Letters

An Improved Microstrip-to-Microslot Transition

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***Abstract*—An improved broad-band microstrip-to-slot transition has an insertion loss of less than 0.2 dB and is not critical in its dimensions.**

Microstrip transmission lines in combination with slotlines [1] offer an additional degree of freedom in the design of microwave integrated circuitry [2]–[4]. An important component in strip-slot circuitry is the well-matched strip-slot transition with low insertion loss. A standard method to design such a transition is shown in Fig. 1. The slot, terminated in an open circuit, crosses the microstrip in the ground plane. The microstrip, on the other hand, is terminated in a short circuit. A narrow-band version of this type of transition has been analyzed in detail [5].

Experimentally, however, a transition as in Fig. 1 is not completely satisfactory as the insertion loss is fairly high, the VSWR is not particularly low, and the usable bandwidth is not as large as one expects. The reason for the nonoptimum behavior of the transition is certainly due to the fact that, because of mechanical restrictions, the effective plane of the microstrip short circuit is not exactly at the slot plane where it should be. Similarly, the slot open circuit is not exactly beneath the microstrip.

Although different solutions have already been proposed [6], there are also drawbacks.

We shall show that a slight modification of the strip-slot transition of Fig. 1 makes the position of the microstrip short-circuit plane and the slot open-circuit plane far less critical and improves the overall performance of the transition.

In order to explain the necessary modifications, one may start from the double junction of Fig. 2 [7]. This 6-port is matched at port 1 and port 6 if the ports 2–5 are terminated by matched loads, if the characteristic impedances of the strip and slotlines are all the same, e.g., 50 Ω, and if junction effects can be neglected. Port 1 and port 6 are decoupled. Therefore looking from port 1, the slotlines leading to ports 2 and 3 are parallel to one another and in series to the two parallel striplines that lead to ports 4 and 5. As a result of this the input impedance, looking

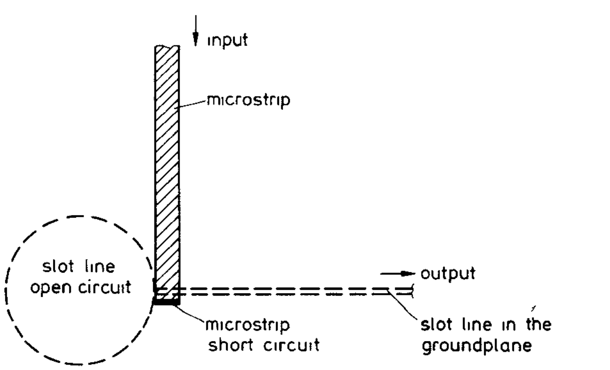

Fig. 1. Standard method for the design of a microstrip-slot transition.

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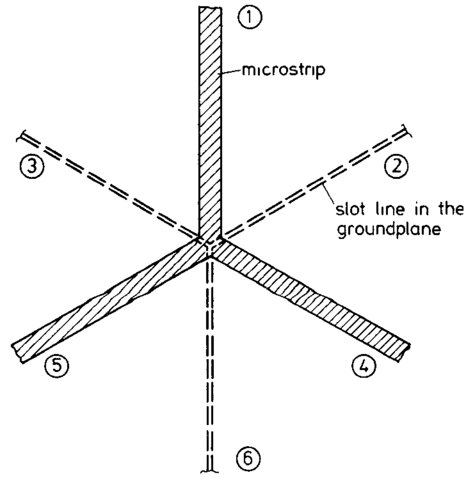

Fig. 2. Strip-slot double junction.

from port 1, is 50 Ω. Similarly, looking from port 6, the slotlines leading to ports 2 and 3 are in series to one another and parallel to the serial striplines that lead to ports 4 and 5. Thus, also looking from port 6, the input impedance into the double junction is 50 Ω. Due to symmetry a similar statement holds for any other pair of opposite ports, i.e., 3 and 4 or 2 and 5. Consider now an input signal at the microstrip port 1. The input signal will be equally transmitted into ports 2–5. If ports 2–5 have the same reflection coefficient, then the signal will be reflected back into port 1. Similarly, an input signal at the slot side (port 6) will also be equally transmitted into ports 2–5, but in phase opposition at ports 2 and 5 with respect to ports 3 and 4. Again, an equal reflection coefficient at ports 2–5 will reflect the signal back into port 6. Therefore a reflection coefficient of $\exp(j\varphi)$ at ports 2 and 5 and of $\exp(j\varphi + 180°)$ at ports 3 and 4 causes a perfect transmission of the input signal (port 1) into the output port 6.

