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# The Status of Printed Millimeter-Wave *E*-Plane Circuits

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***Abstract*—The present state of the printed *E*-plane circuit technology for millimeter-wave integrated circuits is reviewed and trends for the future development in this field are discussed. The general transmission-line properties, types of waveguide housing, discontinuity, and filter problems are discussed in detail. Several circuit components for frequencies between 20 and 170 GHz, such as p-i-n diode attenuators and switches, mixers and detectors, couplers, oscillators, and nonreciprocal devices are examined. Integrated circuit components and subsystems which use these circuits as functional blocks are presented.**

Manuscript received May 3, 1982; revised July 15, 1982.
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## I. Introduction

ALTHOUGH standard microstrip techniques may be applied to millimeter-wave circuits by mere scaling of the linear dimensions, several problems arise. These problems are connected with critical tolerances and very narrow

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conductor strips that are not completely compatible with hybrid devices. This has led Meier to the proposal of "fin-line" as a new transmission line for millimeter-wave integrated circuits to overcome or alleviate these problems [1], [2]. Before this, fin-line had only been proposed for a very special purpose of orthogonal-mode launching in circular waveguides [3], [4]. Since this basic form of fin-line did not provide isolation from the waveguide mount, items such as dc bias, IF, or modulation connections could not be accommodated.

With increasing activities in the millimeter-wave field, more attention was paid to the fin-line technique, among other alternatives, for integrated millimeter-wave circuits. During the last decade, the fin-line medium has been combined with other planar waveguiding structures like microstrip and coplanar line to form quite versatile mixed waveguide integrated circuits mounted in the $E$-plane of a metal waveguide housing.

Nearly all important circuit components have now been realized and lately even complete receive/transmit RF front-ends have been realized on a single printed circuit board. During the last years, several civil and military systems in the U.S. and Europe [5]–[10] have been equipped with key components employing printed $E$-plane circuitry, demonstrating the practicability of the new technology even under severe environmental conditions.

With the growing interest from the component designers' side, there has been an increasing activity concerning the theoretical foundations of the new medium. In the first stage, research has been carried out towards the determination of the propagation coefficient, field distribution, and characteristic impedance. During the last years, several groups have tackled the calculation of discontinuities. This new impetus eventually will lead to a more sophisticated computer-aided design (CAD) of printed $E$-plane circuits.

This paper will review the achieved status of the theoretical as well as the practical printed $E$-plane circuit know-how and will direct the attention of the reader to the remaining problems and desirable directions of future developments.

## II. $E$-Plane Printed Waveguides

$E$-plane integrated circuits usually will have to be compatible with "normal" metal waveguide techniques. As will be shown in this paper, there are some types of circuits that do not lend themselves very well to printed-circuit techniques at millimeter wavelengths. Thus, the integrated circuits will consist of an $E$-plane metal waveguide part and a printed-circuit part.

Besides fin-line, several other printed transmission lines may be used in the printed-circuit part. Fig. 1 shows four types of fin-line, namely: unilateral, bilateral, isolated, and antipodal fin-line, as well as microstrip, coplanar line, and suspended stripline. All of these may be mounted in a metal waveguide housing split in the $E$-plane. All or some of these may be combined in order to exploit the special advantages of those waveguides in special roles within a complex practical $E$-plane integrated circuit. For example, most published mixer designs use microstrip as a medium

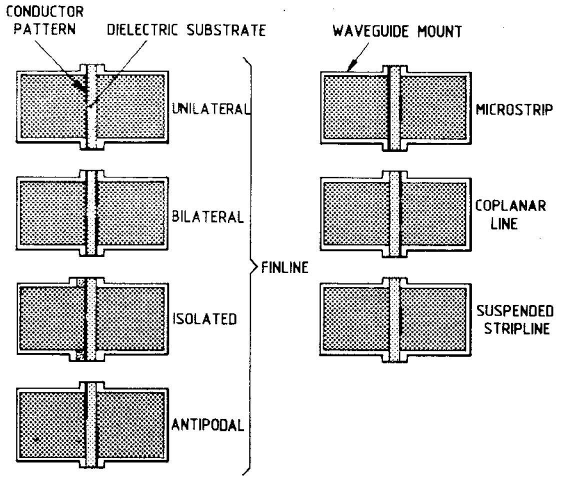



Fig. 1. Several types of fin-line and strip transmission lines supported in the $E$-plane of a metal waveguide. (From Bates *et al.* [35].)

for the IF-filter and some mixer designs combine unilateral fin-line with coplanar line to arrive at a very broad-band 180° hybrid junction (see Section III-B). Since much more is known about the stripline forms, in this section only the characteristics of fin-line shall be discussed.

### A. Wave-Guiding Characteristics

Fin-line basically is a shielded slotline where, in contrast to the conventional slotline, the field confinement is achieved by the use of a metal shield (waveguide housing) rather than by use of a thick substrate with high permittivity. If the dielectric is completely removed, in a unilateral or bilateral fin-line circuit, the resultant waveguide is the well-known ridge guide with very thin ridges, and both waveguides exhibit very similar characteristics. Our present knowledge concerning the propagation coefficients and the characteristic impedances of fin-lines stems from many experimental and theoretical investigations [2], [11]–[30].

The losses in fin-lines partly result from conductor losses (skin-effect) in the waveguide housing and, to a greater extent, in the printed metallization pattern where very high current densities are found near the slot. The rest of the loss is due to the dielectric polarization loss ($\tan\delta$) of the substrate which is placed in the electric field maximum.

Several authors [2], [11], [17], [27] have measured the insertion loss and unloaded $Q$ of sections of unilateral and bilateral fin-line. As a rule of thumb, practical fin-lines on 17-$\mu$m copper cladded RT/Duroid 5880 having relatively narrow slots of widths around 0.1 mm—0.4 mm in the $Ka$-band—exhibit attenuation of below 0.1-dB/wavelength (254-$\mu$m substrate thickness), while this figure tends to increase to 0.15-dB/wavelength in the $E$-band (127-$\mu$m substrate). The corresponding unloaded $Q$ of these lines is in the range of 200–300. Lower metallization thickness and thicker substrates increase the line losses; additional gold-plating of the fin-line circuit leads to still lower losses[1]. The quoted loss figures represent a compromise of the lower

[1] Much higher line-$Q$ may be achieved in wide-slot fin-lines (see Section III).

loss figures encountered in conventional metal waveguide and those higher figures encountered in conventional microstrip lines, topping those of fin-line by a factor of up to 3.

The determination of the effective permittivity and the characteristic impedance has attracted considerably more interest, especially from the theoretical side. Some approximate field solutions to the fin-line problem have been proposed [14], [16], [26], but these have been made obsolete by the successful development of the rigorous spectral domain analysis [18]–[20], [28]. In this approach, the original field theoretical solution of Hofmann [12], [13] was extended, which used moment method formulation and Galerkin's solution in the space domain to calculate the fields of an idealized model with zero metallization thickness, and neglecting the influences of the waveguide housing clamping region.

The spectral domain method employs a Fourier transform of the coupled integral equations in the space domain. A reduction of the number of equations (in both methods) may be achieved by an appropriate choice of expansion functions including an "edge" term (e.g., [86]) to describe the asymptotic behavior of the fields near the metal edges. It is important that this formulation does not suffer from relative convergence. Furthermore, the formulation of the spectral domain field theoretical solution has been made even more flexible and elegant by the introduction of the equivalent transmission line concept [87].

Two alternative rigorous calculation methods have been presented which take into account the finite metallization thickness of the fin-line circuit [15] and additionally, a finite longitudinal slit in the waveguide housing (clamping region) [23]. As was to be expected from experience with microstrip line calculations, there is a discrepancy between the results from this calculation and strictly planar formulations like the spectral domain approach. For practical dimensions of fin-line circuits, however, it is found that the effective permittivity $\epsilon_{\mathrm{eff}}$ and the characteristic impedance $Z_C$ are given only a few percent too high by the idealized model computations.

