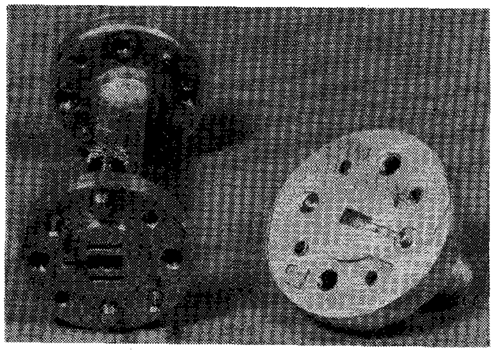

Fig. 3. Advanced mixer mount.

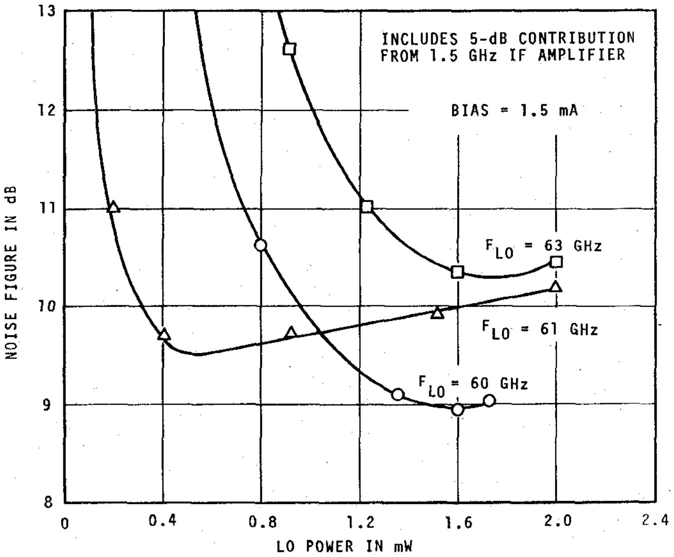

Fig. 4. Noise figure of mixer with GaAs diode.

the noise contributed by the IF amplifier. Fig. 4 shows how the measured noise figure varied with respect to LO power, with the LO frequency as a parameter. The 1.5-GHz IF contributed 5 dB to the measurement, and the dc bias voltage was varied to maintain a constant current of 1.5 mA through the diode. The measured noise figure was 9–10 dB for LO frequencies in the 60–61-GHz range for drive levels of 0.9–1.8 mW.

It is believed that an overall noise figure of 6 dB is feasible in future models with superior FET IF amplifiers. Owing to the low-loss characteristics of oversized microstrip, its freedom from tight tolerances and compatibility with hybrid devices, this new transmission line is well suited to a wide variety of integrated circuits throughout the millimeter spectrum.

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Simplified 12-GHz Low-Noise Converter with Mounted Planar Circuit in Waveguide

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***Abstract*—A 12-GHz low-noise converter consisting of a planar circuit mounted in waveguide is described. This circuit consists of a metal sheet with proper patterns that is inserted in the middle of a waveguide parallel to the *E* plane. All circuit elements required for the converter are pressed or etched. This circuit is very useful for low-cost mass production and good performance. A measured noise figure of 4.5 dB was obtained with a 12-GHz signal frequency and a 420-MHz intermediate frequency.**

I. INTRODUCTION

A low-noise 100-GHz converter comprising a microwave integrated circuit mounted in a waveguide was reported by Konishi and Hoshino in 1971 [1]. A low-noise 12-GHz converter based on a similar principle is described in this short paper. It consists of a planar circuit mounted in a waveguide and is suitable for low-cost mass production.

Millimeter components using integrated circuits mounted in a waveguide were also developed by Meier in 1972 [2], where unloaded Q of this transmission line takes a value less than 900 at X band [3].

A new type of filter, that is, a planar circuit mounted in waveguide, that we have developed has an unloaded Q factor of 2000–2500 at X band. This new type of filter is used in our 12-GHz converter to achieve low-noise performance.

The 12-GHz converter described in this short paper has a mixer conversion loss of 3.5 dB and a total noise figure of 4.5 dB, including the contribution of an intermediate frequency amplifier with a noise figure of 2.0 dB at 420 MHz.

II. 12-GHz LOW-NOISE CONVERTER WITH MOUNTED PLANAR CIRCUIT IN WAVEGUIDE

A high-sensitivity and low-cost converter was required that could be constructed simply and be mass produced. The construction of the circuit we developed is such that every necessary circuit element is arranged on a metal sheet merely by pressing or etching, and the metal sheet is inserted into a waveguide.

We will describe the results of the experiment that was carried out on our converter with a mounted planar circuit.

