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Coplanar Waveguide Transitions to Slotline: Design and Microprobe Characterization

Wes Grammer and K. Sigfrid Yngvesson

***Abstract*—Three types of coplanar waveguide/slotline baluns, suitable for monolithic integrated circuits, are designed and characterized to 40 GHz, using microwave wafer probes. The results from each are compared, and methods of biasing the slotline discussed.**

I. Introduction

The use of coplanar waveguide has several distinct advantages over microstrip as a medium for monolithic microwave integrated circuits. Its uniplanar structure eliminates the need for ground vias, which introduce an undesirable parasitic inductance and limit the performance at high frequencies. Coplanar lines can also be scaled to smaller widths for a given impedance, without the need for substrate thinning. Lastly, integration of devices in series or shunt is simple. Integration with other uniplanar structures, such as tapered-slotline antennas [1] or slot rings [2]–[3], requires a suitable circuit transition from coplanar waveguide (CPW) to slotline.

Three different types of CPW–slotline transitions were identified in the literature. These are shown in Figs. 1–3, along with their approximate equivalent circuit models. The first, referred to here as Type I, is based on a CPW-slotline T-junction [4], [5]. A variation on this type uses a grounded substrate and finite-width strips for the slotline, exciting a coupled microstrip-like quasi-TEM mode [6]. The second (Type II) is a uniplanar realization of the compensated Marchand balun [7]–[9]. The third (Type III) is a CPW equivalent of the double-junction microstrip-slotline balun [10]. The elements Z_{1-6} in Fig. 3 refer to the input impedances of the stubs and transmission lines, as seen from the junction. All of these designs are uniplanar, but require one or more bond wires or air bridges at the junction for odd-mode suppression in the CPW.

The aim of this work was to investigate how the earlier designs could be translated to MMIC-type circuits (on high-resistivity silicon substrates), and toward higher frequencies, up to 40 GHz.

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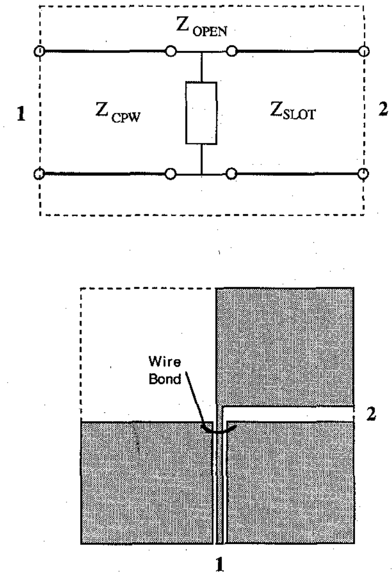

Fig. 1. Type I balun and equivalent circuit.

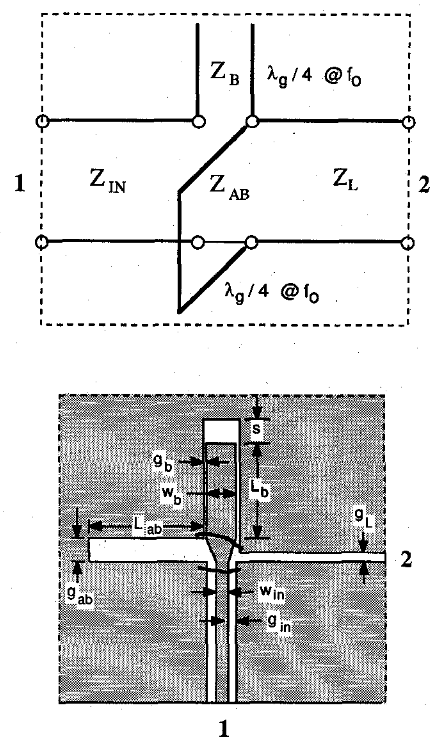

Fig. 2. Type II balun and equivalent circuit.