Some words are due here concerning the definitions of the effective permittivity and the characteristic impedance. From analogy considerations, some workers [11], [19], [23] prefer definitions compatible with the ridge-guide conventions, namely

$$\lambda'/\lambda = 1\sqrt{\epsilon_{\mathrm{eff}} - (\lambda/\lambda_c)^2}$$

and

$$Z_C = Z_{C\infty} \Big/ \sqrt{\epsilon_{\mathrm{eff}} - (\lambda/\lambda_c)^2}$$

where $\lambda, \lambda'$ are the wavelength in free space and in the fin-line, respectively, $\lambda_c$ is the cutoff wavelength of the fin-line, and where $Z_{C\infty}$ is the characteristic wave impedance of the ridge-guide substructure (fin-line without dielectric) for infinite frequency. This definition, from a field theoretical point of view, has two disadvantages: Firstly, it relies on an independent calculation of the characteristics of another waveguide (ridge-guide), and secondly, the effective permittivity in this model is the permittivity of a homogeneously loaded ridge guide. Since the real fin-line is not homogeneously filled by a dielectric, the resultant frequency dependence is only described approximately by the model. Besides that, the impedance formula is only valid for quasi-TE-modes, i.e., higher order modes and the quasi-TEM-modes of, e.g., coupled slot structures, may not be described. Alternative definitions are analogous to those used for microstrip line. Here, the relative permittivity is the squared ratio of the free space—and the guided wavelength $\epsilon_{\mathrm{eff}} = (\lambda/\lambda')^2$, while for the line impedance there is a choice of three definitions, namely, the voltage/current, the voltage/power, and the current/power definitions, which all differ for this non-TEM waveguide.

Since in many applications the line impedance is used in connection with the matching of semiconductor devices, which are contacted across narrow slots, it has been pointed out that a definition using the voltage across the slot should be most significant. Thus, Hofmann [13] used $Z_C = V/I$, where $V$ is the voltage across the slot in the plane of the metallization layer and $I$ is the longitudinal current integral over the complete fin surface, while Schmidt and Itoh [18] use $Z_C = U/2P$, where $P$ is the mode power. These definitions basically differ only for large slot widths due to the neglected portion of current on the waveguide walls in the voltage/current definition [13]. Also similar to results known from microstrip calculations, the frequency dependences of the various impedance definitions differ markedly. It is interesting to compare the theoretical characteristic impedances with experimental data.

Willing and Spielman [25] used a coaxial 50-Ω probe transition to fin-line or chip resistors contacted across the fin-line slot. Best agreement has been found with Knorr's [19] calculations, who uses the ridge-guide definition. An earlier measurement by Meinel and Rembold used a p-i-n diode bonded across the fin-line slot to load the waveguide by its well-defined small-signal impedance [30]. From the results of this experiment, the voltage/current definition appears well confirmed.

It thus seems that the characteristic impedance case is still open; especially, more experimental work on this subject is needed.

For those who wish to perform a calculation algorithm for the unilateral and bilateral fin-line on a pocket calculator or wish to include an algorithm in some optimization program for $E$-plane circuits, Sharma and Hoefer [24] have provided useful arithmetic formulas for both $\epsilon_{\mathrm{eff}}$ and $Z_C$, valid over a wide range of parameters and correct to within some $\pm 2$ percent. This "formula" approach in the past has also been helpful in the proliferation of basic know-how concerning microstrip techniques and later has been vital to CAD techniques.

It has often been stated that the dimensional tolerances of printed $E$-plane circuit housings are less tight than in metal waveguide circuits.

Theoretical analysis [21] and practical experience have confirmed that this is true, mainly due to the high concentration of fields and currents near the slot/strip structures. Only in the clamping region of the housing special care has to be taken (see Section II-B), leading to possibly tighter tolerance requirements than needed in metal waveguide circuits.

The forms and characteristics of fin-lines are so various that it would be impossible to discuss the results from many publications in detail here. One example shall suffice in an attempt to highlight the status of the theoretical techniques concerning the waveguiding characteristics of fin-lines. In Fig. 2 the cross-sectional structure of a very general printed $E$-plane transmission line with three slots (and one strip) is shown. The effective permittivities and the characteristic impedances for the quasi-TEM coplanar mode (even) and the slot mode (odd) have been calculated using the spectral domain approach [22]. This example suggests that practically every slot/strip combination for use as an $E$-plane waveguide now is within the reach of our analytical tools. As a consequence, many researchers now have directed their attention to the next difficult and next important issue of fin-line discontinuities (see Section III), and to new forms of circuit, like, e.g., fin-line on ferrite or on semiconductor substrate.

### B. Mechanical Considerations

Two different styles of housing are used in practice, as shown in Fig. 3. In Fig. 3(a), the type originally used by Meier [1], [2], the circuit board is positioned and held in place by dowels in the split waveguide block.

In Fig. 3(b), the circuit board is cut to fit in a narrow slit in the block [31]. The main difference of the two designs is in the clamping region, where the waveguide housing clamps the fin-line substrate (circuit board). Contrary to pure metal waveguide there is a considerable transverse current flow at this place, which will cause heavy attenuation and/or spurious resonances if bad electrical contacts are found here in the type of housing shown in Fig. 3(b). This means rather tight tolerances for the slit height $d$ (roughly $\pm 10$ $\mu$m) to achieve proper contacting pressure at the split block meeting faces.

In the type of housing shown in Fig. 3(a), the problem is to avoid power flow through the clamping region. Choosing the length of the clamping region as $c \approx \lambda_\epsilon/4$, where $\lambda_\epsilon$ is the wavelength in the dielectric, or incorporating a serration pattern in the circuit-board metallization or milling a choke-slot in the waveguide block, will prevent such wave propagation outside the waveguide structure.

While in the type of housing shown in Fig. 3(a) it is straightforward to make dc-connections to the fin-line pattern by wiring to the circuit-board printed pattern in the field-free area, in the type of housing shown in Fig. 3(b), connections either have to be fed through into the clamping region, or a strip of the circuit board has to be allowed to penetrate into the waveguide block similar to the type of housing shown in Fig. 3(a).

It is clear that both designs have their own advantages, but certainly the type of housing shown in Fig. 3(b) is much more space-economical than the type of housing design shown in Fig. 3(a).

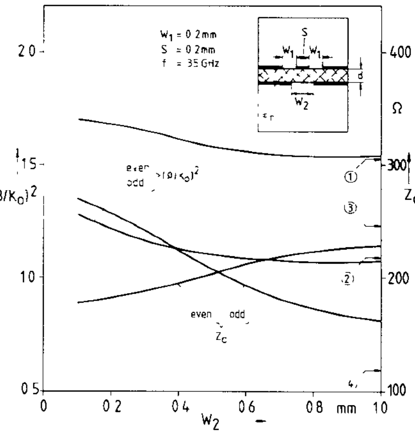



Fig. 2. Characteristics of the three slot structure versus slot width $w_2$. 1–4: Results for the unilateral case (identical with $w_2 = 3.556$ mm). (From Schmidt and Itoh [20].)

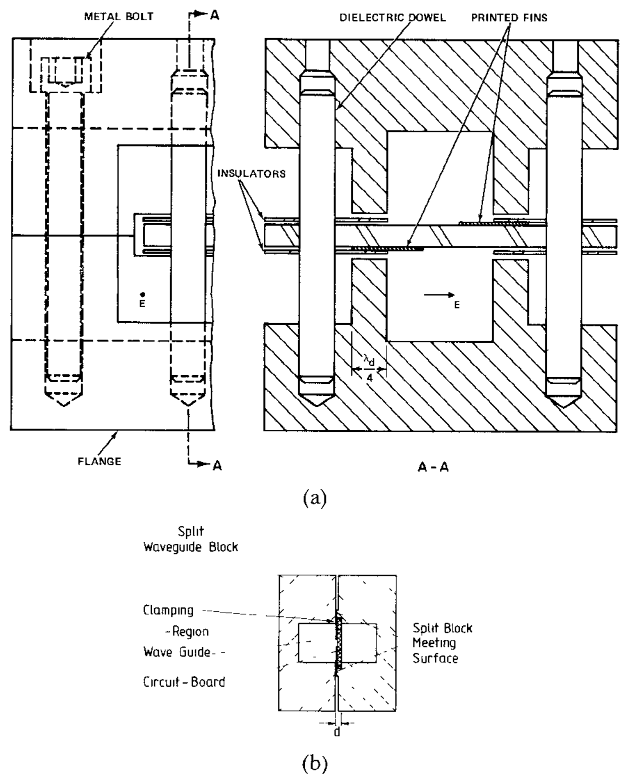



Fig. 3 Two different types of fin-line housing with (a) open, and (b) closed clamping region

As a substrate material in most practical circuits RT/Duroid 5880 (($\epsilon_r \approx 2.22$) with thicknesses of 254 $\mu$m for, e.g., $Ka$-band (26–40 GHz) to 127 $\mu$m for, e.g., $E$-band (60–90 GHz)) is used. Since this material is of rather low mechanical strength (soft substrate), there is the danger of breaking the delicate leads of beam-lead semiconductor devices, either during production handling, temperature cycling of the complete circuit, or under high $g$ forces. The answer to this problem is to use small chips of hard substrates (e.g., quartz) for mounting the semiconductor, and integrating the device into the soft substrate circuit as an upside-down overlay.

