waveguides; Variational method and simpler results [Russian]," *Radiotekh. Elektron.*, vol. 21, pp. 1137–1148, 1976. (Engl. trans. in *Radio Eng. Electron. Phys.*, vol. 21, no. 6, pp. 1–11, 1976.)

[4] E. I. Nefedov, N. P. Santalov, and A. T. Fialkovskii, "Dispersion characteristics of nonradiating waves in a microstrip waveguide [Russian]," *Radiotekh. Elektron.*, vol. 24, pp. 17–23, 1979. (Engl. transl. in *Radio Eng. Electron Phys.*, vol. 24, no. 1, pp. 15–20, 1979.)

[5] D. C. Chang and E. F. Kuester, "An analytic theory for narrow open microstrip," *Arch. Elek. Übertragung.*, vol. 33, pp. 199–206, 1979.

[6] E. F. Kuester and D. C. Chang, "Closed-form expressions for the current or charge distribution on parallel strips or microstrip," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 254–259, 1980; addendum, *ibid.*, p. 1143.

[7] S. Y. Poh, W. C. Chew, and J. A. Kong, "Approximate formulas for the line capacitance and characteristic impedance of microstrip line," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 135–142, 1981.

[8] W. C. Chew and J. A. Kong, "Asymptotic formula for the capacitance of two oppositely charged discs," *Math. Proc. Camb. Phil. Soc.*, vol. 89, pp. 373–384, 1981.

[9] E. F. Kuester, R. T. Johnk, and D. C. Chang, "The thin-substrate approximation for reflection from the end of a slab-loaded parallel-plate waveguide with application to microstrip patch antennas," *IEEE Trans. Antenna Propagat.*, vol. AP-30, pp. 910–917, 1982.

[10] H. A. Wheeler, "Transmission-line properties of a strip on a dielectric sheet on a plane," *IEEE Trans. Microwave Theory Tech.* vol. MTT-25, pp. 631–647, 1977.

[11] D. C. Chang and E. F. Kuester, "An analytic theory for narrow open microstrip," Electromagnetics Laboratory, Dept. of Elec Eng., Univ. of Colorado, Boulder, Sci. Rep. 28, May 1978.

W-Band Low-Insertion-Loss *E*-Plane Filter

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Abstract **— Computer-optimized design data for a four-resonator metallic *E*-plane filter are given with a midband frequency of about 94 GHz. The method of field expansion into suitable eigenmodes used considers the effects of finite insert thickness and higher order mode interaction. The measured minimum passband insertion loss of a metal filter prototype is 1.5 dB.**

I. Introduction

Increasing interest in the 3-mm region of millimeter waves has stirred the need for low-cost low-loss *W*-band filters. Pure metal inserts [1]–[4] placed in the *E*-plane of rectangular waveguides without any substrates (Fig. 1) achieve low-loss designs. Moreover, metal etching techniques [3], [4] can be applied for mass production. This paper yields computer-optimized design data for such a *W*-band filter. The computer-aided design algorithm is based on field expansion into eigenmodes [3]–[4] which, in contrast to the residue-calculus technique [5], includes the finite strip thickness considerably influencing passband ripple behavior and midband frequency, as has already been demonstrated in [3]. The measured frequency response of the metal insert filter that has been designed and operated at 94 GHz shows good agreement with theory.

II. Theory and Design

As in [3], [4], the fields for each subregion (ν) at the corresponding step discontinuities of the filter structure (Fig. 1) are derived from the x-component of the magnetic Hertzian vector

Manuscript received May 6, 1983; revised August 23, 1983.

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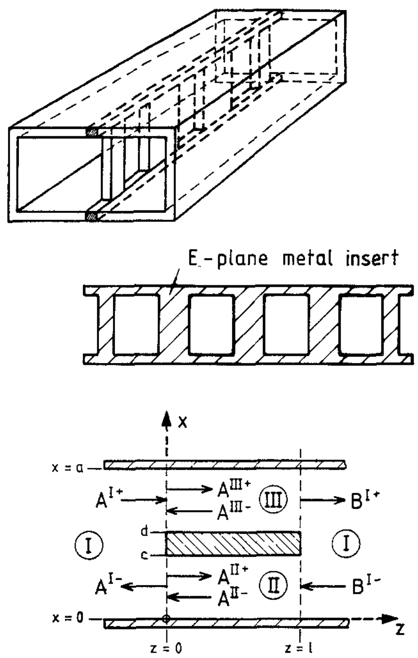


Fig. 1. Low-insertion-loss *E*-plane metal insert filter.