II. Analysis and Design

A. Type I Balun

Assuming the unterminated side of the T-junction to be a decent broadband open, the conductor gap and width dimensions of the transmission lines in Fig. 1 are chosen to give an approximate 50 Ω match. The CPW dimensions are made small to minimize parasitic effects at the junction. Using the transmission line analysis program PCAAMT [11], for a silicon substrate ($\epsilon_r = 11.9$) with a thickness $h = 0.33$ mm, a width and gap size of 25 μm was found to be

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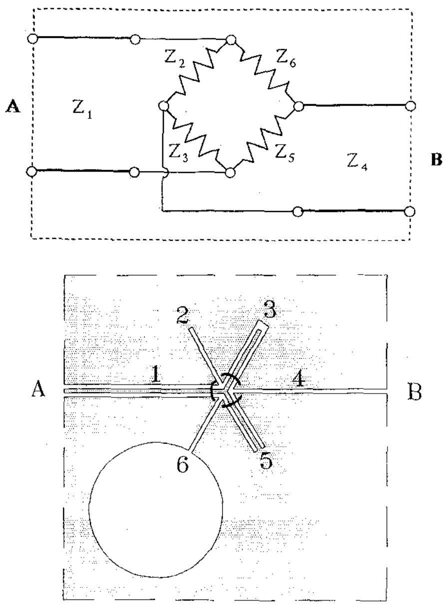

Fig. 3. Type III balun and equivalent circuit.

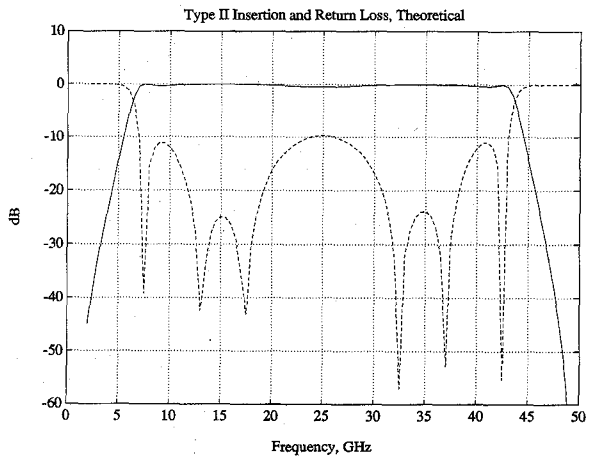

Fig. 4. Modeled response of back-to-back type II baluns.

appropriate. The width of the slotline fins is made sufficiently wide ($\lambda_g/2$ at $f_o = 25$ GHz) to reduce fringing effects where the guide wavelength λ_g is [12]

$$\lambda_g/\lambda_o \approx \sqrt{2/(\epsilon_r + 1)}\,. \tag{5}$$

A single air bridge connecting the grounds is required at the junction. Additional pairs are needed at all right-angle bends.

B. Type II Balun

The compensated balun is designed with a 4:1 bandwidth, or two octaves. As shown in Fig. 2, two relatively long air bridges are required, spanning the center conductor of the CPW input line and open stub. For this design, the lower cutoff frequency $f_{x1} = 10$ GHz, and the upper cutoff frequency $f_{x2} = 40$ GHz. The center frequency is then $f_o = (f_{x1} + f_{x2})/2 = 25$ GHz. The next step is to determine the optimum characteristic impedances of the lines and stubs. From [8], making the impedance Z_{ab} (and hence the slot width g_{ab}) as large as possible maximizes the bandwidth. However, this dimension must be small relative to λ_g at the upper band edge f_{x2}. From (1), $\lambda_g \simeq 2.95$ mm at 40 GHz, assuming the same silicon

TABLE I
DESIGN VALUES, TYPE II AND TYPE III BALUNS

Param.	Value	Units	Param.	Value	Units
ϵ_{ab}	4.57[a]	—	l_6	390[c]	μm
ϵ_b	6.25[a]	—	s	200	μm
g_{ab}	205	μm	W_b	200	μm
g_b	25	μm	$W_{\rm in}$	100	μm
$g_{\rm in}$	70	μm	Z_{ab}	100	Ω
g_L	75	μm	Z_b	30.9	Ω
L_{ab}	1400	μm	$Z_{\rm in}$	53	Ω
L_b	1200	μm	Z_L	70	Ω
l_2	390[c]	μm			
l_3	415[b]	μm	$Z_{\rm CPW}$	50	Ω
l_5	415[b]	μm	$Z_{\rm SLOT}$	50	Ω

[a]At $f_o = 25$ GHz; $\epsilon_r = 11.9$.
[b]Measured from center cond. bend to discontinuity.
[c]Measured from discontinuity to bend.

substrate. Using this as a guide, a realistic choice works out to be $Z_{ab,\max} \simeq 100\ \Omega$, with $g_{ab} = 205\ \mu$m. Next, a slotline impedance $Z_L = 70\ \Omega$ is arbitrarily selected; note that with this balun, Z_L and $Z_{\rm in}$ do not have to be equal. Now the optimum value for Z_b can be computed from [8] as

$$Z_{b,\rm opt} = \left[\frac{Z_{ab}\tan^2\left(\frac{\pi}{2}f_x/f_o\right)}{1+\left(\frac{Z_{ab}}{Z_L}\right)^2\tan^2\left(\frac{\pi}{2}f_x/f_o\right)}\right] = 25.4\ \Omega \tag{6}$$

where f_x is either f_{x1} or f_{x2}.