On the other hand, the whole printed circuit may be realized on a single hard substrate. While this tightens the tolerance requirements of the waveguide housing to prevent breaking of the substrate under clamping forces, such circuits have withstood acceleration levels up to 30 000 $g$'s for 2 ms, without harmful damage either to the semiconductors or to the substrate as a whole [32]. Circuits using soft substrates, on the other hand, have withstood accelerations of 2000 $g$'s for 0.5 ms without damage [33] or shocks up to 4000 $g$'s [34].

There are, however, alternatives to the use of hard substrates [34]. One is to use soft gold beam-leads in the production of the semiconductor devices rather than the conventional, more rigid, leads. Of course, since such devices are not commercially available at the moment, this is only feasible if in-house production capabilities exist.

## III. E-Plane Printed Circuit Discontinuities

The consideration of $E$-plane printed circuit discontinuities has started with investigations of the properties of the $E$-plane inductive strip as an element in printed $E$-plane filters. While Meier [36] used a completely empirical approach to characterize inductive strips printed on a dielectric substrate, Konishi [37] treated the same structure without a dielectric layer using a field expansion approach. Both investigations have led to design charts for the equivalent-circuit elements of the strip discontinuities. Using these results, band-pass filters have been realized demonstrating the feasibility of the employed equivalent-circuit approach.

Due to the dissipative losses in the dielectric and the low metallization thickness characteristic of conventional circuit board materials, the pure metal insert filter, using relatively thick sheet material, exhibits an unloaded $Q$-factor several times that of the dielectric carrier counterpart, e.g., 1700 versus 350 at 30 GHz [38] for narrow slot fin-line inductive strip filters or 800 for a wide slot filter. On the other hand, filters printed on a substrate will certainly be attractive for integrated circuits on a single circuit-board.

Few experimental and theoretical investigations concerning the various forms of filter structures have been conducted since [17], [39]. Recently, a theoretical approach has been presented by Arndt *et al.* [40], [88] which takes into account both the finite metallization thickness and the interaction of the inductive strips (higher order mode interaction). Three-resonator band-pass filters have been realized using gold plated quartz as the substrate [40]. Insertion losses of 0.25, 0.5, and 1.3 dB have been measured for scaled versions at midband frequencies of 15, 34, and 66 GHz, respectively. Pure metal insert filters (without dielectric substrate) [88] have been realized with 0.7-dB insertion loss in a three-section 1-GHz bandwidth version at 76 GHz. It has to be cautioned, though, that the dimensional tolerances of the waveguide mount in inductive strip filters are completely the same as in pure metal waveguide filters. Less tight tolerances are only permissible if narrow slot fin-line with inductive strips is used. It thus seems highly desirable that a simple fine-tuning mechanism be devised for high-$Q$ inductive strip filters using pure metal inserts.

As an example of the state-of-the-art, in Fig. 4 the

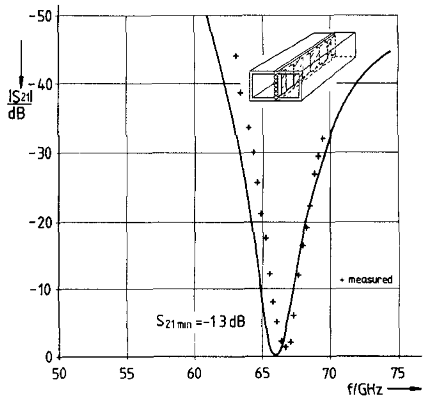



Fig. 4. Calculated and measured insertion loss of an $E$-plane filter for 66 GHz, with three resonators printed on an quartz-substrate. (From Arndt *et al.* [40].)

calculated and the measured transmission coefficient of a $V$-band filter is shown. It is believed that there will be no drastic improvement possible in the future concerning the insertion loss of this class of filters.

A discontinuity element closely related to the inductive strip is the fin-line short circuit. Konishi, again, has presented a rigorous field-expansion approach for this problem if no dielectric carrier is present [41]. If the dielectric layer is included the field theoretical solution presents a really formidable problem. Hoefer and Pic [42] have carried out a series of resonator measurements to determine the elements of an appropriate equivalent circuit for the short-circuit end-effect and have derived analytical formulas for inclusion, e.g., in a computer network analysis and optimization program. Two approximate field solutions have been proposed using a TLM-method [43] and a rectangular waveguide equivalent formulation [44]. These solutions have been made obsolete by two rigorous calculation methods: Knorr [45] has computed the eigenfrequencies of fin-line resonators using the spectral domain method and has deduced equivalent-circuit descriptions for the short-circuit from these calculations. Koster and Jansen [46] have solved the more general problem of an inductive strip of finite width which the short-circuit is a limiting case. They also use a spectral-domain approach and arrive at equivalent-circuit descriptions for the discontinuity. From their results it can be confirmed that the empirical formulas developed by Pic and Hoefer [42] are exact enough to be used for design purposes.

The latest and probably, for the progress of the art, the most important development of field solutions deals with an extended class of transverse discontinuities and three- and four-port discontinuities in fin-line. For example, a single or a double step in the slot width, symmetrical or unsymmetrical, has been described by El Hennawy and Schunemann [47] using a mode-matching technique. In this method, the field distributions of the fundamental and a number of higher order modes in the various fin-line cross-sectional regions are calculated and matched at the respective boundaries of the fin-line regions. As an example, the slot pattern, the employed equivalent circuit for the symmetry plane of the structure, and the calculated results for the equivalent-circuit elements are plotted in Fig. 5.

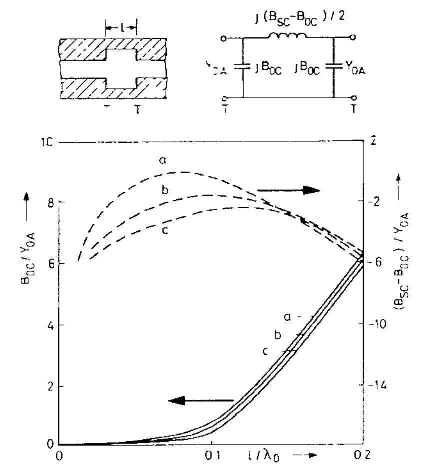



Fig. 5. The slot pattern, the equivalent-circuit, and the theoretical results for the equivalent-circuit elements of a symmetric double step discontinuity. (From El Hennawy and Schünemann [47].)

Since this method only relies on the exact calculation of the fields of the various fin-line modes (see Section II), much more activity may be expected in this field within the near future. From this work, it is hoped that analytical formulas will be derived similar to those derived by Hoefer for the fin-line propagation characteristics [24] and the end-effect [42]. Already, El Hennawy and Schunemann have demonstrated that a CAD of integrated millimeter-wave fin-line circuits is feasible using the results of their field theoretical treatment. In [48], the CAD of *Ka*-band detector circuits, p-i-n diode phase modulators, p-i-n attenuators, and Gunn oscillators has been reported.

From analogy considerations concerning microstrip discontinuity calculations, and from preliminary comparisons of theoretical and experimental results for fin-line discontinuities, it may be concluded that theoretical discontinuity models neglecting the finite metallization thicknesses in practice can be in error by several percent in the values of the resultant scattering parameters. This problem will have to be treated further in order that the method be of real value for CAD techniques.

One problem that has often been misinterpreted is connected with resonance fields in the waveguide housing excited by fin-line discontinuities [49]. Such modes occur, e.g., in directional couplers with T- and Y-junctions of the metal waveguide housing, where the fin-line slot pattern deviates from certain symmetry planes. Since this problem is common to all planar waveguides with metal shielding, the methods for the suppression of the spurious resonances are similar, namely, to reduce the cross-sectional dimensions of the junction or to disturb the main constituents of the resonance modes by, e.g., conducting inserts in the metal waveguide channel.

