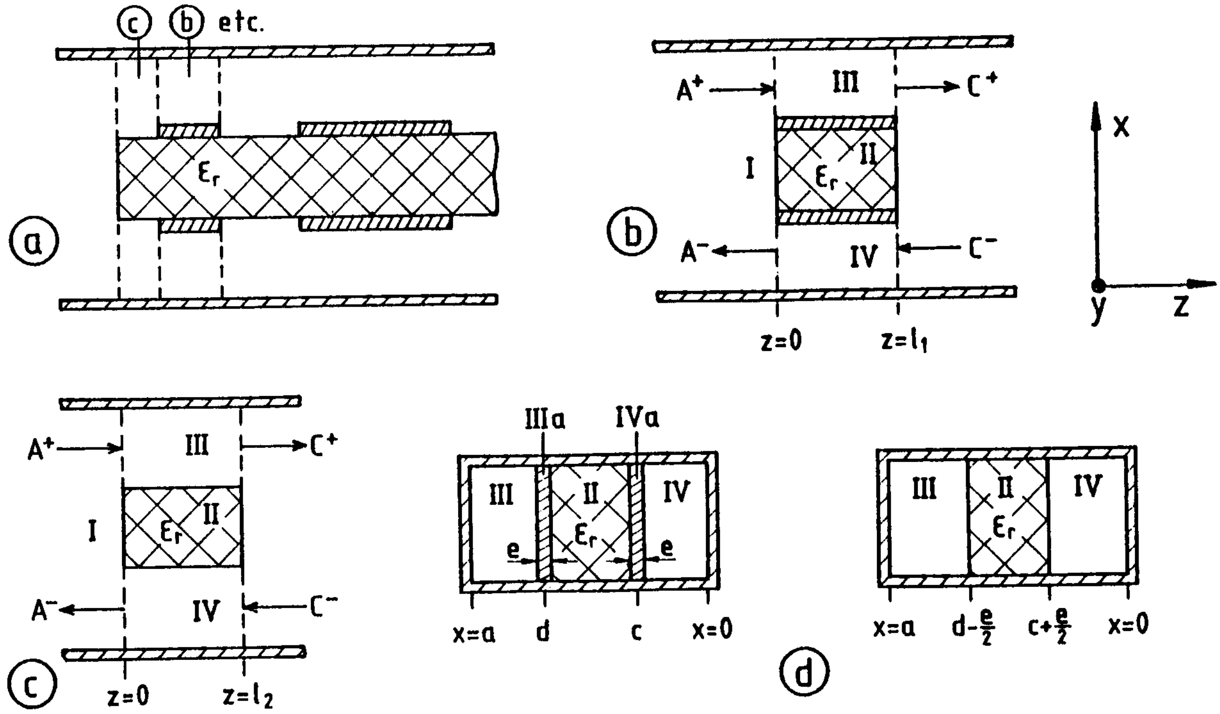


Fig. 2: Filter configuration for the field theory treatment
a alternating waveguide structures: waveguide with a dielectric slab (c) and three parallel waveguides (b)
b three-waveguide-structure of finite length within the waveguide
c dielectric-slab-structure of finite length within the same waveguide
d corresponding cross-sections

For each subregion ν = I,II,III,IV of the three-waveguide-structure (Fig. 2b) the fields are derived from the x-component of the Hertzian vector potential $\vec{Q}$ which is assumed to be a sum of suitable eigenmodes

$$Q_x^\nu = \sum_{m=1}^{\infty} A_m^\nu \sin\left(\frac{m\pi}{p^\nu} f^\nu\right) \cdot e^{-jk_{zm}^\nu \cdot z} \quad , \tag{1}$$

with (cf. Fig. 2d)