The necessary physical dimensions for each of the above transmission lines are synthesized using PCAAMT, and are summarized in Table I. Given the constraints on realizable dimensions, a 4:1 bandwidth is about the maximum achievable in our case. Also shown are the computed $\epsilon_{\rm eff}$ of the lines and quarterwave stub lengths of the preliminary design. The stub-end and junction reactances are neglected in the initial design. For the open-circuited coplanar stub, an end gap equal to the center conductor width is arbitrarily chosen. All calculations assume a frequency f_o, and the substrate parameters given previously. The value of Z_b was compromised slightly (30.9 Ω), due to the dimensional limitations of the line.

A simulation of the balun equivalent circuit was performed, using the values in the table. Fig. 4 shows the theoretical insertion and return loss responses of a back-to-back (CPW-slot-CPW) configuration. The lines were assumed ideal (no attenuation or dispersion); also, the reactances of the stub ends, air bridges, and junction were neglected. The passband is over two octaves wide and very flat, and is periodic on $2f_o$. By shortening or lengthening the stubs, the center frequency may be shifted up or down. The simulation revealed that the transmission line impedances were not extremely critical; varying them over $\pm 20\%$ (well beyond the tolerances from fabrication) had only a slight (≤ 1 dB) effect on the response.

C. Type III Balun

This design has the advantage of an inline port geometry, convenient in monolithic array applications and probing. However, this design requires three tightly-spaced air bridges at the junction, across the grounds of the CPW stubs and feed. These are again to suppress the CPW odd mode.

The equivalent circuit in Fig. 3 of the Type III balun can be analyzed with elementary circuit theory [13]. Thus, the normalized input impedance at port 1 of the junction can be shown as

$$Z_{\rm in} = \frac{1}{Z_2' + Z_3' + 1}\left[Z_2'Z_3' + \frac{(Z_2'+Z_3')(Z_2'+1)(Z_3'+1)}{(Z_2'+Z_3'+2)}\right] \tag{7}$$

assuming $Z_2 = Z_5$ and $Z_3 = Z_6$, and that all impedances are normalized to Z_4, i.e.,

$$Z_n = Z_n' Z_4 . \tag{8}$$

For a perfect match ($Z_{\rm in} = 1$), it can be shown that (3) reduces to the condition

$$Z_2' Z_3' = 1 \tag{9}$$

or, in terms of reflection coefficients,

$$\Gamma_2 = -\Gamma_3 . \tag{10}$$

A significant fact revealed in the analysis is that the stub impedances are specified relative to one another, and not as fixed quantities. In theory then, if all stubs are the same electrical length, then a perfect match is maintained regardless of frequency. In reality, the bandwidth is limited by imperfect stub terminations, the unequal dispersion between slotline and CPW, junction parasitics, and mode conversion. Additionally, all stubs and transmission lines must have the same characteristic impedance, to maintain the match over a broad band. So, unlike the Type II design, this balun cannot transform impedances. To minimize loss in the balun, the stub terminations are made purely reactive (shorts and opens).

For the physical circuit, the line impedance of both CPW and slotline is chosen to be approximately 50 Ω, to match the wafer probes. The appropriate dimensions are again synthesized with PCAAMT; all relevant design information is summarized in Table I. The stub lengths are kept short ($\lambda_g/8$) to minimize attenuation and the effects of dispersion. The open- and shorted-end reactance of both the CPW and slotline stubs are assumed ideal; for the open-circuited coplanar stub, an end gap of twice the center conductor width is chosen arbitrarily. For the slotline open, the same kind of circular resonator is used as in [10], with the proper diameter determined empirically. This type of open circuit has been shown to have good bandwidth, though radial stub resonators have also been used [14]. All design calculations assume a center frequency $f_o = 35$ GHz and the silicon substrate parameters previously given.