## IV. *E*-Plane Millimeter-Wave Circuits

Since printed *E*-plane waveguides are basically planar slot and strip media, it is clear that from the beginning the idea behind the new technology was to implement semiconductors in beam-lead or chip form. Consequently, the first integrated *E*-plane circuits were p-i-n diode attenuators, mixers, and detectors. Other circuits, such as oscillators, have not performed as successfully as these, partly because the corresponding semiconductor elements do not conform to the slot/strip structure as do beam-lead devices (they are only available commercially as packaged devices).

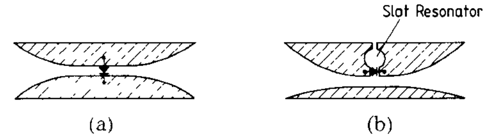



Fig. 6. Sketch of the (a) parallel-type and the (b) series-type p-i-n diode attenuator.

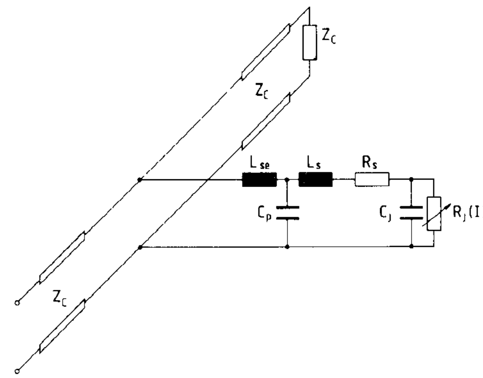



Fig. 7. Equivalent circuit for fin-line p-i-n diode attenuator. (From Meier [11].)

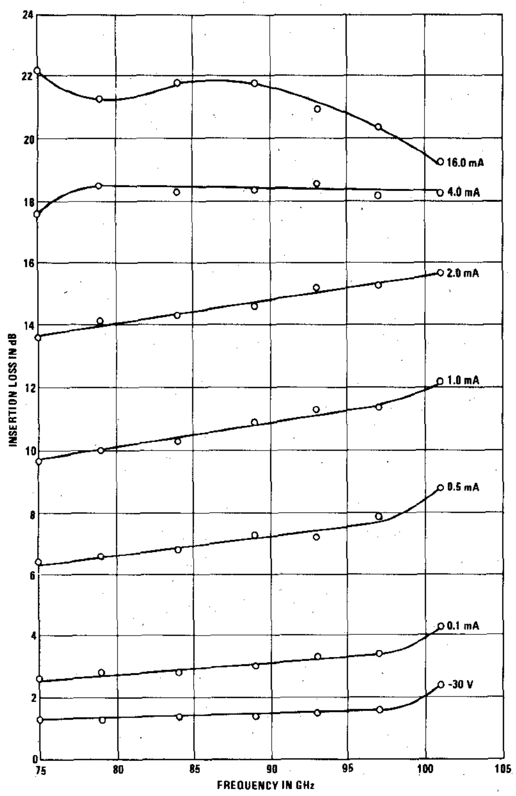
### *A. p-i-n Diode Attenuators and Switches*

p-i-n diode attenuators may be designed as parallel-type circuits (Fig. 6(a)), where the diode is shunt-mounted across the fin-line slot and as series-type circuits (Fig. 6(b)), where the diode is series mounted, backed by a slot resonator.

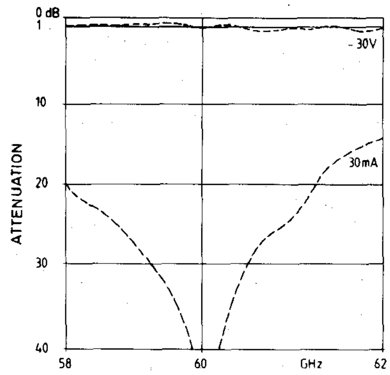
An equivalent-circuit of a parallel-type p-i-n diode attenuator has been determined experimentally by Meier [11], [36] (Fig. 7). While the characterization of the semiconductor device is well known, the impedance levels encountered in such a circuit are worth mentioning. Using conventional unilateral fin-lines with narrow slots compatible with beam-lead dimensions, the line impedance $Z_0$ is found in the range of 150–200 Ω, while depending on the manufacture of the diodes, the series inductances and the parallel capacitances represent reactances of some few ohms and several hundred ohms, respectively. The ratio of the capacitive reactance of the "off"-state diode (no current) and the inductive reactance of the "on"-state diode (current flowing), establishes the limits for the dynamic range of the electronic attenuator. Using commercially available p-i-n diodes, the insertion loss of single-diode circuits may be varied from about 1 dB to a maximum of about 15 dB nearly frequency independent over waveguide bands from *X*-Band to as high as 110 GHz.

The series-type attenuator has been devised in order to tune out the reactance of the diode, thus achieving a pole in the attenuation.

In order to both decrease the "through"-insertion loss per diode, and increase the maximum attenuation, several p-i-n diodes may be used in one circuit, with the diodes




(a)




(b)

Fig. 8. Insertion loss of (a) a two-diode parallel-type p-i-n diode and (b) series-type p-i-n attenuator using fin-line. ((a) from Meier [53], (b) from Callsen [55].)

spaced approximately one quarter-wavelength and with appropriate impedance steps between the diodes. Several authors have presented p-i-n diode attenuators using two up to four diodes in a parallel- or series-type circuit [30], [38], [39], [50]–[55], [89]. Both basic types of p-i-n diode attenuators have been realized up to 110 GHz with commercially available diodes. As an example in Fig. 8, the insertion loss of a two-diode parallel-circuit attenuator for the *W*-band, (Fig. 8(a)) and for a two-diode series-circuit attenuator for the 60-GHz band (Fig. 8(b)) are shown. Using a p-i-n diode with the dc-current adjusted in such a way that the diode real impedance is approximately equal to the fin-line characteristic impedance, absorbing attenuators have been realized also [54].

Other publications [8], [35], [39], [52], [89] have shown

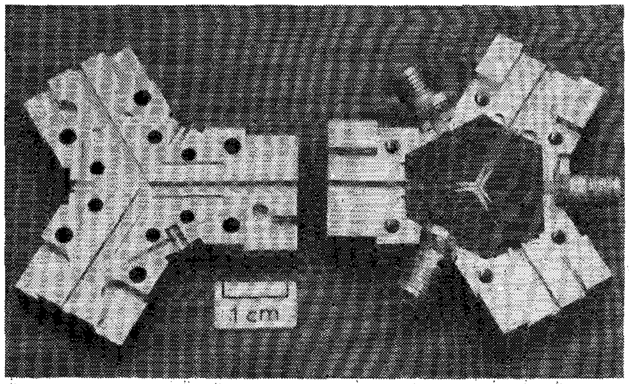



Fig. 9. A SPDT switch for 80–90 GHz using two p-i-n diode attenuators in a fin-line *Y*-junction. (From Bates *et al.* [35].)

that two or three fin-line attenuators can be combined with a 3- or 4-port fin-line junction to form SPDT or SP3T switches. As an example, in Fig. 9 a SPDT switch is shown for 80–90 GHz exhibiting 2.5-dB/20-dB insertion loss [35], but full waveguide bandwidth may also be achieved.

The switching time achievable with p-i-n diode switches is known to depend mainly on the construction of the diodes. Using "fast" p-i-n diodes, few tens of ns are achievable; using Schottky-barrier mixer diodes instead reduces this time to below 1 ns [30].

For the future, progress in the p-i-n diode circuit field may be expected from two sources: firstly, improved devices with smaller size, higher switching speed, closer tolerances, and lower parasitic reactances; and secondly, from computer-aided design.

Using the developing field theoretical techniques, (see Section III), multi-diode structures may be optimized. This approach may be rewarding since there are some basic properties of p-i-n diode circuits that are not yet fully understood; e.g., the limit of the maximum attenuation (isolation) in multi-diode circuits is influenced by interaction of evanescent field modes excited by the diodes and thus the spacing of the diodes as well as the dimensions of the waveguide channel have a strong influence.

As an area of application where research so far has been totally neglected, the use of p-i-n diodes as power limiters for millimeter-wave frequencies has been identified. With increasing power levels in transmitter devices, such circuits may gain some importance in the near future.

### *B. Mixers*

Printed *E*-plane circuit mixers have first been realized for *Ka*-band by Meier [5], [50] as a single-ended mixer and Gysel [56] as a balanced mixer.

