An Interdigitated 3-dB Coupler with Three Strips

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Abstract—**An interdigitated microstrip coupler on alumina substrate with three strips and two bond wires is described that provides an alternative to the four-strip Lange coupler. The useful coupling range is between 3 and 6 dB. Experimental results for a 3-dB coupler in the frequency range 4.5–9 GHz are presented together with theoretical results for 3–5-dB coupling.**

I. Introduction

THE interdigitated microstrip 3-dB coupler as first described by Lange [1] has now been widely accepted [2], [3]. However, the interdigitated coupler needs four bond wires, when it is designed to have the direct and the coupled ports adjacent, and half the bonds are within the structure, where the space is limited. This leads to technological problems at higher microwave frequencies, when standard thin-film and bond techniques are applied. In addition an "unfolded" Lange coupler has been proposed [4] that only needs two bond wires. This structure, however, does not have the direct and coupled ports adjacent.

We therefore raised the question whether it might be possible to obtain 3-dB coupling in thin-film technique with only three strips instead of four. This is, indeed, the case, as shall be shown below.

A design with the direct and coupled ports adjacent is shown in Fig. 1. Only two bond wires are needed and additional space may be provided for the bondings (Fig. 1). The gap width for this type of coupler on an alumina substrate 0.51 mm thick is in the order of 20 μm, which can be realized with present technology. Of course, a backward coupler design is also possible. Furthermore the interdigitated coupler with three strips seems to be particularly suited to achieve coupling values between 3 and 6 dB.

A similar structure has already been suggested [5] and -8.34 dB coupling was realized in stripline.

II. Theoretical Approach

A. General Considerations

In a system of three coupled microstriplines, there exist three independent modes with different phase velocities. If we assume a symmetrical structure, the modes can be characterized as follows: I: quasi - even mode, II: quasi - odd mode, III: odd mode between outer conductors (the middle conductor is at ground potential). Further information about this structure is given in [6].

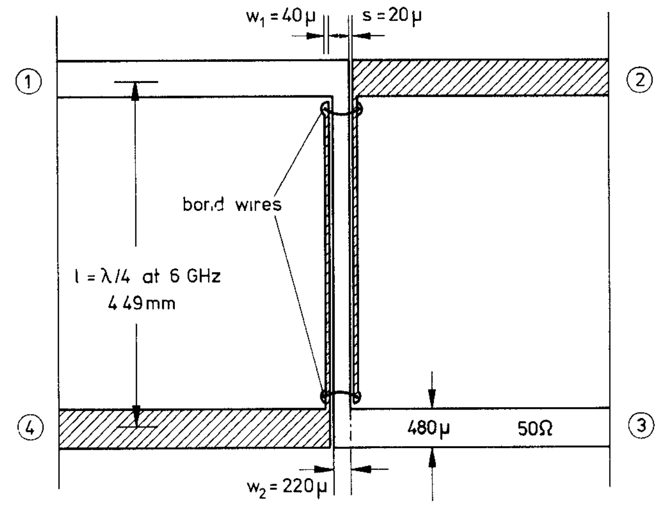

Fig. 1. Calculated geometrical layout of an interdigitated three-strip 3-dB coupler. Substrate thickness 0.51 mm; Al_2O_3; $\epsilon_r = 10$.

Modes I and II can be described as "natural" modes of the coupler. Mode III is supposed to be suppressed by the boundary conditions that are imposed by the bonding connections. In order to optimize the coupler over a given frequency band, we have three degrees of freedom, i.e., (Fig. 1) w_1/h (relative width of the outer strips), w_2/h (relative width of the middle strip), s/h (relative spacing between strips), where h is the thickness of the substrate. The permittivity ϵ_r is determined by the selected material, and the length l of the coupler determines the midband frequency.

From a physical viewpoint the coupler can be considered as two parallel-working two-line couplers. Unfortunately this does not lead to a simplification, because these two-line couplers have a magnetic wall at one vertical boundary. Results for such couplers are not known and therefore a proper analysis of the given system was necessary.

B. Computation of the Performance

The structure is not too complicated for a computer analysis of the scattering parameters at the four-ports. The only simplification made was the quasi-TEM assumption, which, as is well known, leads to good results for design

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purposes. From given cross-sectional dimensions, we have to calculate the distributed static capacitances for the vacuum and the dielectric filled cases.