$$\begin{bmatrix} f^I \\ f^{II} \\ f^{III} \\ f^{IV} \end{bmatrix} = \begin{bmatrix} x \\ d-\frac{e}{2}-x \\ a-x \\ c-\frac{e}{2}-x \end{bmatrix}, \quad \begin{bmatrix} p^I \\ p^{II} \\ p^{III} \\ p^{IV} \end{bmatrix} = \begin{bmatrix} a \\ d-c-e \\ a-(d+\frac{e}{2}) \\ c-e/2 \end{bmatrix}, \quad \begin{bmatrix} k_{zm}^{I2} \\ k_{zm}^{II2} \\ k_{zm}^{III2} \\ k_{zm}^{IV2} \end{bmatrix} = k^{\nu 2} - \begin{bmatrix} (\frac{m\pi}{a})^2 \\ (\frac{m\pi}{p^{II}})^2 \\ (\frac{m\pi}{p^{III}})^2 \\ (\frac{m\pi}{p^{IV}})^2 \end{bmatrix}, \tag{2}$$

$$k^{\nu 2} = \omega^2 \mu \epsilon, \quad \epsilon_\nu = \epsilon_o \cdot \epsilon_r^{(\nu)} .$$

By matching the field components at the corresponding interfaces F^ν of the adjacent subregions (Fig. 2d) at z=0 and $z=l_1$ the coefficients A_m^ν in equation (1) can be determined after multiplication with the appropriate orthogonal function related to each subregion.

If the forward and backward waves at the two steps (Fig. 2b) of the structure of finite length l_1 are suitably normalized and related together, the coefficients can be written in the form of the desired scattering matrix

$$\begin{bmatrix} (A)^- \\ (C)^+ \end{bmatrix} = \begin{bmatrix} (S_{11}) & (S_{12}) \\ (S_{21}) & (S_{22}) \end{bmatrix} \begin{bmatrix} (A)^+ \\ (C)^- \end{bmatrix} \quad . \tag{3}$$

A similar procedure is chosen for calculating the dielectric-slab-structure (Fig. 2c). The common propagation factor k_{zm} in the region II, III, IV for this structure is determined numerically by the boundary conditions along the dielectric slab, which lead to the transcendent equation

$$\frac{1}{k_{zm}^{II}} \cdot \tan(k_{xm}^{II}(d-c)) + \frac{1}{k_{xm}^{III}} \cdot \tan(k_{xm}^{III}(a-d)) + \frac{1}{k_{xm}^{IV}} \cdot \tan(k_{xm}^{IV} \cdot c) - \frac{k_{xm}^{II}}{k_{xm}^{III} \cdot k_{xm}^{IV}} \tan(k_{xm}^{II}(d-c)) \tan(k_{xm}^{III}(a-d)) \tan(k_{xm}^{IV} \cdot c) = 0, \tag{4}$$

where the relation holds

$$\begin{bmatrix} k_{xm}^{II2} \\ k_{xm}^{III^2} \\ k_{xm}^{IV2} \end{bmatrix} = \begin{bmatrix} k_o^2 \cdot \epsilon_r \\ k_o^2 \\ k_o^2 \end{bmatrix} - k_{zm}^2 , \quad k_o^2 = \omega^2 \mu_o \epsilon_o \; . \tag{5}$$

DESIGN

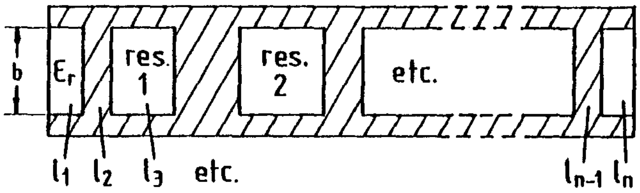
An error function $F(\overline{x})$ is defined

$$F(\overline{x}) = \sum_{NSB} (GSMIN/A_{21}(SB))^2 + \sum_{NPP} (A_{21}(PP)/GPMAX)^2 \tag{6}$$

where NSB, NPP = given number of frequency sample points in the stopband, and passband, respectively; GSMIN, GPMAX = given minimum stopband, and maximum passband attenuation, respectively; A_{21} (SB), A_{21}(PP) = calculated amount of the fin-line filter attenuation in stopband, and passband, respectively. For a given thickness of the dielectric, and of the copper cladding, the parameters $\overline{x} = l_1, l_2, \ldots, l_n$ (cf. table) are suitably optimized, using the evolution strategy method[6],[7] . This method compared e.g. with the Fletcher-Powell method has the advantage, that no differentiation step in the optimizing process is necessary, which reduces the computing time involved.