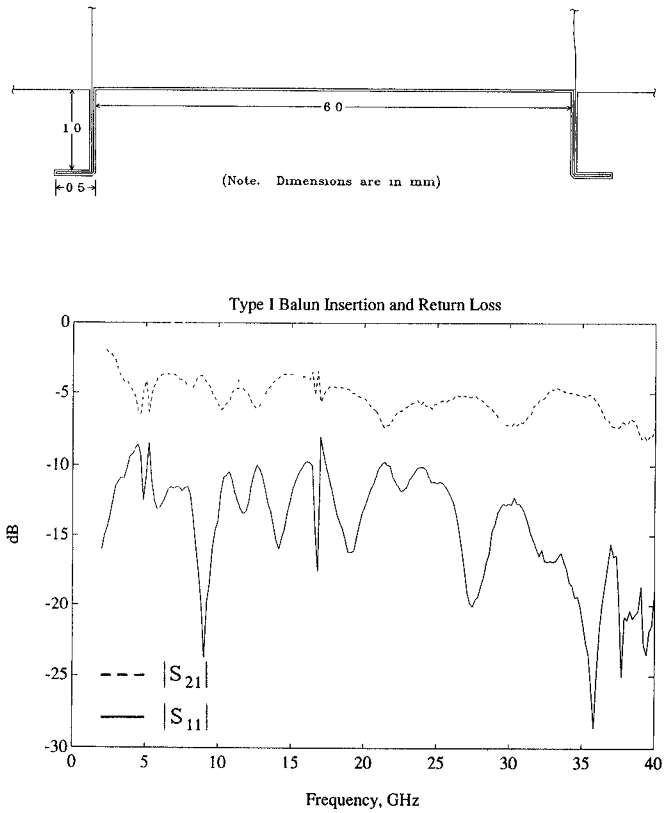

Fig. 5. Type I balun circuit layout and measured response.

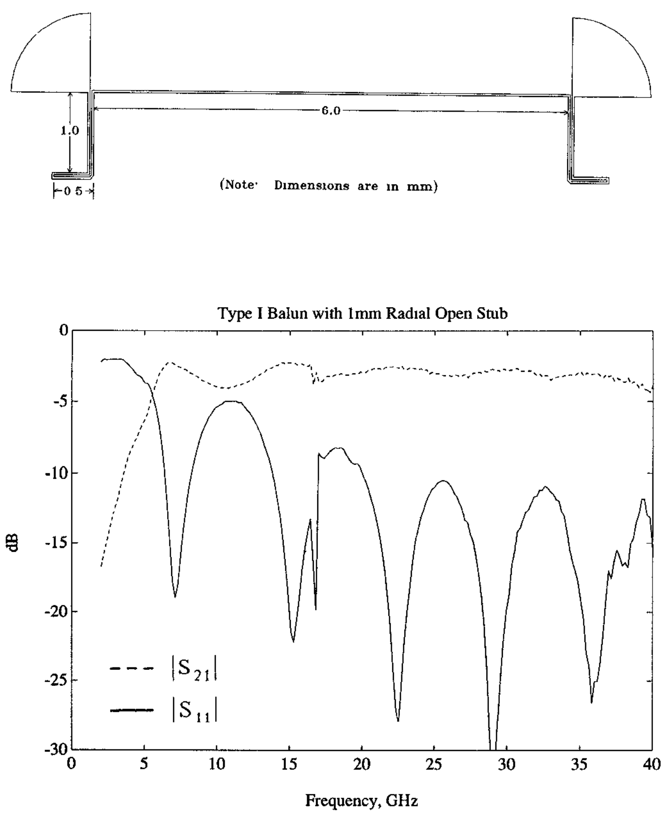

Fig. 6. Type I balun with radial open, measured response.

III. Circuit Fabrication and Measurement

The test circuits are fabricated on high-resistivity silicon, with thickness $h = 0.33$ mm. A thin (∼200 Å) adhesion layer of TiW is deposited on the surface, followed by a 0.25 μm layer of copper. The wafer is then processed like any other printed circuit, using standard photolithographic procedures. A thin layer of gold is deposited onto the copper, to inhibit oxidation and assure good repeatable probe contact. Finally, the necessary wire bonds are attached at the balun junctions and right-angle bends. It was later discovered during the measurements that extra bonds were needed at quarter-wavelength intervals on the CPW lines to completely suppress the odd mode.