$\vec{\Pi}_h$, which is assumed to be a sum of suitable eigenmodes

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

with

$$(f^{(\nu)})' = (x, x, a-x), \quad (p^{(\nu)})' = (a, c, a-d)$$

$$k_{zm}^{(\nu)} = \sqrt{k^2 - \left(\frac{m\pi}{p^{(\nu)}}\right)^2}, \qquad k^2 = \omega^2 \mu \epsilon.$$

$A_m^{(\nu)\pm}$ are the still unknown eigenmode coefficients which are suitably normalized with the factor $T_m^{(\nu)}$ [3], [4], so that the power carried by a given wave is 1 W for a wave-amplitude coefficient of $\sqrt{1W}$.

By matching the tangential field components at the step discontinuity interfaces, the coefficients $A_m^{(\nu)\pm}$ are determined after multiplication with the appropriate orthogonal function. This leads directly to the scattering matrix of each discontinuity. The overall scattering matrix of the total filter section is calculated without introducing transmission matrices by directly combining the single scattering matrices [3], [4], which preserves numerical accuracy.

As has already been introduced for low-loss fin-line filters [6], the computer-aided design is carried out by an optimizing program applying the evolution strategy method [7], where no differentiation step is necessary, which varies the input parameters until a desired value of the insertion loss for a given bandwidth is obtained. An error function $F(\bar{x})$ [6] to be minimized is defined (Fig. 2)

$$F(\bar{x}) = \sum_{i=1}^{N_s} \left(a_{s\min}/a_{21}(f_i)\right)^2 + \sum_{i=1}^{N_p} \left(a_{21}(f_i)/a_{p\max}\right)^2 \overset{!}{=} \mathrm{Min} \tag{2}$$

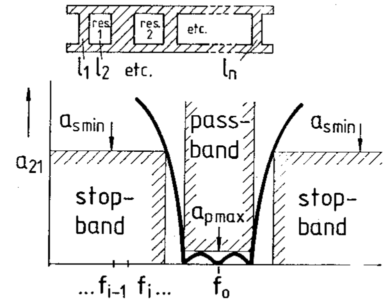
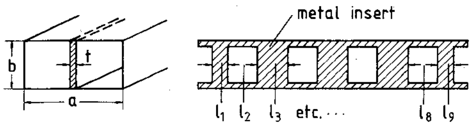


Fig. 2. Scheme for the computer optimization.

TABLE I
DESIGN DATA FOR THE W-BAND FOUR-RESONATOR METAL INSERT FILTER



Frequency-band waveguide housing	Number of resona-tors	Insert thick-ness t	$l_1 = l_9$ (mm)	$l_2 = l_8$ (mm)	$l_3 = l_7$ (mm)	$l_4 = l_6$ (mm)	l_5 (mm)	Midband-frequency (GHz)	3 dB-band-width (GHz)	min. passband insertion-loss (dB)	Remarks
W - band	4	50 µm	0.599	1.439	1.811	1.440	2.008	94	1.2	0.001	computer optimized
a = 2.54 mm b = 1.27 mm WR 10			l_1 0.602	l_2 1.443	l_3 1.818	l_4 1.439	l_5 2.020	93.95	0.9	1.5	measured
			l_9 0.604	l_8 1.444	l_7 1.820	l_6 1.443					

where N_s and N_p are the number of frequency sample points f_i in the stopband and passband, respectively, $a_{s\min}$ and $a_{p\max}$ are the given minimum stopband and maximum passband attenuation levels, respectively, and $a_{21} = 20\log(1/|S_{21}|)$ is the calculated insertion-loss. For given waveguide housing dimensions and thickness of the metallic E-plane inserts, the parameters $\bar{x}$ to be optimized are the insert and the resonator lengths. For computer optimization, the expansion into 15 eigenmodes at each discontinuity has turned out to be sufficient. The final design data are checked up by 45 eigenmodes.

III. RESULTS

A four-resonator W-band E-plane filter is chosen for design example (cf. Table I), where the computer-optimized design data are given. The tolerances of the photoetched metal insert dimensions in relation to the optimized data lie between -1 and $+12$ μm, as has been checked up by a measuring microscope. The material of the 50-μm-thick inserts is 99.9-percent pure copper. Fig. 3 shows the photograph of the filter structure, together with the opened waveguide housing. The insert is held in the split block by alignment pins.

The calculated and measured insertion loss $(1/|S_{21}|)$ in decibels is plotted in Fig. 4 as a function of frequency for the four-resonator computer optimized W-band E-plane filter. The calculated minimum insertion loss is 0.001 dB, the measured minimum value (see detail Fig. 1(b)) is about 1.5 dB at 93.95 GHz, and the measured typical mean passband insertion loss is about 1.8 dB.