For the optimization procedure five eigenmodes in each subregion are considered; the final design values (cf. table) are calculated with twenty eigenmodes. The total computing time for e.g. an optimized five resonator fin-line filter was about 20-30min (SIEMENS-7880 computer). The design table shows optimized filter data for some examples.

Design-table



$n = 2n_R + 3$

Copper cladding: 17.5µ [••]

Fin-line Filter	n_R	S	Substrate thickness t	$l_1=l_n$ 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	l_7 mm	f_0 GHz	Δf* MHz
Ku-band a=15.80mm b= 7.90mm	3	R	0.01"	24.26	1.900	11.30	6.89	11.30			13.55	390
			1/16"	15.660	1.500	8.000	5.840	8.000			15.19	524
Ka-band a=7.11mm b=3.56mm	3	R	0.01"	19.81	0.650	3.815	3.800	3.850			33.585	630
			1/32"	12.260	0.480	3.525	2.970	3.530			33.99	635
	5		0.02"	12.887	0.624	3.834	3.144	3.890	3.676	3.891	32.95	695
V-band a=3.76mm b=1.88mm	3	R	0.01"	14.712	0.350	1.91	2.038	1.92			64.995	920
	5		0.005"	7.2	0.685	1.771	1.64	1.31	1.516	1.756	67 ·	300
E-band a=3.10mm b=1.55mm	1	F	0.558mm	0	0.582	1.101					75	500
	2		0.558mm	0	0.374	1.096	1.123				75.1	1000
	3		0.558mm	10.939	0.365	1.463	1.503	1.461	1.503		75	1500
	5		0.220mm	8.073	0.334	1.452	1.359	1.467	1.583	1.464	74.9	1800

n_R = number of resonators
S = substrate
R = RT/duroid 5880, ϵ_r = 2.22
F = fused silica (quartz), ϵ_r = 3.8

*Δf = 3dB-bandwidth
** for the E-band filters: 5µ

RESULTS

Figs.3 to 5 show the calculated and measured insertion-loss (scattering coefficient $1/S_{21}$) in dB as a function of frequency for designed Ku-, Ka-, and V-band fin-line filters as examples. The frequency displacement between calculated an measured curves is caused by etching errors, which alter the effective resonator lengths. Fig. 6 shows the calculated insertion-loss of an E-band fin-line filter, where fused silica (quartz) is chosen for substrate material because of its lower loss and surface roughness. A photoetched fin-line structure (V-band) is shown in Fig.7.