The circuits are measured using a vector network analyzer and a wafer probe measurement system, with 50 μm-pitch 40 GHz coplanar waveguide probes. The system is calibrated at the probe tips from 2–40 GHz, using LRM (line-reflect-match), a calibration technique similar to TRL that uses precision 50 Ω planar terminations in place of the LINE standard, to obtain a fixed reference impedance over a broad band. The circuit substrate is measured raised off the surface of the wafer stage with a 6 mm-thick spacer ring, to decouple the TM_0 parallel-plate mode.

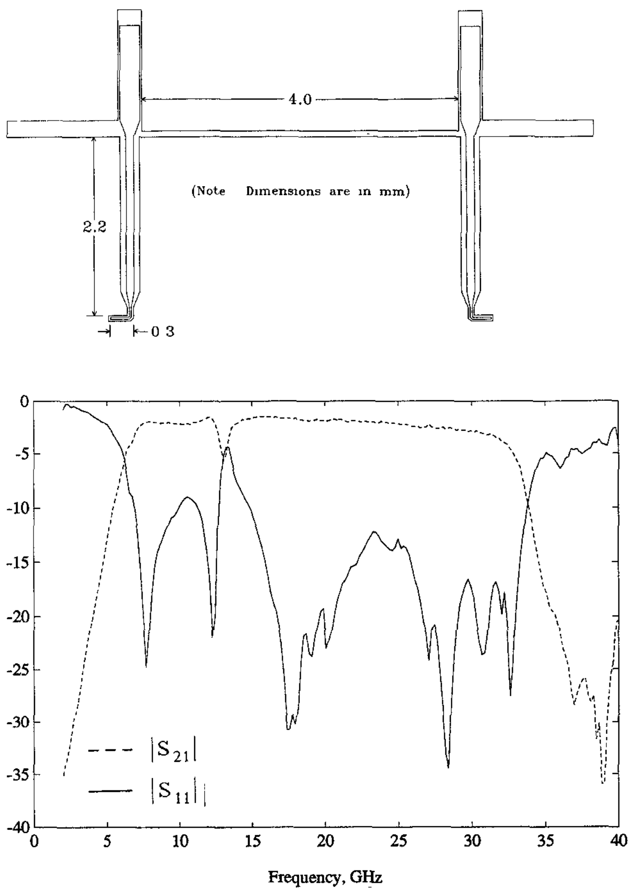

Fig. 7. Type II balun initial design layout, measured response.

IV. Measured Results and Discussion

A. Type I Balun

The initial Type I balun design was fabricated and measured in the symmetric (back-to-back) configuration, shown in Fig. 5 along with the measured insertion and return loss data. The passband of the circuit extends from DC to 40 GHz; however, there are many small dips and ripples in the thru response, and overall loss is relatively high. This is probably due to radiative loss in the unterminated side of each junction.

In an effort to reduce the loss, the transition was modified as shown in Fig. 6. The unterminated side of the junction is replaced with a 90-degree radial open stub, of radius $R = 1$ mm. A radial slotline stub is in an approximate sense the dual of a microstrip radial stub, which is a good broadband RF short, but open at DC. Both structures have been used in the design of broadband microstrip-slotline transitions [14]. The measured response of the modified design is also given in Fig. 6. The insertion loss is much improved—the insertion loss response is smooth and flat, with only a trace of ripple and a shallow dip near the lower band edge (~6 GHz). Further analysis of the measured attenuation revealed that it is due almost entirely to conductor loss, not surprising considering the 0.25 μm metal thickness. The overall return loss is higher; however, the response has the characteristic of a minor line impedance mismatch at each junction. An adjustment to either the coplanar line or slotline gap dimensions could further improve the match. Despite the loss of a passband at DC, this balun still has nearly three octaves of measured bandwidth.