Fig. 1 shows the planar circuit pattern from left to right, a signal frequency bandpass filter, a Schottky barrier diode mount, a local oscillator frequency bandpass filter, and a Gunn diode mount for the local oscillator. A 0.3–0.5-mm-thick copper sheet is favorable for this pattern. When etching is used in forming the pattern, a 0.3-mm copper sheet is utilized for dimensional precision. When pressing is used, with a 0.5-mm-thick metal sheet, dimensions are maintained within 20 μm.

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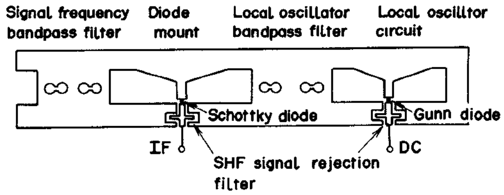

Fig. 1. Construction of 12-GHz converter with planar circuit in waveguide.

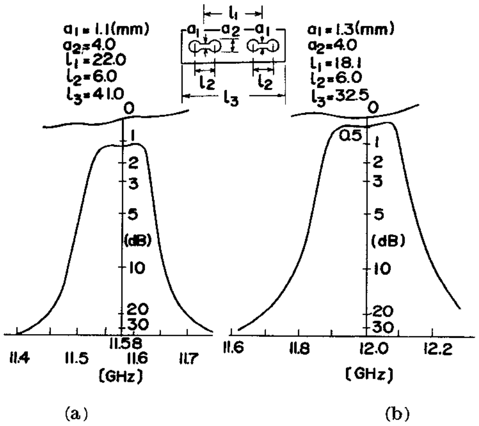

Fig. 2. Bandpass filter (BPF). (a) Characteristic of oscillator BPF. (b) Characteristic of signal BPF.

A. Bandpass Filter

The construction of the bandpass filters for both the signal and the local oscillator frequencies is shown in Fig. 1. The diameter of the circle or the gap is adjusted to set the center frequency.

Fig. 2(a) and (b) shows the characteristics of the local oscillator frequency and the signal frequency bandpass filters. The signal frequency bandpass filter has a 12-GHz center frequency and a 3-dB bandwidth of 230 MHz. Its insertion loss is 0.3 dB. The attenuation is more than 30 dB for the local oscillator frequency and the image frequency. The local oscillator frequency bandpass filter has an 11.58-GHz center frequency and a 3-dB bandwidth of 140 MHz. Its insertion loss is 1.2 dB. Attenuation is more than 30 dB at the signal and image frequencies. It has a narrow band in comparison with the signal frequency bandpass filter. As a result, its insertion loss is higher. An unloaded Q of approximately 2000–2500 is provided for both the bandpass filters.

B. Schottky Diode Mount

Fig. 3(a) shows the Schottky diode mount circuit. A beam lead diode or a Schottky diode packaged for microwave integrated circuits is employed, allowing the diode to be directly mounted on the planar circuit.

Round and flat posts were considered as diode mounts. When round or flat posts are used alone, the height of the waveguide determines the length of the post, and unwanted reactance is included. Matching is more difficult, and the bandwidth will inevitably be narrower. When the diode is mounted at the foot of an open-tip probe that couples to the waveguide mode, the length of the antenna can be varied to match to impedances between 50 and 200 Ω. The width of the antenna determines the frequency bandwidth.

Fig. 3(a) shows the construction that allows the diode to be mounted directly with the metal sheet alone. As shown in Fig. 3(a), a distributed line transformer and a ridge taper are provided. The flat post that is located close to the $\lambda/4$ line is connected to the ridge taper, which has been set vertically in the middle of the H plane of the waveguide, and the diode is connected between the waveguide bottom wall and the flat post, as shown in Fig. 1. In this case, the flat post inserted along the E plane functions approximately as a distributed line transformer of $\lambda/4$. That is, where the diode impedance is Z_D and the $\lambda/4$ line impedance is Z, the impedance viewed toward the diode from the point A in Fig. 3(a) is Z^2/Z_D. As Z is varied by changing the post width, impedance matching becomes possible for a low-impedance diode, too.

The curves in Fig. 3(b)–(d) show the impedance viewed from the diode terminal when the post width W and the height of the ridge taper h are changed. Fig. 3(b) shows the impedance characteristics for various post widths W in the case of $\lambda/4$ post length. An approximate change is made as was described previously. Fig. 3(c) shows the impedance when W of the flat post remains constant at 4 mm and a change is made in the height of the taper. Fig. 3(d) shows the impedance when the width W is changed only by the flat post, the tip of which is shortened. In Fig. 3(d), in comparison with Fig. 3(b), there is an increase in reactance and the bandwidth proves to be narrower. According to those charts, we can obtain arbitrary impedance in wide bands by selecting appropriate W and h.