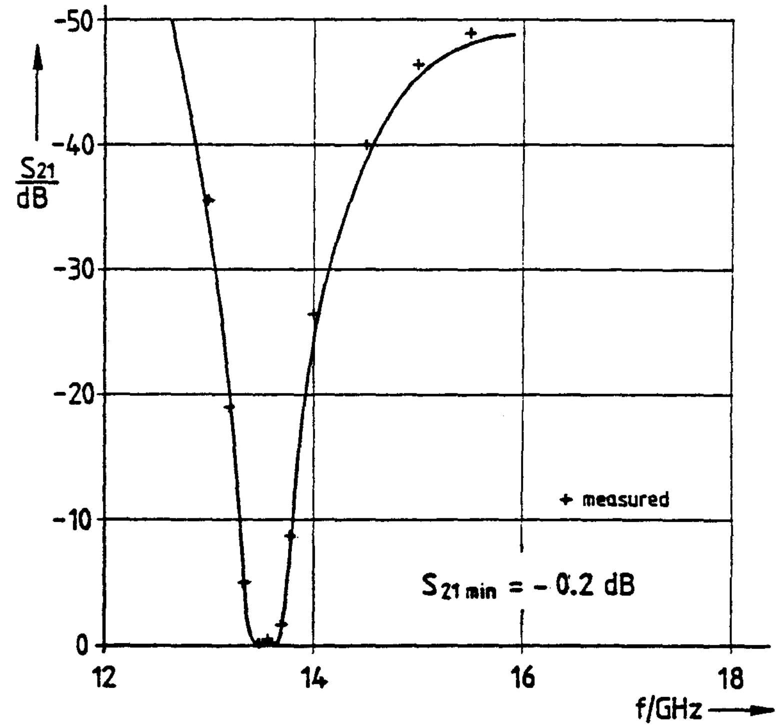


Fig. 3: Calculated and measured insertion-loss of a Ku-band fin-line filter
(3 resonators, RT/duroid 0.01", cf.table)

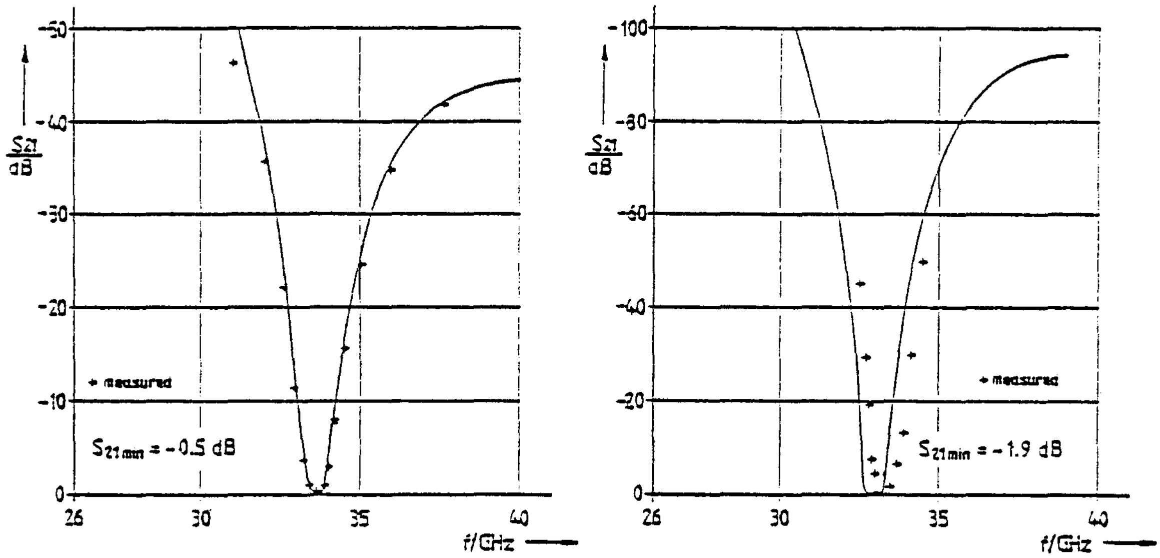


Fig. 4: Calculated and measured insertion-loss of Ka-band fin-line filters
a 3 resonators, RT/duroid 0.01", cf. table
b 5 resonators, RT/duroid 0.02", cf. table

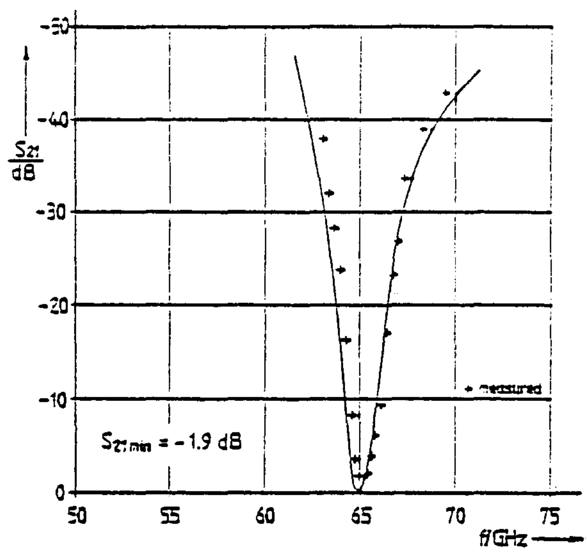


Fig. 5: Calculated and measured insertion-loss of a V-band fin-line filter (3 resonators, RT/duroid 0.01", cf. table)

