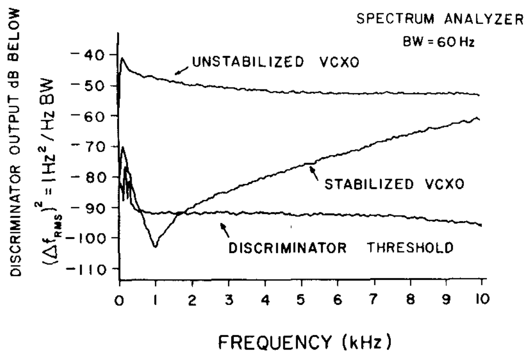


Fig. 4. Discriminator output showing FM noise spectrum up to 10 kHz.

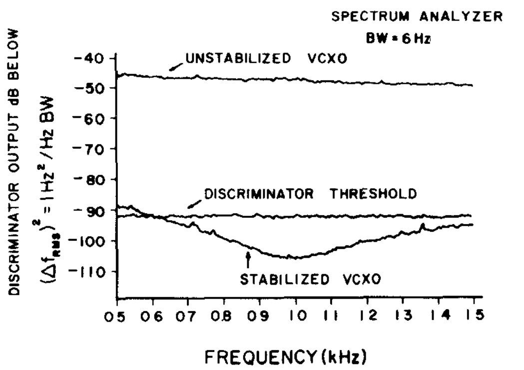


Fig. 5. Discriminator output showing FM noise spectrum centered on 1 kHz.

unstabilized FM noise of the VCXO: 0.004 $Hz/Hz^{1/2}$, at 1 kHz. The measured discriminator threshold corresponds to $2.5 \times 10^{-5}$ $Hz/Hz^{1/2}$ in good agreement with the estimate from (2), and is essentially flat up to 10 kHz. This was measured by replacing the resonator and circulator with an equivalent amount of attenuation (25 dB) and, after re-establishing carrier suppression, temporarily modulating the p-i-n attenuator to adjust the reference (LO) phase of the detector for minimum AM response. This threshold is exactly that due to the RF amplifier looking at a room temperature 50-Ω termination, confirming that the varactor phase shifter does not add any detectable near-carrier FM noise into the signal channel. Separate tests on the electronic phase shifter and attenuator indicate that only the AM noise of the attenuator is detectable, about 9 dB above threshold when the reference phase of the phase detector is adjusted for maximum AM response.

With the servo locked at maximum gain (stabilized VCXO), the FM noise is seen in Fig. 4 to be reduced at all frequencies, closely replicating the inverse of the gain curve of Fig. 3, with a dramatic notch at 1 kHz. Fig. 5 shows the FM noise spectrum expanded around the 1-kHz notch, which reveals a maximum servo gain at 1.000 kHz of about 60 dB, as expected. Over the frequency interval 0.5–1.5 kHz, the stabilization is at least 40 dB. Up to 150 Hz either side of 1 kHz, the observed noise is 10 dB below the discriminator threshold, which guarantees the oscillator FM noise is limited by the discriminator, and further increasing the loop gain cannot improve the stabilization over this limited frequency range.

## V. Conclusion

The oscillator circuit described exhibits very low FM noise, at least over a small range of audio frequency offset from the carrier. The feasibility of an active carrier suppression circuit to significantly lower the discriminator noise floor has been demonstrated. Significant improvements could be made to the circuit by redesigning the dielectric resonator for both an unloaded $Q$ near 20000 and a smaller frequency-temperature coefficient. This would lower the discriminator threshold by some 6 dB and broaden the temperature tolerance of the carrier suppression loop.

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# 140-GHz Finline Components

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*Abstract*—**A balanced mixer and a p-i-n diode switch based on finline technique for the frequency range around 140 GHz are described. The mixer exhibits a conversion loss of 7 to 8 dB, whereas the p-i-n diode switch shows an insertion loss of 2 dB and a maximum attenuation of 33 dB.**

## I. Introduction

Balanced mixers and p-i-n diode attenuators are parts of nearly all millimeter-wave systems, especially radar equipment. Considerable work has been done to realize these components with planar integrated techniques, especially in finline, e.g., [1]–[7]. These investigations, during the last years, were mostly concentrated to the frequency range below 100 GHz. Recently, the highest frequency for finline circuits has been pushed to over 200 GHz for detector circuits [8].