Fig. 3 shows a realization of this transition with short and open circuits. The lengths l_{strip}, l_{slot}, i.e., the distance from the short or open circuits to the center of the structure, are arbitrary, at least in theory, as long as the electrical lengths are equal, i.e., $\beta_{\text{strip}} \cdot l_{\text{strip}} = \beta_{\text{slot}} \cdot l_{\text{slot}} = \theta$. For $\theta = n \cdot (\pi/2)$; $n = 0,1,2,3,\cdots$ the structure is electrically equivalent to the circuit of Fig. 1. In the experimental realization l was made conveniently small (Fig. 4).

It proved for different distances from the short circuit to the open circuit, i.e., $2l$, that $2l$ should be made smaller than a quarter-wavelength, otherwise the structure would start radiating and the circuit losses would increase.

The following experimental results have been measured on a structure with the dimensions as shown in Fig. 4. Two transitions are in series. The alumina substrate has the dimensions 1 × 1 in and a thickness of 0.51 mm. In Fig. 5 the measured insertion and return losses of the structure of Fig. 4 are plotted versus frequency. The losses, which also include the connector losses, are rather low, going below 0.5 dB. This means that losses for one transition should be below 0.15 dB in the frequency range 2–6 GHz and below 0.2 dB in the range 6–9 GHz. The return loss, which includes the interference of four discontinuities, i.e., two connectors and two transitions, is mostly below −20 dB.

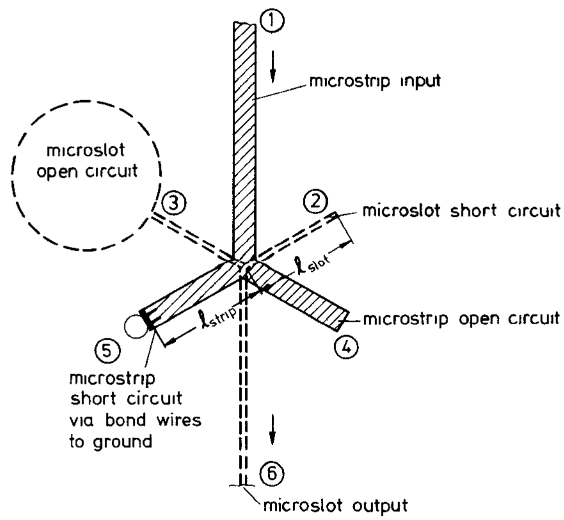

Fig. 3. Microstrip-slot transition with open- and short-circuited lines.

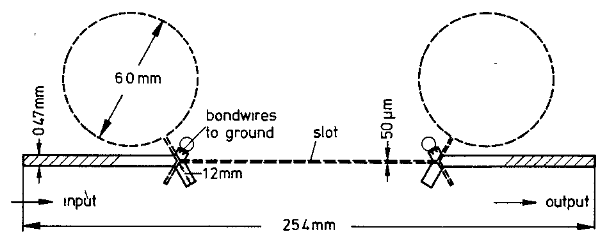

Fig. 4. Dimensions of the circuit layout of two transitions in series. The substrate material is Al_2O_3 of 0.51-mm thickness; $\varepsilon_r \simeq 10$.

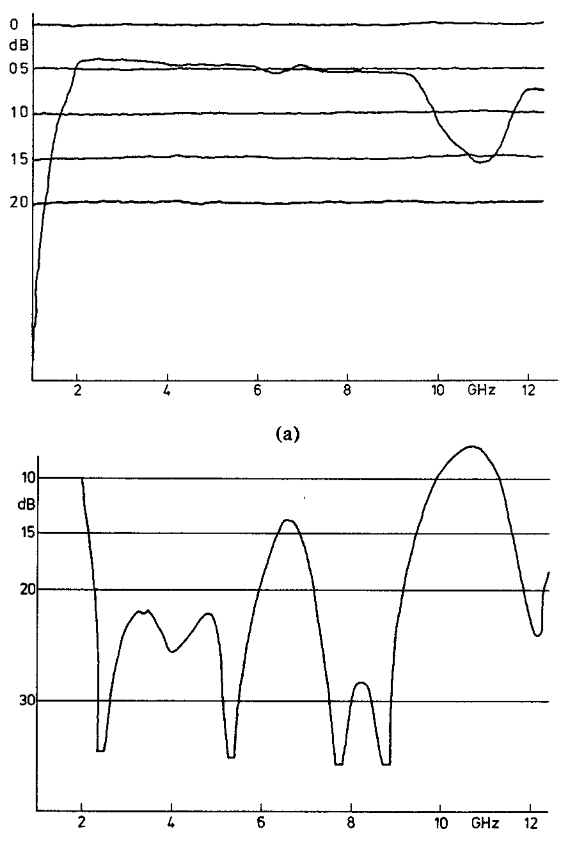

Fig. 5. (a) Measured insertion loss for the two transitions in series as given in Fig. 4. (b) Measured return loss of the two transitions in series as given in Fig. 4.