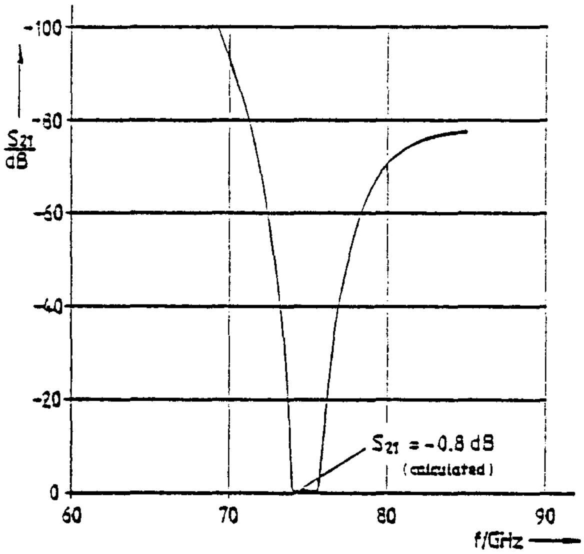


Fig. 6: Calculated insertion-loss of an E-band fin-line filter (5 resonators, quartz 220µm, cf. table)

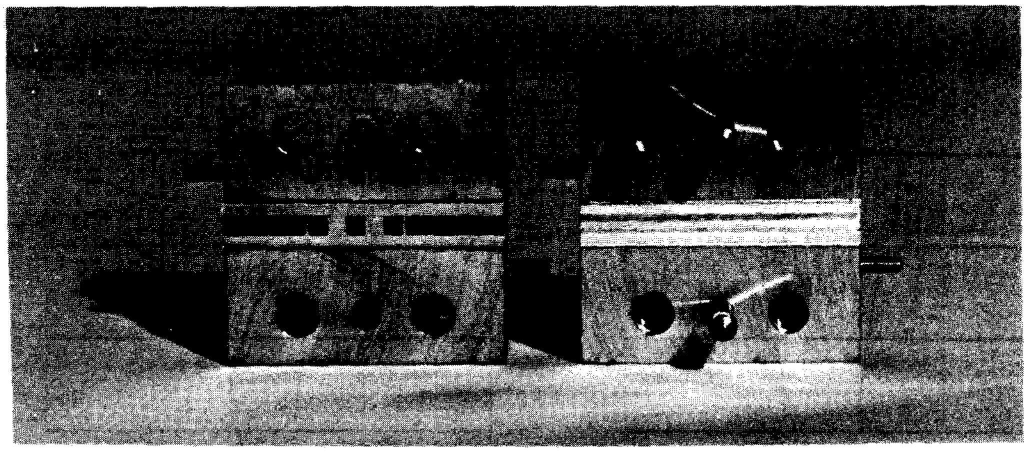


Fig. 7: Photograph of the V-band fin-line filter (cf. Fig. 5)

CONCLUSION

A design theory has been described for low-insertion-loss fin-line filters. The theory is based on the field expansion into suitable eigenmodes, which allows both higher order mode propagation and the finite thicknesses of the dielectric slab and the metallic fins to be included. A synthesis procedure provides design data for given input parameters. The measured insertion-loss of designed filters verifies the described theory. RT/duroid is used for substrate material providing measured minimum insertion-losses in the passband of 0.2 dB (13 GHz), 0.5 dB (34 GHz), and 1.9 dB (65 GHz). For further practical investigations fused silica (quartz) is foreseen because of its lower loss and surface roughness.

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