This paper describes our latest developments concerning finline balanced mixers and p-i-n diode attenuators in the 140-GHz range. Together with other components like Gunn and IMPATT oscillators [9], [10] and circulators [11], the basis for a 140-GHz radar front end is given.

The implementation of finline technique for the mixer and the p-i-n diode switch gives the chance to build high-performance, rugged, but relatively low-cost millimeter-wave circuits. The circuit elements requiring high dimensional tolerances are situated on the planar substrate where high precision is no severe problem. The waveguide mount necessary for finline is much less complicated and requires an order of magnitude lower tolerance compared to standard waveguide circuits.

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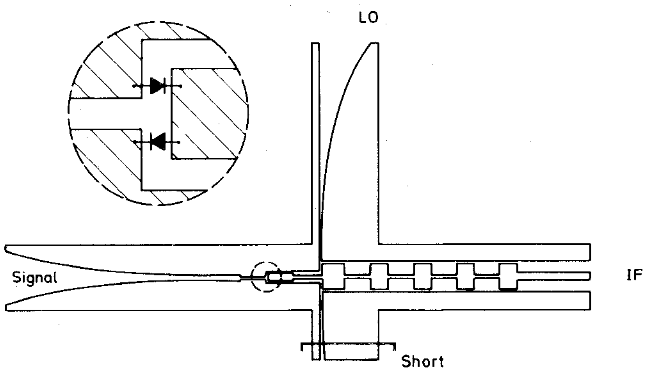


Fig. 1. Basic layout of a balanced finline mixer.

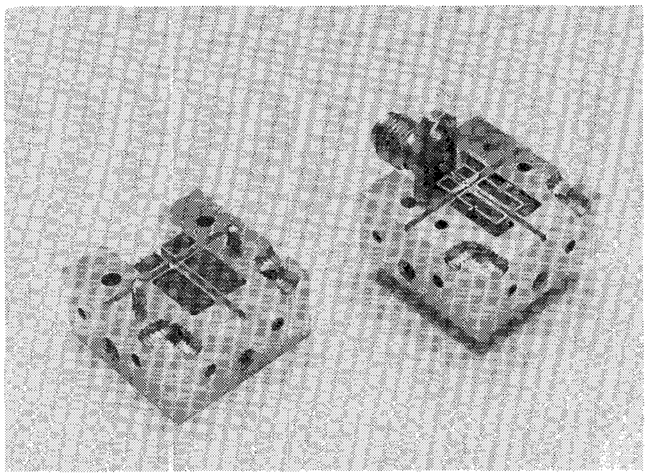

Fig. 2. Photograph of the 140-GHz balanced mixer mount with the planar substrate.

## II. Balanced Mixer

This mixer is built up in a similar way to those at lower frequencies [4]. The planar mixer circuit is etched from a $5\times10$-mm substrate and placed in a waveguide mount. As substrate material, fused quartz of 0.11-mm thickness was used in a first setup [12]. For easier handling, however, RT-Duroid substrates of 0.127-mm thickness were used in the following development steps.

Fig. 1 shows a sketch of the planar mixer circuit. The signal is fed to the diodes via a symmetrical finline. The local oscillator power reaches the diodes via an unsymmetrical finline and a transition from finline to coplanar line. This transition works in a similar way to that of a probe-type transition from waveguide to coaxial line. For proper operation, a short is placed opposite the transition, allowing some tuning of the LO input impedance. An IF microstrip lowpass filter is connected to the coplanar line. A reduced waveguide cross section in the first part of the coplanar line acts as a mode filter; only the coplanar mode is able to propagate to the diodes. On the other hand, this narrow line section provides a reactive termination for the RF signal behind the diodes.

The combination of finline and coplanar line forms a frequency-independent $180^\circ$ hybrid, giving the basis for this balanced mixer. Commercially available beam-lead diodes, in this case the DC 1346 from Marconi, are placed at this junction (Fig. 1). For this frequency band, some problems arise with the size of the diodes. The chip size of the diodes is about $0.25\times0.25$ mm, and their total length including leads amounts to 0.75 mm. In this mixer, two of these diodes have to be placed side by side on a substrate of 0.83-mm width. Furthermore, the Schottky contact of the diode is not situated in the center of the semiconductor chip, leading to an additional longitudinal shift of one diode in relation to the other one. This fact shows the necessity for smaller diodes, especially if even higher frequencies are concerned.