The single-ended mixer is a direct transformation of known metal waveguide techniques into planar fin-line techniques (Fig. 10) and basically draws its attractiveness from the implementation of beam-lead diodes: The low junction capacitance of modern beam-lead mixer diodes and the planar circuit approach permit mounting of the diodes with minimum additional reactances, making wideband matching to the LO and RF ports possible. The balanced mixer design takes it a step further: By ap-

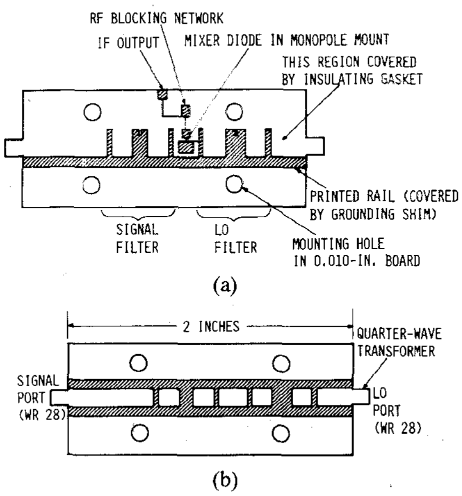



Fig. 10. Single-ended mixer for *Ka*-band with integrated printed *E*-plane filters for image rejection and LO suppression. (a) First side and (b) second side (From Meier [5], [50].)

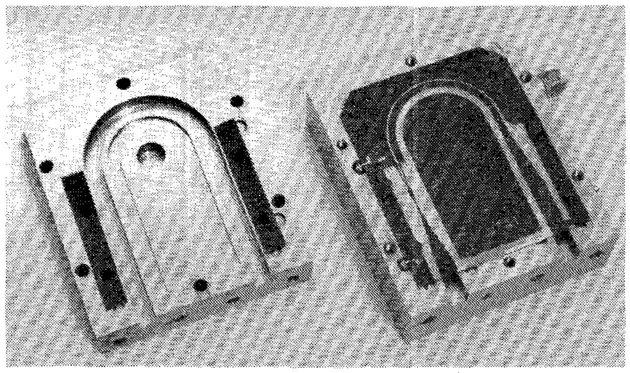


Fig. 11. *E*-plane balanced mixer for *Ka*-band. The LO-port (right) has a transition from rectangular metal waveguide to suspended strip line and to (unbalanced) coplanar line, while the signal (left) is coupled via (balanced) unilateral fin-line to the mixer diodes. The signal port is matched using a fin-line transformer section (From Bates *et al.* [35].)

propriately joining fin-line as a balanced transmission line for the signal and coplanar line as an unbalanced transmission line for the LO, a broad-band 180° hybrid is formed that is ideally suited for the implementation of two beam-lead diodes.

Several modified and scaled designs for frequencies up to 100 GHz have been reported since then [8], [35], [39], [51], [57], [58], [90]. As an example, a design that has been used for 35-GHz and 85-GHz receivers is shown in Fig. 11. The waveguide housing has been produced either in aluminum or in surface metalized ABS-plastic with total weights of 140 g and 70 g, respectively.

A unique double-balanced mixer design [59] operating over more than 10–40 GHz has been awarded the European Microwave Prize at the European Microwave Conference 1980. It uses bilateral fin-line for the LO and antipodal fin-line for the signal. Four beam-lead diodes are connected across the resultant balanced/balanced junction on both sides of the circuit board. The IF port is coupled to the fin-lines via a balanced stripline perpendicular to the plane of the *E*-plane circuit, thus essentially establishing a three-dimensional circuit.

Balanced mixers have also been built using a 3-dB directional coupler as a 90° hybrid for both LO and signal [60], [61]. The directional coupler can use either *E*-plane probe coupling or continuous coupling of two slots in one waveguide channel (coupled fin-lines), (see Section IV-E).

The achieved conversion losses and noise figures of the reported mixer circuits mainly depend on the cutoff frequencies of the diodes used. The state-of-the-art conversion loss figures vary from 5 dB at 35 GHz to 6.5 dB at 94 GHz, using commercially available dc 1309 diodes. The best noise figures reported so far are 6 dB (SSB) at 35 GHz and 7.5 dB (SSB) at 85 GHz including IF contribution [91], both using specially fabricated MOTT diodes [62]. A recently published fin-line subharmonic mixer [92] has 10-dB conversion loss at 94 GHz.

More detailed information on the design and the performance of recent 60-GHz and 94-GHz *E*-plane mixer developments by Meier [34], and Menzel and Callsen [63] is presented in separate submissions to this Special Issue. Due to the broad-band hybrids, most of the reported mixer designs are inherently broadband, i.e., with IF-ranges from near zero to more than 10 GHz.

Problems arise here from the difficult matching of the mixer diodes to the signal. The pumped mixer diodes present impedance levels around or below 50 Ω, including parasitic reactances. That means the signal input match without a transformer is found in the range of 3 to 10 dB. A broad-band impedance match may only be achieved by using a suitable transformer section very close to the diodes. There still have been found no satisfactory methods to make such a transformer variable or adjustable, firstly, due to the fact that variable discontinuity elements, like plugs penetrating through the metal waveguide housing, only disturb the fin-line fields effectively if they come very close to the slot. Secondly, discontinuity slot patterns (fin-line transformer) may only be varied after a complete disassembly of one of the housing blocks, or even the complete substrate has to be changed with a varied line pattern. In this context it also appears that it is extremely difficult to even measure exactly the complex reflection coefficient of the mixer signal port. These difficulties basically are connected with relatively high inaccuracies in the determination of the reference plane when tapered transitions to standard metal waveguide measurement equipment are used. Since this measurement aspect seems vital to the use of CAD techniques for matching or image-rejection applications of mixers, more activity in this field should be expected for the future.

## C. Detectors

As the third type of printed *E*-plane circuit, the fin-line detector circuit was first investigated by Meinel. Experiments with bonded chip-diodes have yielded high sensitivity detectors only in short-circuit tuned versions due to the high bond wire reactances at millimeter wavelengths. High, broad-band sensitivity has been realized only using commercially available zero-bias beam-lead diodes [64]. These are soldered across the slot of a unilateral fin-line which

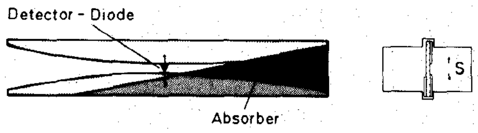



Fig. 12. Basic setup of fin-line detector circuit. (From Meinel and Schmidt [64].)

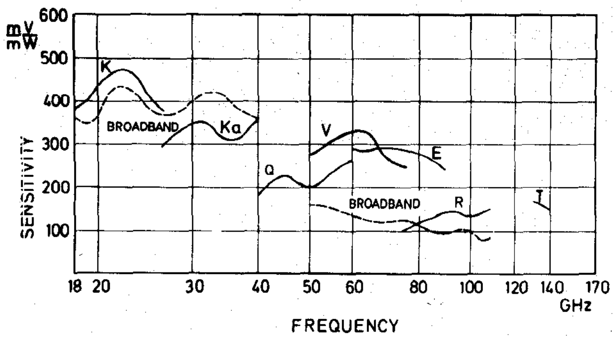



Fig. 13. Measured sensitivity of fin-line detector units (single-band and double-band) versus frequency. (From Meinel and Schmidt [64].)

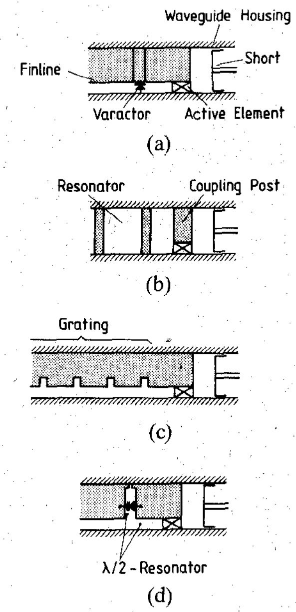



Fig. 14. Various alternatives for printed $E$-plane oscillator circuits.

incorporates tapered transitions to metal waveguide on both sides. In Fig. 12, the basic setup for a fin-line detector is shown. In this circuit, the tapered transition behind the diode is painted with absorber material to yield a broadband termination. Such detectors for a range of bands extending from 18 GHz ($K$-band) to 170 GHz ($T$-band) are presently available commercially from two companies (AEG-Telefunken and Mullard (Philips)). In Fig. 13, measured sensitivities of several detector units in these bands are plotted versus frequency.