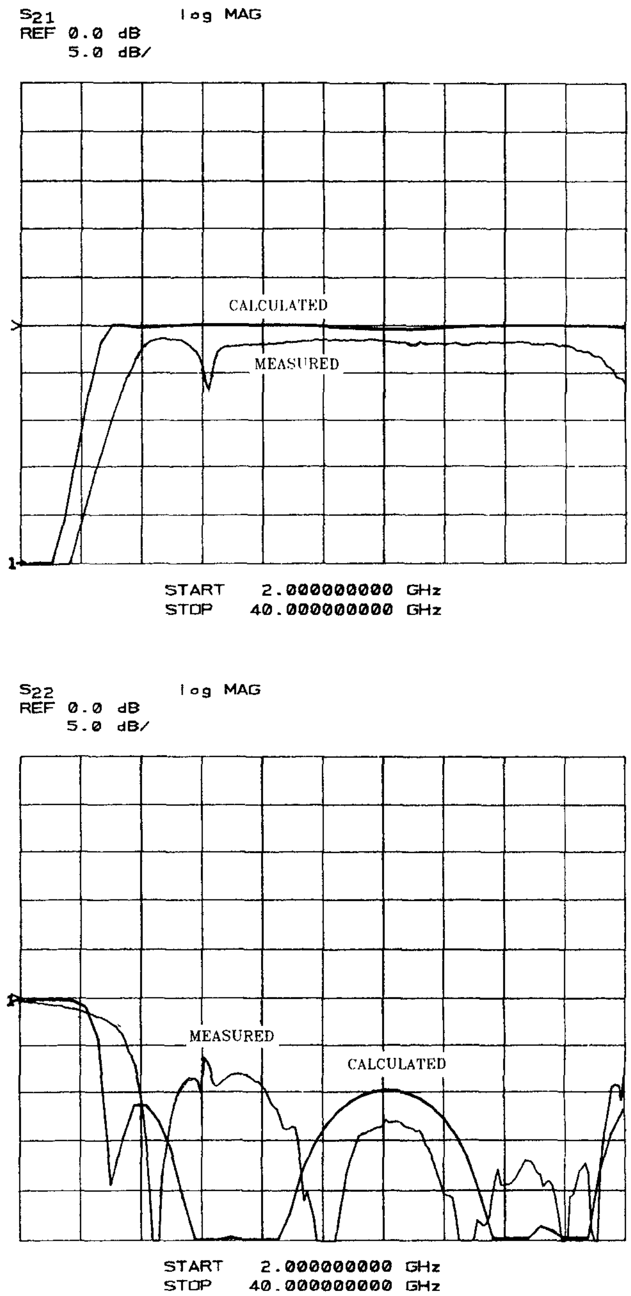

Fig. 8. Modeled versus measured response, type II balun (modified).

B. Type II Balun

The measured insertion and return losses for the initial Type II design are shown in Fig. 7. The passband is very broad and flat, with the lower band edge at 6.3 GHz and the upper edge near 32.8 GHz. Clearly, the measured response exhibits a lower center frequency (f_o = 19.55 GHz) than what was designed for; this is probably due to the fringing reactances of the open and shorted ends of the CPW and slotline stubs, which were neglected in the initial design. Using the circuit model described earlier, these effects can be modeled empirically from the measured data, and the design modified to readjust f_o to the desired value. From analysis, it was determined that shorter stubs were required: the open stub length was decreased 0.4 mm to 1.2 mm, and the shorted stub by 0.3 mm, to 1.4 mm. Additional baluns were then fabricated using the revised dimensions and measured. From these results, the optimum response was obtained

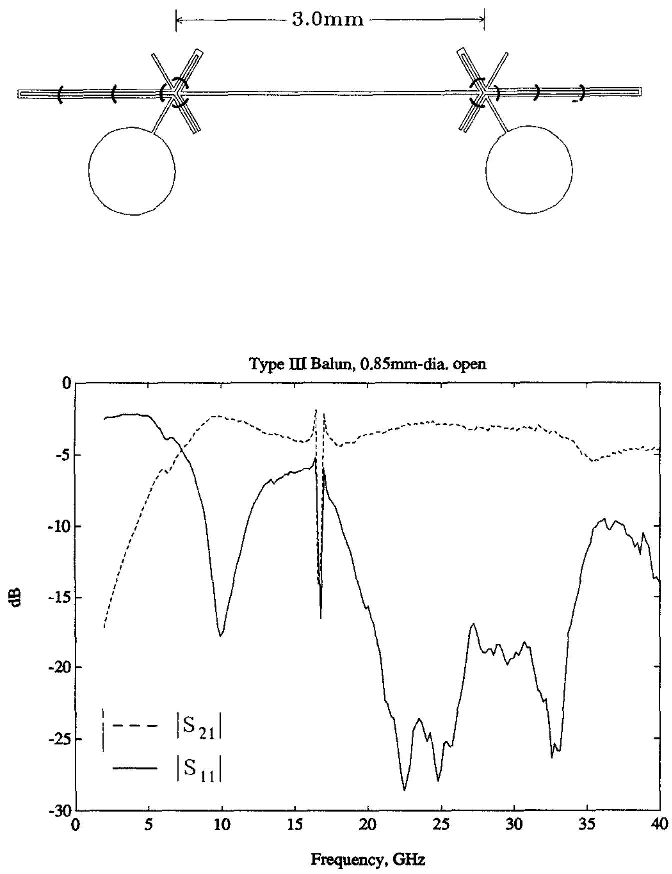

Fig. 9. Type III balun initial design with mode suppression.