C. Consideration of Image Impedance

It is necessary to consider the image impedance in reducing the conversion loss [4]. At the image frequency, the signal frequency and the local oscillator frequency bandpass filters, which are arranged at both sides of the diode mount, produce a short circuit at an inside point a little off the end surface of the filter. Accordingly, the distance between the two filters varies the image frequency impedance that is presented to the diode. In general, where the distance between the filters is selected equal to $n\lambda/2$ at the image frequency, the image impedance is an open circuit. As the position of the signal frequency and the local oscillator frequency bandpass filters are adjusted to optimize the signal frequency and local oscillator frequency impedances, it is not possible to provide the exact image frequency impedance for minimum conversion loss.

For the converter described in this short paper, the image impedance remains a little off the open condition and capacitive.

D. Local Oscillating Circuits

The local oscillator source used a Gunn diode, and the construction described in Section II-B is used as a mount.

III. RESULTS OF OUR EXPERIMENT ON PROPOSED CONVERTER

We will describe the results of our measurements on the experimental converter with the metal sheet on which every foregoing pattern is formed, as shown in Fig. 1.

Fig. 4 shows the characteristics of the converter we have developed. The local oscillator drive is 8 mW. The signal frequency is 12 GHz and the intermediate frequency is 420 MHz. We were successful in obtaining a minimum real conversion loss of 3.2 dB. An intermediate amplifier with approximately a 2.0-dB noise figure was connected and a total noise figure 4.5 dB was obtained. This value coincides with the result of the noise figure equation according to the Appendix.

APPENDIX

Overall receiver noise figure F_t can be expressed as

$$F_t = L_r L[t + F_{if} - 1] \tag{1}$$

where

- L_r RF circuitry loss;
- L mixer conversion loss;
- F_{if} noise figure of IF amplifier;
- t mixer noise temperature.

In the case of the image reactive load, the noise figure F of the mixer diode is given by [6]

$$F = 1 - n + nL \tag{2}$$

where

- $n = \dfrac{e}{2kT_0\alpha}$ equals noise ratio;
- e charge on an electron;
- k Boltzmann's constant;
- T_0 temperature;
- α parameter of Schottky barrier diode (see [5]).