Unfortunately, no real millimeter-wave beam-lead zero-bias detectors are available commercially at the moment. Devices designed for up to only 18 GHz have to be used instead, resulting in problems with impedance match and sensitivity at higher frequencies. It was found that such diodes, depending on manufacturing tolerances, may possess parallel capacitance values up to 100 $f$F, making impedance match very difficult with fin-line characteristic impedances around 150 Ω. Improved performance in sensitivity as well as VSWR may be expected for the future if semiconductor manufacturers come up with smaller low capacitance diodes in beam-lead form, as indicated by Anand [85]. Until then, more or less narrow-band matching of the diodes may be employed using, e.g., fin-line transformers [48].

### *D. Oscillators*

Gunn-elements and IMPATT-diodes may be mounted in planar $E$-plane circuits to realize oscillators. Several circuit designs have been proposed [51], [52], [65]–[68], a number of which are sketched in Fig. 14. The circuit of Fig. 14(a) uses a metal waveguide-to-fin-line transition and a backshort to match the source impedance of a Gunn element. This circuit presents a real conductance to the active element which is equivalent to the characteristic conductance of the fin-line ($Y_C \approx (1/150) \cdots (1/200)S$). At lower frequencies, using conventional packaged elements, the source impedance of the Gunn elements is in the range of $-10$ to $-30$ Ω, while at frequencies above 40 GHz the package parasitic reactances transform these values to a level of some $-100$ Ω. Thus, the circuit in Fig. 14(a) may be used for wide-band tunable oscillators at higher millimeter-wave frequencies. Cohen and Meier [65] have reported Gunn oscillators using this circuit at 60 and 70 GHz, employing GaAs and InP-Gunn elements to produce 60 and 50 mW, respectively. A varactor has been integrated in this circuit as sketched in Fig. 14(a) to give FM capability.

A circuit capable of stronger impedance transformation than the first one is sketched in Fig. 14(b). It uses a planar post structure to transform the source impedance of the active elements and incorporates an $E$-plane planar filter to realize a cavity stabilized oscillator [66]. Such a circuit in practice is difficult to adjust properly since the transformer post dimensions (width and distance to cavity) depend critically on the reactive components of the active elements which may vary considerably due to manufacturing tolerances and depend on the form of the connection of the printed circuit to the packaged element (direct soldering or bond wiring). The same applies here concerning the problems of creating a variable or adjustable means of tuning as has been discussed in context with mixer circuits (Section IV-B).

The "grating" oscillator [67], due to Hofmann in Fig. 14(c), overcomes this problem basically by presenting the active element a strongly varying impedance with frequency. The frequency of oscillation is determined within a narrow band by the distance of the stubs (stopband) while the transformation ratio is mainly determined by the length and number of the stubs. This structure is an adaptation of the distributed reflector LASER design known from integrated optics. Its main disadvantage obviously results from its distributed character: The efficiency of this oscillator circuit due to dissipation losses in the

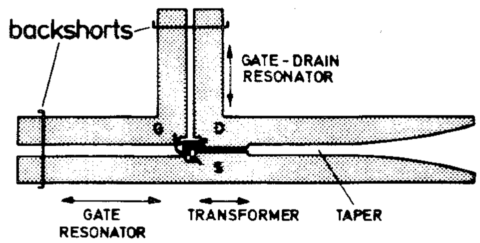



Fig. 15. Fin-line FET oscillator using a 0.5-$\mu$m gatewidth transistor to yield 2 mW at 30 GHz. (From Meinel [71].)

"grating" structure may be considerably lower (10–30 percent) than in more concentrated circuits, as, e.g., in Fig. 14(d), where a single stub is used. This circuit acts as a half-wavelength resonator between the active element and the stub short-circuit with an output-load tap approximately midway [68]. Using this circuit, 500-mW pulse and 130-mW CW have been produced at 35 GHz using a 150-mW minimum power Gunn-element MA 47177. Since the design consists of a single tuned circuit, it is straightforward to incorporate wide-band electronic tuning as sketched in the figure.

Up to now only one report is known to the author describing IMPATT-diode oscillators utilizing printed $E$-plane circuits. The reason for this may be two-fold: Firstly, the relatively low quality factor of the printed waveguides may prohibit effective impedance transformation to the extremely low levels necessary for IMPATT-diode oscillators. Secondly, IMPATT-diode oscillators are known to respond very critically to mismatched loads, thus making it necessary to tune the circuit especially carefully. Experiments by Sicking [32], employing a grating structure for impedance match of an IMPATT diode at 30 GHz, have yielded oscillators with only low power and extremely poor repeatability of the oscillation condition.

Oscillators for frequencies above 70 GHz, using GaAs Gunn elements have to rely on the harmonic mode of operation, as known from several publications describing standard metal waveguide oscillators [69]. In such a circuit, a resonator for the fundamental frequency has to be provided in a waveguide below cutoff. The harmonic output power propagates in the waveguide via an impedance transformer to the extremely low harmonic source impedance. Although this concept may be realized as a printed $E$-plane circuit as well, again in this instance, the problems connected with the adjustment of the printed circuit and the limited useful impedance transformation range have prevented experiments in this direction from being successful. A combination of fin-line with a radial disk resonator inserted between the top of the Gunn-element package and the printed fin has been reported by Cohen [70] to have yielded 5 to 10 mW between 80 and 100 GHz.

FET oscillators may also be realized as a printed $E$-plane circuit. Fig. 15 shows the circuit pattern of the first fin-line FET oscillator for 30 GHz reported by Meinel [71]. The planar structure of the transistor makes it ideally suited for a combination with fin-line, even though the low impedance level of the semiconductor device would call for impedance transformation schemes more similar to those used for Gunn oscillators.

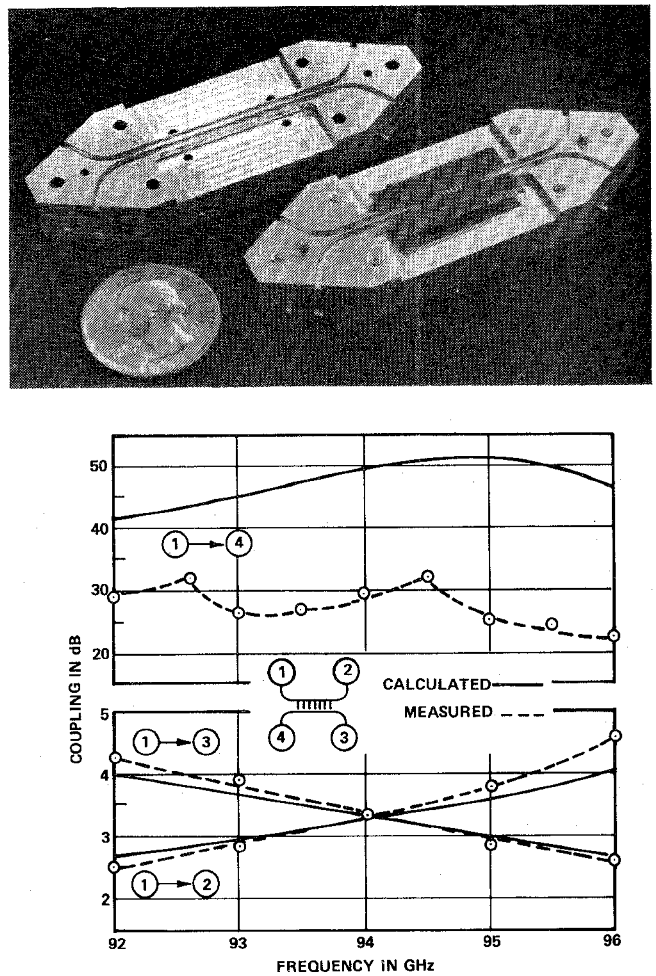



Fig. 16. 94-GHz printed $E$-plane hybrid coupler. (From Meier [74].)

The future progress in the oscillator field may be strongly connected to the development of practical tuning and adjustment techniques for printed $E$-plane circuits. At the moment, due to the much higher flexibility of standard metal waveguide oscillator designs, printed $E$-plane circuit oscillators are not feasible or only little attractive in several applications, especially above 60 GHz.