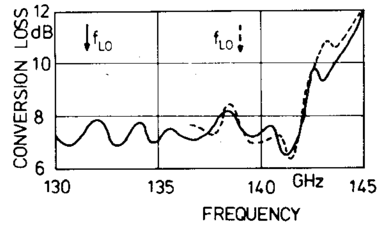


Fig. 3. Conversion loss of the 140-GHz mixer: —— LO frequency 131.5 GHz, ---- LO frequency 139.0 GHz.

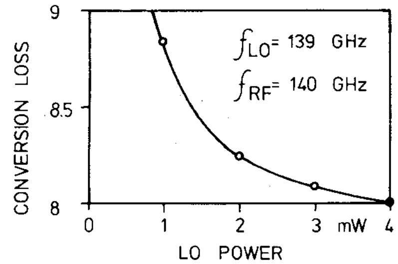


Fig. 4. Conversion loss versus LO power.

An approximate analysis of the diode impedances including the parasitics reveals a relatively low real part (caused by an impedance transformation by the parallel capacitance and series inductance) and a considerable imaginary part of the diode impedances. To compensate for this, impedance matching sections were included in the LO and signal paths. Fig. 2 shows a photograph of the open mixer mount and the quartz substrate.

For self-bias operation of the mixer, an LO power of about 10 to 15 mW is required. The generation of such a power at 140 GHz, however, provides difficulties and, for this purpose, is not acceptable. So, the mixer diodes are biased by applying a voltage between the two fins of the signal input. To separate the IF from the bias circuit, a capacitor is used in the IF output. In this way, an LO power of only 2 to 5 mW is required for this mixer. Nevertheless, a first measurement was made with an IMPATT oscillator as LO, delivering 12 mW at 131.5 GHz, sufficient for self-bias operation. The conversion loss versus frequency for this case is given in Fig. 3 as a solid line. Between 130 GHz (and possibly even below) and 142 GHz, the conversion loss lies around 7.5 dB; above 142 GHz, however, it increases rapidly due to an increasing RF mismatch. In a next step, an IMPATT local oscillator at 139 GHz with a power of 4 mW was used, applying an additional bias of 2 mA to the mixer diodes. The conversion loss of this arrangement is also included in Fig. 3 (broken line). As can be seen, a rather close agreement between the two results is found, indicating that both operation modes lead to an equivalent performance. Furthermore, no deterioration of the conversion loss at high IF frequencies compared to low values can be found. The dependence of the conversion loss on LO power was investigated with the second configuration at 140.5 GHz, using again an LO frequency of 139 GHz (Fig. 4). Below 2 mW of LO power, the conversion loss increases rapidly, whereas above 4 mW, an improvement of only a few tenths of a decibel can be achieved.

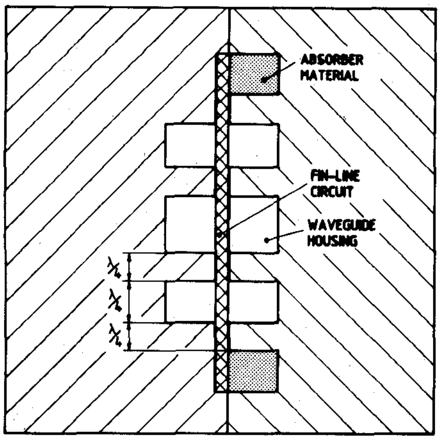


Fig. 5. Cross-sectional view of the finline p-i-n diode attenuator.

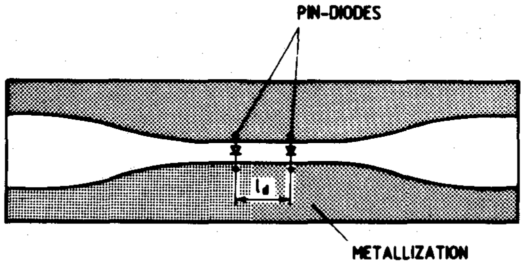


Fig. 6. Sketch of the planar circuit of the p-i-n diode attenuator.

The RF return loss is better than 10 to 15 dB between less than 130 and 141.5 GHz. The LO return loss amounts to about 15 dB within a bandwidth of approximately 0.5 GHz; its center frequency, however, can be tuned with the backshort (Fig. 1) over a bandwidth of more than 15 GHz. The isolation between LO and RF port is about 25 dB, in spite of the problems with the (mechanical) diode symmetry.