by shortening the open stub an additional 0.1 mm, to 1.1 mm. Fig. 8 shows the measured response of the revised design, compared with the theoretical response from the circuit model. The actual circuit has a slightly narrower bandwidth than the model; this is probably due to dispersive effects not included in the model (the value of ϵ_{eff} for both CPW and slotline increases with frequency, so at high frequencies the lines and stubs appear electrically longer than if ϵ_{eff} remained constant. Therefore, the upper band edge would be shifted downward as a result. The opposite effect occurs at the lower band edge, which shifts up slightly). The asymmetry in the response may be due to an increase in parasitic effects from the junction at higher frequencies. The spike in the passband is due to a defective cable, and is a relic of the calibration, not part of the actual response.

C. Type III Balun

The back-to-back circuit layout and measured response of the Type III design is given in Fig. 9. From tests of several shapes and sizes, a 0.85 mm-diameter circular slotline open was selected as giving the optimum passband response and return loss. The insertion loss is fairly flat across the band, except for several shallow dips at 16 and 35 GHz (the spike is from the bad measurement cable). The return loss is greater than 15 dB from 20 to 35 GHz. Overall loss is about the same as with the Type I design.

V. Biasing Methods/Measurements

In many cases, there is a need for DC isolation between the slotline fins, to permit biasing of shunt-mounted devices. This is a requirement in some mixers, where a bias current is necessary for the proper operation of the device [1]. Except for the initial Type I design, none

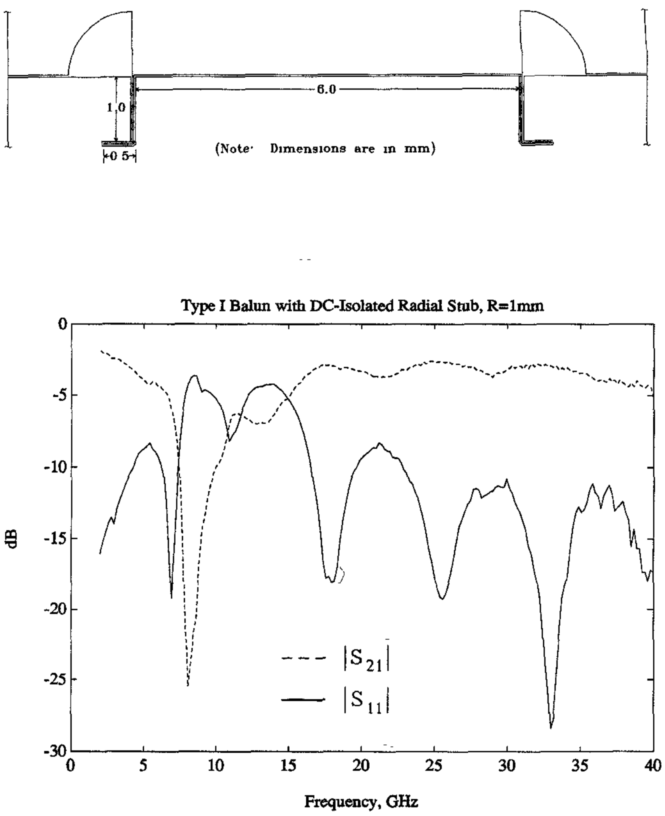

Fig. 10. Type I balun with DC-isolated radial open, measured response.

of the above transitions has a DC passband. In the following sections, various means will be discussed to extend operation of the baluns to DC.

A. Type I Balun

In an attempt to regain DC isolation and retain the good high-frequency response of the terminated design, the configuration in Fig. 10 was developed. The radial arc of the open stub is split at one end by a narrow (25 μm) slit of length 1 mm, out to the circuit metal boundary. The insertion and return loss responses of this circuit are shown in the same figure. Other than a deep dip at 8 GHz and a shallower dip near 13.5 GHz, balun performance is quite satisfactory at or below 2 GHz and above 18 GHz. Compared to the response in Fig. 6, this circuit has virtually the same insertion loss (3–4 dB) from 18 to 40 GHz.