IV. CONCLUSION

The efficient computer-aided higher order mode design of E-plane integrated filters, introduced by the authors in 1981 [6], leads to optimum low-insertion-loss pure metal insertion W-band filters. A four-resonator optimized prototype fabricated by metal etching techniques exhibits a measured minimum passband insertion loss of about 1.5 dB at 93.95 GHz. Since the finite metal thickness is included in the computations, measured characteristics agree excellently with theory.

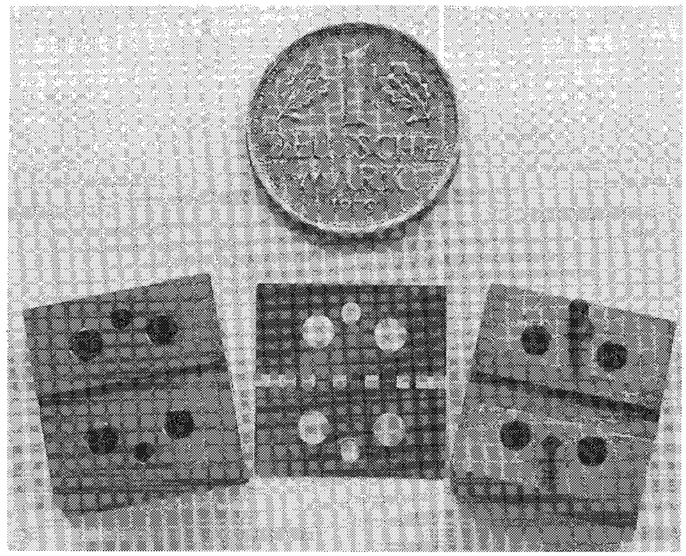


Fig. 3. Photograph of the W-band metal insert filter (photoetched metal insert together with the opened waveguide housing).

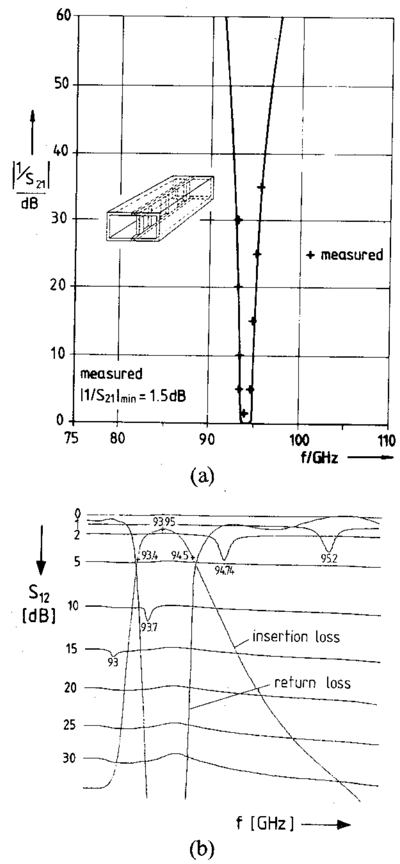


Fig. 4. Computer optimized four-resonator E-plane W-band metal insert filter. (a) Calculated and measured insertion-loss as a function of frequency. (b) Expanded view of measured insertion and return loss.

Acknowledgment

The *W*-band waveguide block has been fabricated and the filter has been measured in the microwave laboratory of AEG-Telefunken, Ulm, West Germany. The authors are greatly indebted to Dr. B. Rembold, the head of the laboratory, and the members of his staff, especially Dr. W. Menzel, for this aid.

References

[1] Y. Konishi and K. Uenakada, "The design of a bandpass filter with inductive strip—Planar circuit mounted in waveguide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 869–873, Oct. 1974.

[2] P. J. Meier, "Planar multiport millimeter integrated circuits," in *Dig. IEEE 1977 G-MTT Symp.*, June 1977, pp. 385–388.

[3] R. Vahldieck, J. Bornemann, F. Arndt, and D. Grauerholz, "Optimized waveguide *E*-plane metal insert filters for millimeter-wave applications," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 65–69, Jan. 1983.

[4] J. Bornemann, R. Vahldieck, F Arndt, and D. Grauerholz, "Optimized low-insertion-loss millimetre-wave fin-line and metal insert filters," *Radio Electron. Eng.*, vol. 52, no. 11/12, pp. 513–521, Nov./Dec. 1982.

[5] Yi-Chi Shi, T. Itoh, and L. Q. Bui, "Computer-aided design of millimeter-wave *E*-plane filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 135–142, Feb. 1983.