In the case of $\alpha = e/(kT_0)$, the noise ratio is equal to 0.5. There-

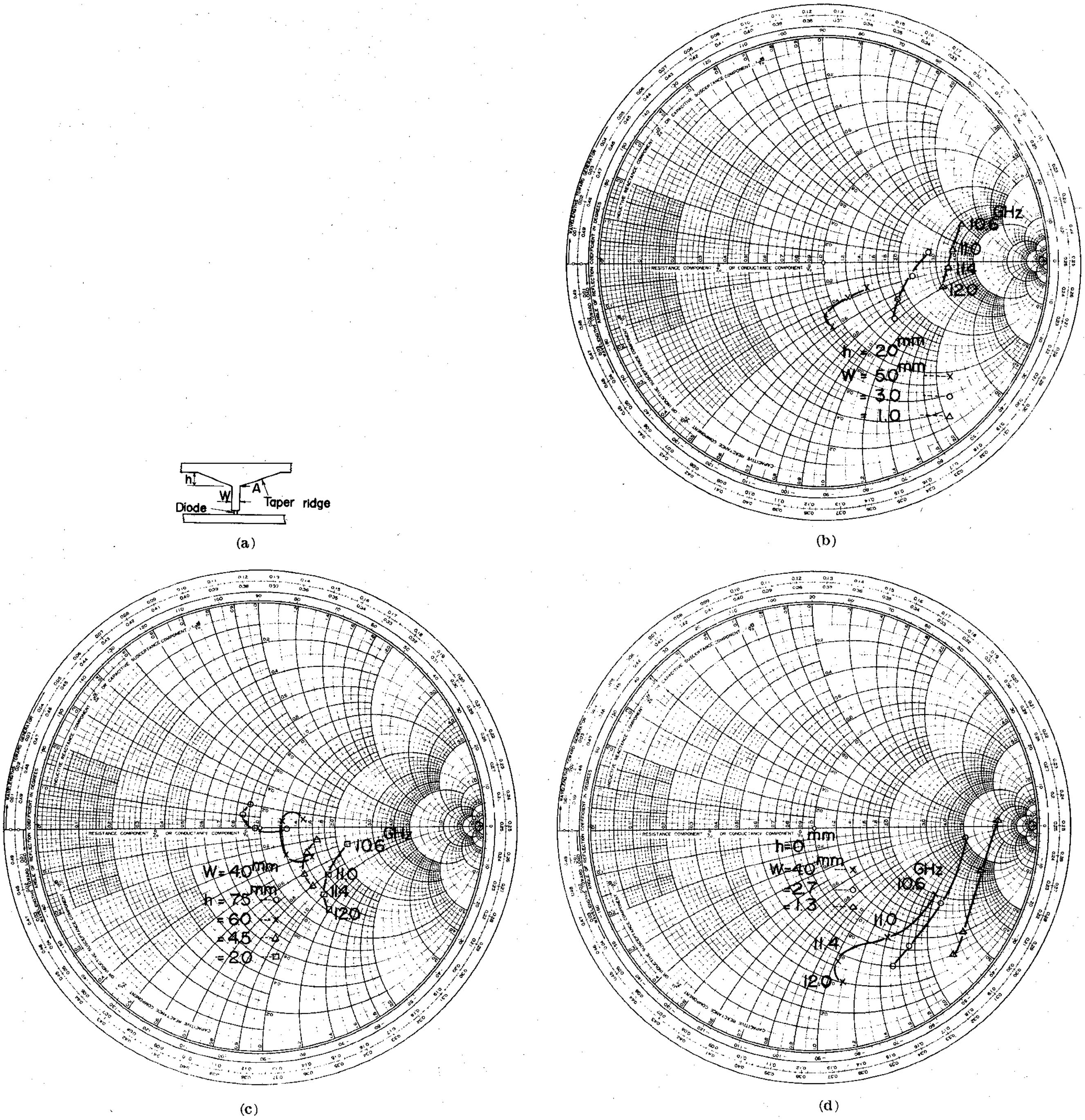

Fig. 3. Relationships between diode mount and impedance. (a) Diode mount. (b)–(d) Impedance or admittance coordinates.

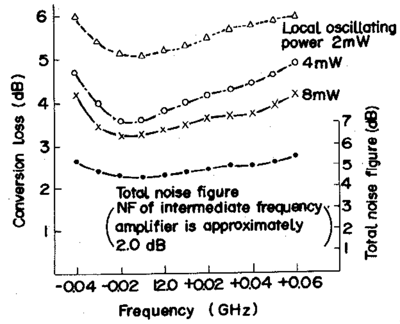

Fig. 4. Characteristics of 12-GHz converter with mounted planar circuit in waveguide.

fore, the noise figure F is given by

$$F = \tfrac{1}{2}(1 + L). \tag{3}$$

This is the same formula as in [5]. Also, the equation relating t, L, and F is

$$t = F/L. \tag{4}$$

Substituting formula (2) into (4) yields

$$t = \frac{1}{L}(1 - n) + n. \tag{5}$$

Measured values for the mixer were

$$L_t = L_r L = 2.09 \qquad (3.2\ \text{dB})$$

$$F_{if} = 1.585 \qquad (2.0\ \text{dB})$$

$$\alpha = 33.0$$

$$F_t = 2.82 \qquad (4.5\ \text{dB}).$$

Calculated values were

$$n = 0.591$$

$$t \cong \frac{1}{L_t}(1 - n) + n = 0.7855$$

$$F_t = 2.86 \quad (4.57\ \text{dB}).$$

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Efficient Numerical Computation of the Frequency Response of Cables Illuminated by an Electromagnetic Field

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Abstract—**Computationally efficient numerical methods for determining the frequency response of uniform transmission lines consisting of a large number of mutually coupled conductors in homogeneous and inhomogeneous media, and illuminated by an electromagnetic (EM) field are presented.**

I. INTRODUCTION

Conductors connecting electronic subsystems on aircraft, missiles, and ground electronic systems are generally grouped into large, closely coupled cable bundles and it is not uncommon to find bundles of over 100 conductors on modern avionics systems (for example, F-4, F-111, and F-15 aircraft) [3]. Determination of the frequency response of these large cable bundles illuminated by high-power radars as well as an electromagnetic pulse (EMP) from nuclear detonations is becoming of increasing importance [1]. Computation of the frequency response of these large bundles illuminated through an aperture such as a landing gear door can be quite costly for only one frequency. However, it is generally necessary to determine the response for many frequencies so the concern here is to minimize the per-frequency computation time for bundles consisting of a large number of conductors. Flat pack and woven flat cables are being used more frequently to connect electronic subsystems and it is not uncommon to find over 35 mutually coupled conductors in these types of cables [7].