B. Type II Balun

In this design, the conductor must be split somewhere in the circuit to isolate the slotline fins, without drastically affecting the loss and bandwidth of the circuit. Since the ground planes on the coplanar stubs must remain connected to suppress the odd mode, the only possible locations for such a slot are at the junction, in series with the slotline stub, or at its shorted end, in a cascade configuration. An equivalent circuit model was used to analyze the various combinations, and determine whether they might work. The result was that neither location could retain the broadband characteristics of the original design and also be a physically realizable circuit.

An alternative to a simple slot is a monolithic capacitive termination of the slotline stub, in place of a short. The capacitance must be large enough to present a low impedance within the desired passband. One possible structure is a set of overlapping parallel plates, separated by a thin dielectric layer. A vertical structure like this is difficult

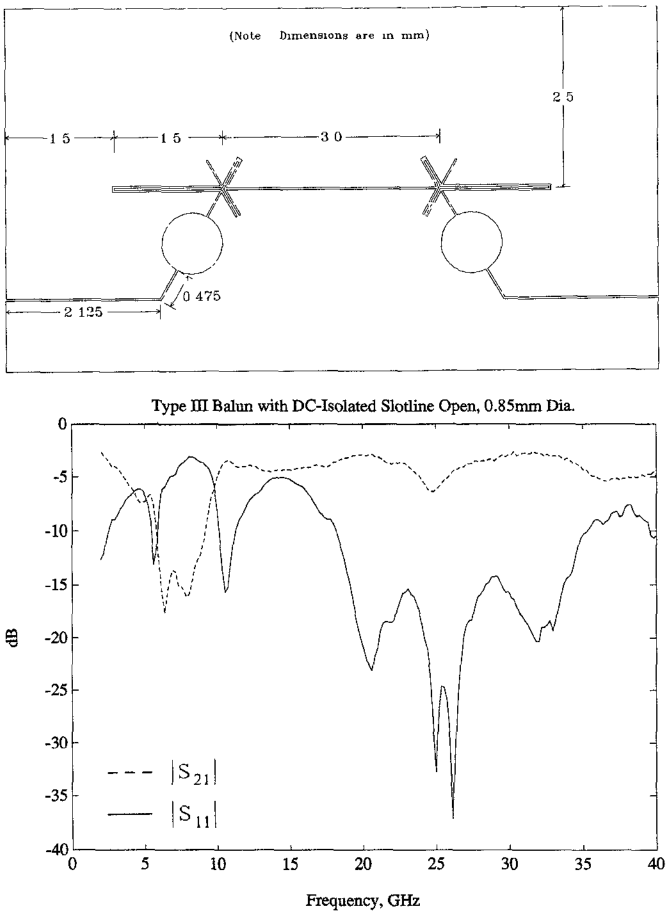

Fig. 11. Type III balun, modified for DC bias, measured response.

to fabricate, due to the number and complexity of processing steps required, and was not attempted.

C. Type III Balun

Provision for DC bias may be introduced in the same way as with the Type I balun, due to their common use of a slotline open stub. A narrow (25 μm) slit is placed directly opposite the line entering the circular open from the junction, and runs to the outer edge of the circuit metal, as shown in Fig. 11. The corresponding insertion and return losses are shown at the bottom of the figure. A low-frequency passband from DC to 2 GHz has been added, with only slight degradation of the original response. Other than a minor dip at 25 GHz, the insertion loss is less than 5 dB from 10 to 40 GHz, and the return loss greater than 10 dB form 18 to 35 GHz.

VI. Conclusions

We have demonstrated in this paper that three different efficient, wideband transitions can be designed between the CPW and slotline circuit media, up to 40 GHz. It is noteworthy that no surface wave problems were encountered up to 40 GHz, although the silicon substrates were 0.33 mm thick. We estimate that the same substrates could be utilized up to about 100 GHz. The information generated here should be useful for uniplanar MMIC design at millimeter wave frequencies, and with thinner substrates could be applied at THz frequencies. Biasing of devices across the slotline is feasible in MMIC versions of all three designs. We have recently employed the biasable Type III balun to characterize a new hot electron mixer device at 300 K as well as 77 K, using a Cascade wafer probe [13].

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