### E. Directional Couplers

Directional couplers have been employed in printed $E$-plane mixers [60], [72], [34], a multiplexer [51], and in a unique p-i-n diode phase modulator [73]. Cohen and Meier [38], [65] have investigated empirically capacitive $L-C$ and loop-probe coupling of two fin-lines separated by a common metal wall. Recently, a 94-GHz 3-dB hybrid has been realized using computer-aided design techniques on the basis of equivalent-circuit data gained by experiment [74]. The coupler employs seven printed capacitive probes on a RT/Duroid substrate (Fig. 16) yielding cross-over coupling with high directivity and additional insertion loss of 0.3 dB. Alternative forms of couplers use a double-slot structure in a single waveguide housing (continuously coupled fin-lines, (Fig. 17). In this waveguide structure, the

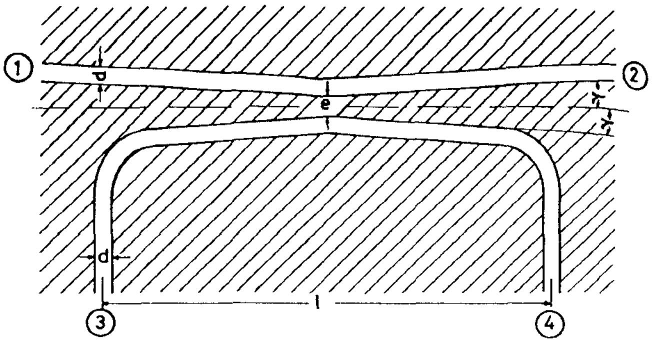



Fig. 17. Slot pattern of coupled fin-line hybrid. (From Kpodzo *et al.* [73].)

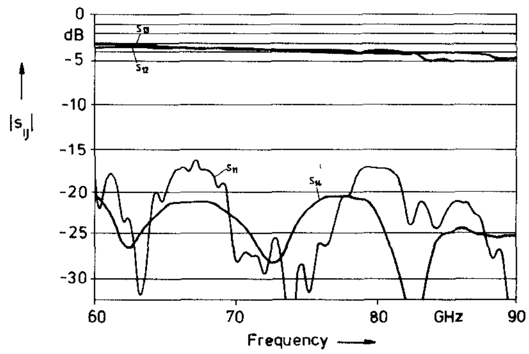



Fig. 18. Measured scattering coefficients of a *E*-band 3-dB hybrid employing coupled fin-lines. (From Solbach [75].)

quasi-TEM (even) mode interferes with the (odd) slot mode to produce practically full band coupling with insertion losses directly corresponding to the attenuation of the coupled fin-lines (about 0.15 dB/λ at 75 GHz), plus the attenuation due to the tapered transitions to the metal waveguide ports [75] (Fig. 18).

Other forms of couplers have been derived from de Ronde's hybrid microstrip coupler. Two designs have been reported operating at *Ka*-band [76], [77]. These circuits are very compact using quarter-wavelength coupling strip/slot combinations. There are, however, technological problems when these circuits are to be scaled to higher millimeter-wave frequencies due to the narrow line dimensions.

In the future, it may be expected that more use will be made of the existing directional coupler designs, especially in integrated circuits.

### *F. Nonreciprocal Devices*

As the printed *E*-plane circuit technology has matured, the need for nonreciprocal devices for integration in oscillator circuits (isolators) or RF front ends (circulators) has become stronger.

A first proposal for a field-displacement ferrite isolator compatible with fin-line has been published by Solbach and Beyer [78], and a field theoretical method is presented, useful for the optimization of the mixed layered fin-line structure.

In Fig. 19, the cross-sectional structure of an experimental model for the *X*-band is shown together with the approximate field distributions for the forward direction

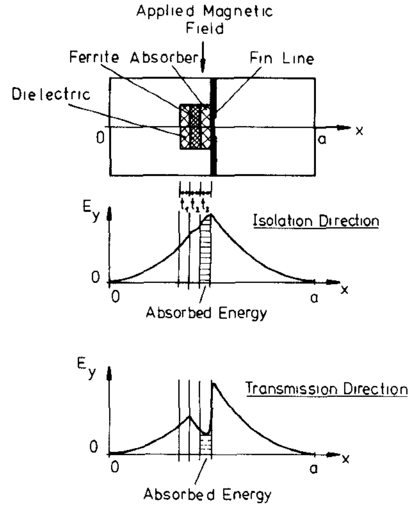



Fig 19. Cross-sectional structure and the desired field patterns for operation of the fin-line field displacement isolator. (From Beyer and Solbach [78].)

(transmission) and the reverse direction (isolation). Basically, the structure is similar to the standard metal waveguide isolator with the difference that the gyromagnetic material is placed near the center of the waveguide rather than near the narrow wall of the metal waveguide. Due to the high field concentration near the slot, the magnet needed to bias the ferrite slab may be incorporated within the waveguide housing to yield a compact "drop-in" unit for integration with other printed *E*-plane circuits.

A resonance isolator using the same principle setup has been described by Beyer and Wolff [79] with a figure of merit of 1 dB/17 dB over a 4-GHz band around 31 GHz. The fundamental difficulties prohibiting the application of such isolator structures in practical millimeter-wave circuits is due to the fact that hexagonal ferrites are not available at the moment. This material with its strong anisotropy field is needed in millimeter-wave isolators to reduce the bias field requirements from 8 and 20 k Gauss (30 and 100 GHz, respectively) to practical values around 1 to 3 k Gauss.

For the future, it may be expected that the companies presently carrying out research programs in this direction will overcome the technological problems so that the fin-line isolator eventually will become feasible for integrated millimeter-wave circuits.

Until that time, junction circulators may fit the gap, since they do not suffer from exacting bias-field requirements. Meier [5] has experimented with a fin-line Y-junction circulator achieving 1.3-dB insertion loss versus 21-dB isolation at 71 GHz. Braas and Schieblich [80] have presented three configurations for mounting the ferrite disk resonator in a fin-line Y-junction (Fig. 20). The fin-line may incorporate transformer sections to increase the bandwidth of the circulator to over 10 percent in a *X*-band model with 0.5-dB insertion loss.

It has been observed, though, that such a structure is very sensitive to misalignment with respect to the symmetry axis of the ferrite disk and the Y-junction; the disk has

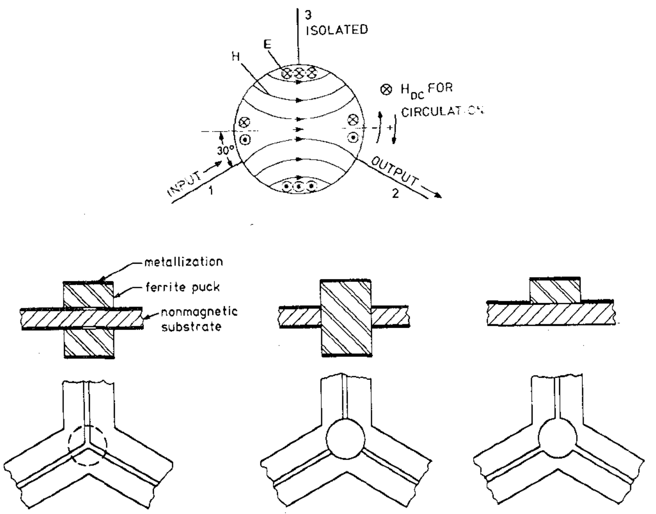



Fig. 20. Schematic diagram of field patterns in the ferrite resonator and some configurations for mounting the resonator disk in bilateral and unilateral fin-line. (From Braas and Schieblich [80].)

to be centered with a tolerance of some $10^{-3}$ referred to the wavelength to achieve proper symmetric operation of the circulator. It is this extremely high accuracy requirement which, for the moment, makes the ordinary metal waveguide *E*-plane junction circulator [81] much more attractive for integration in *E*-plane circuits. *H*-plane circulators are not compatible with *E*-plane circuit design due to their different branching planes. Bandwidths of 2–3 GHz have been achieved in *Ka* up to *E*-band with insertion losses between 0.3 to 0.5 dB.

Recently, an efficient way to combine fin-line with a ferrite junction having a field distribution very similar to that known from microstrip circulators has been found [93]. This idea leads to broad-band performance and may be interesting for future developments.

## V. Millimeter-Wave Integrated Circuits

Considerable reductions in size and weight of millimeter-wave subsystems may be realized by integrating separate circuits on one substrate in a single housing. Additionally, in high-volume production, a considerable cost advantage may be realized in spite of the higher development costs. From an "electrical" point of view, bringing circuits closer together is useful, since lower transmission losses and higher over-all bandwidths are achievable.