## III. p-i-n Diode Attenuator

p-i-n diode attenuators and switches are part of nearly all radar and communication systems. They are used, e.g., as STC's (sensitivity time control) in the receiver front ends of radar equipment, as TR-switches, as ASK-modulators or in conjunction with a circulator, as PSK-modulators in communication systems.

Fig. 5 shows a cross-sectional view of the switch. The waveguide mount is composed of two counterparts containing the finline substrate in its center part.

A sketch of the planar circuit board is given in Fig. 6. S-shaped tapers form the transitions to standard rectangular waveguide. Two beam-lead p-i-n diodes with a spacing $l_d$ are soldered across the slot. To enable bias supply, the fins are dc-isolated by anodizing the aluminium housing.

In the ON state of the switch, the p-i-n diodes are unbiased; the fin line is loaded with the diodes parasitic capacitances. In the OFF state, bias is applied to the diodes, and as the RF resistance of the p-i-n diodes decreases with the bias current, the finline is nearly short-circuited. The isolation is limited because of the series resistance of the diodes and the inductance of the leads. Design criteria are to reach low transmission loss in the ON state and high isolation in the OFF state. The ON state is critical and determines the design. An equivalent circuit is given in Fig. 7 for this case. The parallel impedance of the parasitic capacitance ($\approx -j66\Omega$) of one p-i-n diode has approximately half the value

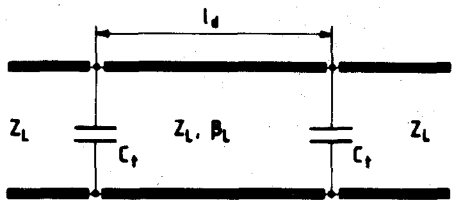


Fig. 7. Equivalent circuit for the p-i-n diode attenuator.

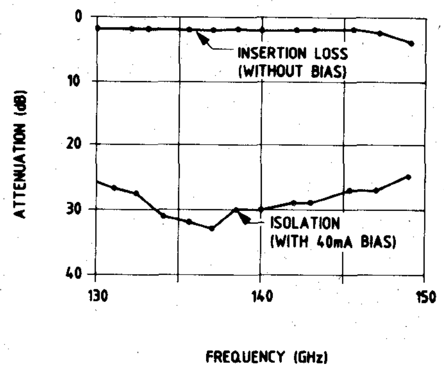


Fig. 8. Transmission characteristics of the 140-GHz p-i-n diode attenuator.

of the lowest waveguide impedance which can be realized in this configuration. For impedance matching, the spacing between the two diodes is given by

$$l_\alpha = \frac{\lambda}{2\pi}\cos^{-1}\left|\frac{Z-1}{Z+1}\right| \qquad Z = \frac{1}{1+jwC_D\cdot Z_L}. \tag{1}$$

To have maximum bandwidth, the finline with minimum impedance has to be chosen, that is the finline with the smallest slot enabling the implementation of the diodes. Also, the OFF-state performance improves with reduced slot width, because the distance between the p-i-n diodes calculated from (1) stays close to a quarter wavelength, which is the optimum distance for isolation. Earlier experiments at lower frequencies revealed that higher mode coupling decreases the isolation if the diodes are placed too close together. To prevent higher order mode coupling, one-half wavelength was added to the distance calculated from (1). The waveguide impedance and propagation constant are calculated from [13].

The switch described here is built with commercially available beam-lead p-i-n diodes with $C = 17$ fF, $R_s = 5\Omega$, and 200 ns minority carrier life time. As substrate material, fused quartz of 0.11 mm thickness was used; it should not provide severe difficulties, however, to use RT-Duroid material for the switch, too.

The insertion loss and the isolation of this p-i-n diode switch are plotted as a function of frequency in Fig. 8. The loss is 2 dB over a bandwidth of 15 GHz between 130 and 145 GHz. Insertion-loss measurements of the finline without diodes revealed 1–1.2-dB loss due to the line, so the loss due to the diodes is only 0.8–1.0 dB. Some improvement seems to be possible reducing the length of the complete planar structure to a minimum. Maximum isolation of $29 \pm 4$ dB is achieved with a bias current of 40 mA applied to the p-i-n diodes. Experiments at lower frequencies have proven that, if more isolation is required, an additional p-i-n diode can be added to obtain higher isolation with only a small increase of loss. In this case, the distances between the diodes have to be modified slightly for optimum performance.