Taylor *et al.* [2] considered the problem of two conductors illuminated by a nonuniform electromagnetic (EM) field. For two conductors, the per-frequency computation times are practically minimal. For larger numbers of conductors, we encounter per-frequency computation times which are functions of n^3 for an $(n + 1)$ conductor line so that reduction of the per-frequency computation times becomes an important concern for large numbers of coupled conductors and many computed frequencies.

We will cast the equations to be solved at each frequency into particularly efficient forms as well as introduce computational procedures peculiar to these forms which allow an efficient solution. Perhaps many of the results here will be considered fairly straightforward to obtain but our purposes will be to unify the particular formulations and also point out some perhaps not so obvious techniques for reducing computation times.

Consider an $(n + 1)$ conductor uniform transmission line consisting of $(n + 1)$ parallel lossless conductors of length $\mathcal{L}$ imbedded in a lossless nondispersive medium with the $(n + 1)$st conductor designated as the reference conductor (usually a ground plane or overall shield). The transmission line is described for the TEM mode by the following $2n$ strongly coupled complex differential equations [3], [5]:

$$\begin{bmatrix} \dot{\mathbf{V}}(x) \\ \dot{\mathbf{I}}(x) \end{bmatrix} = -j\omega \begin{bmatrix} {}_n\mathbf{0}_n & \mathbf{L} \\ \mathbf{C} & {}_n\mathbf{0}_n \end{bmatrix} \begin{bmatrix} \mathbf{V}(x) \\ \mathbf{I}(x) \end{bmatrix} \tag{1}$$

where $\dot{\mathbf{V}}(x) = (d/dx)\mathbf{V}(x)$ and ${}_i\mathbf{0}_j$ is the $i \times j$ zero matrix. The distance along the conductor structure and parallel to it is denoted by x; the complex currents $\mathbf{I}(x)$ are directed in the direction of increasing x and the ith elements of the $n \times 1$ vectors $\mathbf{V}(x), V_i(x)$, and $\mathbf{I}(x), I_i(x)$ are the complex potentials (with respect to the reference conductor) and currents, respectively, associated with the ith conductor, $i = 1,\cdots,n$. The parameter ω is the radian frequency of excitation under consideration and the $n \times n$ real symmetric constant matrices $\mathbf{L}$ and $\mathbf{C}$ are the per unit length inductance and capacitance matrices, respectively [3], [5].

The boundary conditions at the ends of the transmission line are in the form of n ports and are characterizable by "generalized Thevenin equivalents" as

$$\mathbf{V}(0) = \mathbf{E}_0 - \mathbf{R}_0\mathbf{I}(0) \tag{2a}$$

$$\mathbf{V}(\mathcal{L}) = \mathbf{E}_{\mathcal{L}} + \mathbf{R}_{\mathcal{L}}\mathbf{I}(\mathcal{L}) \tag{2b}$$

where $\mathbf{E}_0$ and $\mathbf{E}_{\mathcal{L}}$ are $n \times 1$ complex vectors of the equivalent open circuit port excitations and $\mathbf{R}_0$ and $\mathbf{R}_{\mathcal{L}}$ are $n \times n$ real symmetric hyperdominant (and therefore positive definite) matrices representing passive termination networks.

Initially, we must solve (1) and then we must incorporate the boundary conditions of (2). Differentiating the second equation in (1) with respect to x and substituting the first we obtain

$$\ddot{\mathbf{I}}(x) = -\omega^2\mathbf{C}\mathbf{L}\mathbf{I}(x). \tag{3}$$

If we define a change of variables $\mathbf{I}(x) = \mathbf{T}\mathbf{I}_m(x)$ where $\mathbf{T}$ is an $n \times n$ nonsingular matrix and $\mathbf{I}_m(x)$ represent "modal currents" then we obtain

$$\ddot{\mathbf{I}}_m(x) = -\omega^2\mathbf{T}^{-1}\mathbf{C}\mathbf{L}\mathbf{T}\mathbf{I}_m(x). \tag{4}$$

We will show that it is always possible to diagonalize $\mathbf{CL}$ for lines immersed in linear isotropic media and thus uncouple the mode currents $\ddot{\mathbf{I}}_m(x)$ by the similarity transformation $\mathbf{T}$ such that

$$\mathbf{T}^{-1}\mathbf{C}\mathbf{L}\mathbf{T} = \boldsymbol{\gamma}^2 \tag{5}$$

where $\boldsymbol{\gamma}^2$ is an $n \times n$ diagonal matrix with real positive and nonzero scalars γ_i^2 on the diagonal, i.e., $[\boldsymbol{\gamma}^2]_{ii} = \gamma_i^2$ and $[\boldsymbol{\gamma}^2]_{ij} = 0$ for

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