Several successful attempts have been made in the past years to integrate two to five separate circuit functions on a single printed *E*-plane circuit substrate. Actually, even the mixer circuits discussed in Section IV-B may be termed "integrated circuits", since they integrate a 180° hybrid and the IF low-pass filter. In truly integrated circuits, the mixer function is coupled with a p-i-n diode attenuator for STC purposes, or with a Y-junction SPDT switch with integrated reference load for Dicke-switch radiometer applications, as produced by a group at Philips Research Laboratories [35].

Fig. 21 shows two halves of the machined waveguide housing and the IF amplifier in a separate case. A wave-

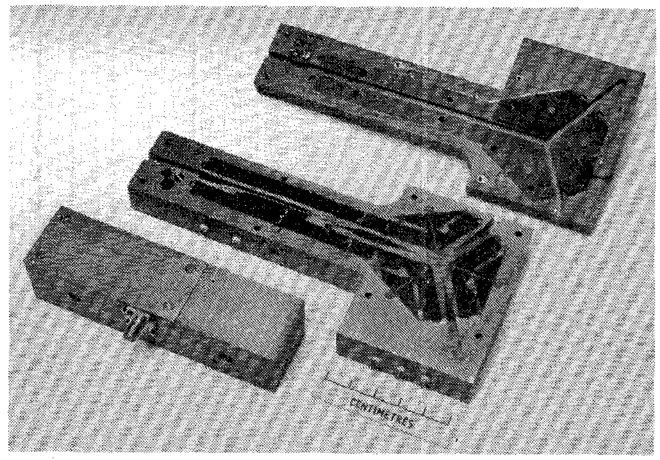



Fig. 21. Printed *E*-plane radiometer head for 35 GHz. (After [35].)

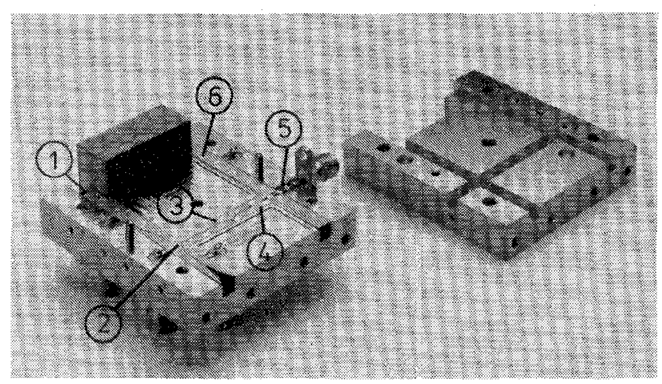



Fig. 22. Printed *E*-plane pulse radar head for 35 GHz.

guide Gunn oscillator (not shown) bolts onto the left of the assembly. The subsystem exhibits a DSB noise figure of 5.5 dB at 35 GHz with the insertion loss of the p-i-n diode switch included.

Other groups have realized combinations of the mixer function with the LO Gunn oscillator for frequencies from 15 GHz to 40 GHz [72], [82], [83] or even mixer function, LO and p-i-n switch [84]. The highest level of integration achieved so far by our group at AEG-Telefunken realizes a complete medium-distance (5–1000 m) pulse radar frontend for 35 GHz on a single substrate [94]. Fig. 22 shows a photograph of the subsystem.

The module consists of a transmit Gunn-pulse oscillator (1), a p-i-n diode SPDT-switch (2) as TR-switch, a p-i-n diode STC (3) for the balanced mixer (4) with IF-filter (5) and local oscillator (6).

The pulse oscillator produces 500 mW using a MA 47177 Gunn element with 10-V bias pulses of 10–100 ns at a pulse repetition rate of 200 kHz. The SPDT switch incorporates two diodes in the oscillator arm, and three diodes in the receiver arm, giving a high isolation between transmitter and receiver greater than 30 dB and, together with the STC, providing a sensitivity control for close-range reflections. The LO pumps the balanced mixer at about 20 mW to yield a conversion loss of 7.5 dB at an intermediate frequency of 500 MHz (conversion loss includes insertion loss of switch and STC).

For the future, more "medium scale" integration of RF subsystems may be expected with a) an increase in operating frequencies towards 94 GHz and b) more incorporation of classical metal waveguide passive components like circulators. In circuits, for frequencies above 60 GHz, for the short term, there will probably be no alternative to metal waveguide oscillators, so that these components will not be integrated.

There may as well be progress made in the direction of integration by employing separate single-function circuits combined in one housing. Such an intermediate step would enable separate testing and optimizing of the circuits but appropriate connection methods have to be found if "lossy" tapered transitions to metal waveguide are to be avoided.

## VI. Concluding Remarks

There are still some open questions regarding the limits of printed *E*-plane circuits. Experts in the field are not sure about the practicability of this technology above 100 GHz. While Meier [34] and Meinel and Schmidt [64] have successfully produced fin-line circuits up to 120 GHz and 170 GHz, the limiting factors may turn out to be a) the printing accuracy needed on the order of $\pm 10$ $\mu$m and b) the physical dimensions of the beam-lead semiconductor devices. At AIL, mixer diodes have been developed [34] that are physically smaller than commercial devices and could be mounted in fin-line housings as small WR-4 (WG-31) for potential applications up to 260 GHz. Such a reduction in size is also found necessary for, e.g., p-i-n diodes which appear to have rather long "intrinsic" zones in the order of several percent of the wavelength around 100 GHz. Other semiconductor devices are not yet commercially available in beam-lead form (e.g., tuning and multiplier varactors), making efficient printed *E*-plane circuit design very difficult since packaged or bonded chip devices with their high parasitic reactances have to be used instead. Other properties of the new waveguiding technique, especially of fin-line, have not yet been investigated. For example, it is not known what the limits are concerning high-power transmission.

Nevertheless, it is clear from what has been presented in this paper that the printed *E*-plane technology is right in the transition from being merely a "promising" new approach, to becoming a standard technology for millimeter-wave circuitry. The range of applications of the new technology includes communications and radar equipment up to the 90-GHz band at present. The 140-GHz band still presents a challenge but high-volume production applications (e.g., military terminal guidance and small-size radars) demand compact, inexpensive millimeter-wave components which may well be realized in the future using printed *E*-plane techniques.

## Acknowledgment

The author wishes to thank his group leader and his collegues at AEG-Telefunken for many helpful discussions in connection with this work. Thanks are due also to many collegues at other companies and universities for contributed photographs and drawings and for helpful comments and views.

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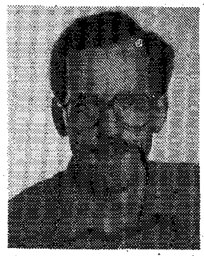

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# 50-GHz IC Components Using Alumina Substrates

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*Abstract*—**This paper discusses the feasibility of employing alumina substrates instead of fused quartz or sapphire substrates in millimeter-wave integrated circuits (IC's), an attractive prospect since alumina boasts considerable advantages over either of the other materials.**

**Millimeter-wave 50-GHz components were developed on alumina substrates. These included passive components, a mixer, an ASK modulator, and an oscillator. Empirical results for both oscillator stabilization using a dielectric resonator and a new application of a GaAs FET in a millimeter-wave oscillator-doubler are presented.**

**Examples of integrated systems using millimeter-wave IC's are also presented. These systems include a compact Doppler radar front-end for an automobile ground-speed sensor, and a transmitter/receiver for digital radio equipment. All of them are fabricated on alumina substrates.**

## I. Introduction

IN 1979, THE WORLD Administrative Radio Conference adopted a frequency utilization plan for the millimeter-wave spectrum beyond 40 GHz. This plan opened the door to a broad range of commercial applications for millimeter-wave radio in the 1980's.

Integrated circuit (IC) technique in the millimeter-wave range will be the key to achieving compact and cost-effective systems. Until now most of the millimeter-wave IC's have employed fused quartz, sapphire, and/or other substrate materials, but it would be difficult to commercialize millimeter-wave IC's utilizing these materials. Quartz requires special handling because of its low mechanical strength. Sapphire has high mechanical performance but is very expensive. Other materials such as copper-clad are easy to handle, but it is difficult to accurately form tiny IC patterns and thin-film resistive materials on the substrate. Alumina ceramic material is predominantly used as the IC substrate in the microwave range.

This paper shows that alumina can be used for millimeter-wave IC substrates. The empirical design equation developed for microwave frequencies together with basic properties of microstrip lines on substrates can be extended to the millimeter-wave frequencies.

Manuscript received May 5, 1982; revised August 2, 1982.
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