[6] F. Arndt, J. Bornemann, D. Grauerholz, and R. Vahldieck, "Low-insertion-loss fin-line filters for millimeter-wave applications," in *11th Eur. Microwave Conf.* (Amsterdam, Netherlands), Sept. 7–10, 1981, pp. 309–314.

[7] H. Schmiedel, "Anwendung der Evolutionsoptimierung bei Mikrowellenschaltungen," *Frequenz*, vol. 35, pp. 306–310, Nov. 1981.

S- and *X*-Band GaAs FET Mixers with Thin-Film Lumped Elements

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***Abstract*—The design and performance of 2- and 11-GHz band mixers with a single-gate GaAs FET are presented in this paper.**

A mixer configuration in which the local oscillator (LO) signal is applied to the source is used. Matching networks are constructed with thin-film lumped elements fabricated on alumina.

An SSB noise figure of 6.2 dB, with an associated conversion gain of 10 dB, has been achieved at the 11-GHz band, and SSB noise figures of less than 6 dB and a conversion gain of more than 8 dB over a 40-percent bandwidth are obtained at the 2-GHz band.

I. Introduction

The operation and experimental results of GaAs MESFET's as microwave mixers have already been reported [1]–[3].

The performance of FET mixers has advantages in terms of conversion gain over Schottky-diode mixers, which makes the mixer nearly insensitive to the IF amplifier noise figure.

FET mixers have already been realized with single-gate or dual-gate FET's. In the dual-gate mixer configuration, the RF and local oscillator (LO) signals are applied to the first and second gate, respectively, offering a high degree of isolation between RF and LO ports.

Two single-gate FET mixer types have been reported. One type is a gate mixer in which the local signal is fed to the gate, and another type is a drain mixer in which the local signal is fed to the drain [3], [5].

Manuscript received May 24, 1983; revised August 3, 1983.
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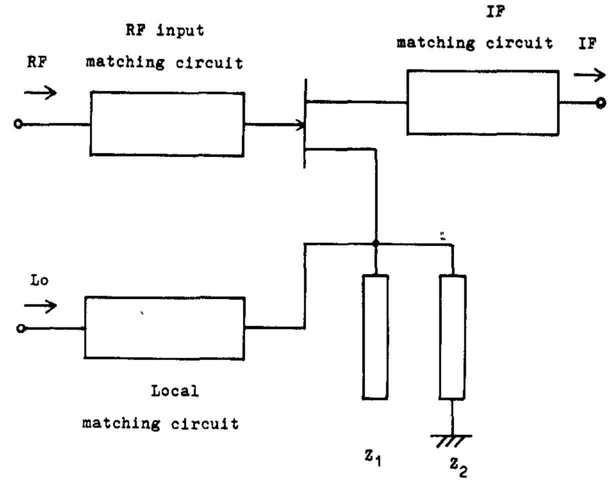

Fig. 1. FET mixer at the *X*-band.

In the case of single-gate FET mixers, one problem is the necessity of introducing both the local and RF input signals into the gate of the FET [2]. Since a coupler with a high coupling ratio is used to provide RF and LO signal isolation, the use of a high LO power level, therefore, is required.

This paper describes and provides performance data on a mixer circuit in which the RF signal is fed to the gate and the LO signal to the source.

The means of construction using a thin-film inductor and capacitor is discussed.

II. Design Principle

In order that a simple circuit can be realized, a single-gate FET mixer configuration has been adopted. The circuit diagram for an *X*-band circuit is shown in Fig. 1.

In the circuit, the LO signal of frequency F_L and the RF input signal of frequency F_{RF} are applied to the source and the gate of the FET, respectively, where mixing takes place in the FET by means of the nonlinearity of the transconductance and drain resistance of the FET.

The mixing product is the IF signal of frequency $|F_L - F_{RF}|$ which is amplified in the FET. The single-gate FET mixer has a simple structure, and does not require couplers or combiners.

To minimize the noise figure in single-gate *X*-band mixers, the RF input frequencies are terminated in a short circuit with a quarter-wavelength open stub (Z_1) at the source of the FET, and the drain is terminated in a short circuit at the RF input frequencies and local signal frequencies. To obtain low-noise performance of the IF amplifier and efficient injection of LO signal into the FET, the source is terminated in a quarter-wavelength short stub (Z_2) for local frequencies. Consequently, the impedance between the source and the common terminal is approximately shorted at the IF frequencies and open at the local frequencies.

Fig. 2 shows a circuit diagram designed for an *S*-band mixer. To obtain wide-band low-noise mixer performance, the source of the FET is terminated in a parallel circuit consisting of a resistor and a quarter-wavelength short stub in the highest local frequencies. Therefore, the impedance of the source at wide-band local