In the structure of Fig. 4 the microstrip, which is short-circuited, is somewhat reduced in length in order to compensate for the inductance of the bondwires to ground. The slot width was either 40, 50, or 100 μm, while the strip width remained unchanged. However, this had only little effect upon the performance, as the characteristic impedance of the slot is only weakly dependent on the slot width. According to calculations in [8], the characteristic impedance of a slotline on a substrate as in Fig. 4 increases from 51 Ω for a slot width of 40 μm to 62 Ω for 100 μm, at a frequency of 6 GHz. It is not really known how good the model of an ideal junction with perfect coupling between the slot and the striplines is, but the experimental results seem to substantiate this simple model. Also, an equivalent circuit for a slot open circuit as used in Fig. 4 is still lacking.

The increase in insertion loss below 2.0 GHz is partly due to the finite diameter of the slot open-circuit area being rather small for that frequency, and to the high-pass behavior of the slotline itself. Above 9.5 GHz it is due to the aforementioned radiation effects. Above this radiation frequency the losses decrease again.

Possible applications are that two of the structures of Fig. 4 may be used to realize a broad-band 180° phase shift. For this purpose the short circuits have to become open circuits and vice versa for one of the four transitions.

There are different ways to construct broad-band-balanced or double-balanced mixers from this transition. This can be accomplished, e.g., with one strip-slot transition in conjunction with the double junction of Fig. 2. Diodes can be put either as terminations to ports 4 and 5 (with ports 2 and 3 passively matched) or to ports 2–5.

Another way to construct a double-balanced mixer is to put an internally crossed-over diode quartet into the slotline, which is interrupted by two open circuits.

Acknowledgment

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Correction to "Efficient Minimax Design of Networks Without Using Derivatives"

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In the above paper,[1] on page 804, the logic of the flow diagram in Fig. 1 appears to be somewhat in error. The corrected flow diagram is shown here in Fig 1.

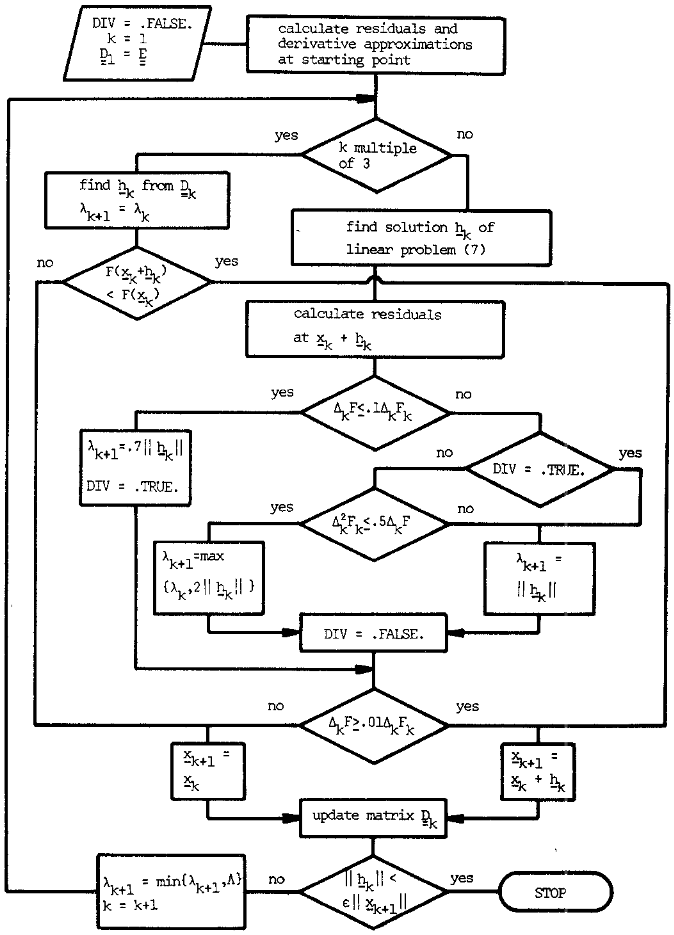

Fig. 1. Corrected flow diagram of minimax algorithm.

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