The switching speed between 50-percent logic to 90-percent detected RF signal is divided into 5-ns delay time and 30-ns rise time. The time from 50-percent logic to 10-percent detected RF signal is divided into 40-ns delay time and 15-ns fall time. The maximum power handling capability was not measured, but it can be assumed that 26–30-dBm power handling is possible due to the diodes total power dissipation of 24 dBm.

## IV. Conclusion

It has been shown that the design of finline mixers and p-i-n diode attenuators is possible at frequencies around 140 GHz with acceptable results. Besides the electrical data, excellent mechanical stability is achieved with this approach. For example, a finline balanced mixer on a quartz substrate with even larger dimensions (*E*-band) than in this case has been successfully exposed to shock accelerations up to 30 000 g, a value unthinkable for whisker-contacted mixers.

Furthermore, production is made easier and cheaper by the planar circuit in conjunction with beam-lead diodes. Last, but not least, integration techniques have become possible, including *E*-plane circulators, leading to very compact front ends with losses lower than the sum of the discrete components as demonstrated at 94 GHz [4], [6].

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# Radial-Line/Coaxial-Line Stepped Junction

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*Abstract* — **A radial-line/coaxial-line junction having a step in the inner conductor in the coaxial aperture is considered. It is shown that the stepped junction may be modeled by an equivalent circuit obtained by a simple modification of that for the unstepped junction. Comparison of theoretical and experimental results has established this representation to be very accurate over a wide frequency range.**

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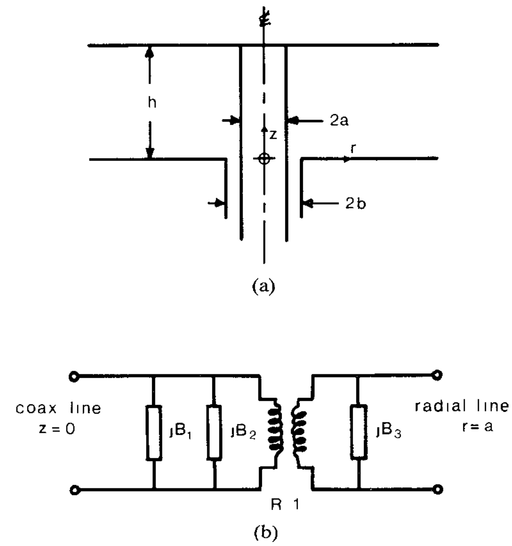


Fig. 1. Radial-line/coaxial-line junction: (a) cross-sectional view, (b) equivalent circuit.

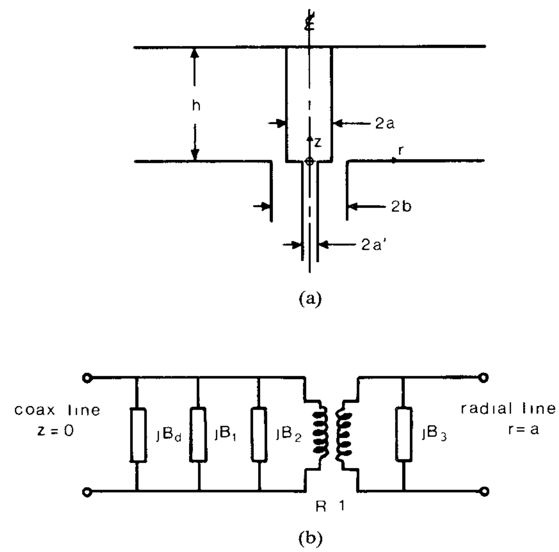


Fig. 2. Radial-line/coaxial-line stepped junction: (a) cross-sectional view, (b) equivalent circuit.

## I. Introduction

There are many microwave devices which employ, in some form or other, a radial-line/coaxial-line junction (e.g., antenna feeds, power dividers, and combiners). In designing such devices, it is advantageous to have an equivalent circuit representation for the junction. However, in spite of the wide application of radial-line/coaxial-line junctions, there have been relatively few studies reported. Indeed, it is only recently that equivalent circuits have been developed for the junction [1]–[3].

Recently, Allison [2], [3] investigated the design of wide-band transitions. He concluded that for the coaxial line he was employing, transitions of the form shown in Fig. 1(a) were unsuitable for wide-band performance, and that it was necessary to enter the