We used the "network analogue method" described by Lennartsson [7], [8]. It is appropriate at this point, to report about a significant improvement to this method which has been successfully tested. Lennartsson subdivides the cross section by an equidistant network in both the horizontal and vertical direction. As this is not necessary [9], we assumed a "network" which had infinitely small node distances in the vertical direction, while the horizontal spacing remained finite. This leads to limiting value formulas for the elements of the matrix $[R_{tot}]$ (see [8]) and we could spare the time consuming recurrence algorithm and directly establish $[R_{tot}]$. Moreover the computing accuracy was increased. For the special case of planar strips at only one interface between two dielectrics ϵ_1, ϵ_2, we found (nomenclature as in [8])

$$(R_{tot})_{i,j} = \frac{1}{N+1} \sum_{k=1}^{N} \rho_k \left\{ \cos\left[\frac{k\pi}{N+1}(i-j) \right] - \cos\left[\frac{k\pi}{N+1}(i+j) \right] \right\}$$

with

$$1/\rho_k = 2 \sin \frac{k\pi}{2(N+1)} \left[\frac{\epsilon_{r1}}{\tanh \frac{h_1}{\Delta x}} + \frac{\epsilon_{r2}}{\tanh \frac{h_2}{\Delta x}} \right]$$

where h_1 and h_2 are the thicknesses of the two dielectrics and Δx is the distance between two network nodes in the horizontal direction. In the special case of the described coupler, the term with ϵ_2 degenerates to 1 because of $\epsilon_{r2} = 1$, $h_2 = \infty$.

Having calculated the distributed capacitance matrix we have to find the admittance matrix $[Y]$. A recent paper by Tripathi [6] gives formulas for this special case, which can be evaluated directly. The six-port is reduced to a four-port by bonding the outer strips together at the line ends. Mathematically this means, that the corresponding rows and columns in the Y-matrix have to be summed. Since the Y-matrix of the coupler does not conveniently describe the coupler performance in a given system, we form the scattering matrix $[S]$ assuming that all ports of the coupler are connected to lines with characteristic impedance Z_0.

$$[S] = \left(\frac{1}{Z_0}[1] + [Y] \right)^{-1} \cdot \left(\frac{1}{Z_0}[1] - [Y] \right).$$

In the present case of a reactance network, $[S]$ is of the form

$$[S] = [A]^{-1}[A]^*$$

which simplifies programming considerably.

C. Optimization and Results

The three parameters w_1, w_2, and s have been determined numerically for some optimum couplers in a range from 2.9- to 5.0-dB coupling. The permittivity assumed was 10.0. For the computation of the capacitance matrix a scaling of $\Delta x = 500$ nodes/mm was chosen. The entire width of the structure has been chosen to be 20 mm, so that N was about 10.000. The selected substrate material had an h of 0.51 mm.

The optimization is simplified by the fact, that the coupling mainly depends on the spacing s. Therefore, we took a given s and optimized w_1 and w_2 in such a way, that the reflected and isolated signals were minimized. At the midband frequency the line length is a quarter wavelength long for the mean of the two mode phase velocities. The results for the coupling parameters and the optimum line widths depending on s are given in Table I. The

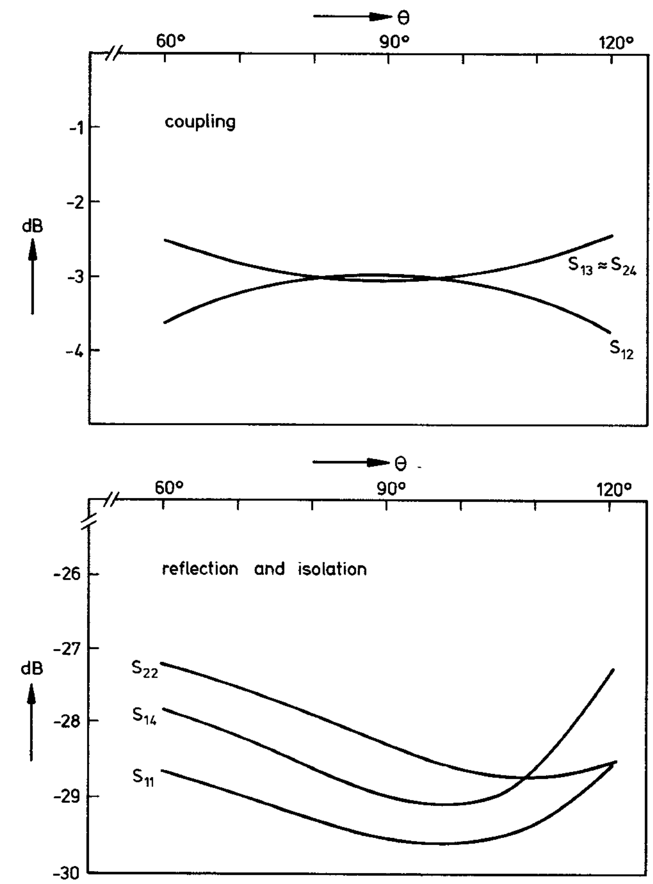

Fig. 2. Calculated performance versus average line angle θ of the three-strip 3-dB coupler with parameters as given in Table I ($s = 0.0225$ mm).

TABLE I
RESULTS FOR $h = 0.51$ mm, $\epsilon_r = 10$

s [mm]	w_1 [mm]	w_2 [mm]	db at midband freq. S_{21}	S_{31}	S_{42}	β_I/β_0	β_{II}/β_0
0.02	0.03	0.17	- 2.86	- 3.19	- 3.19	2.53	2.35
*0.0225	0.0325	0.175	- 2.98	- 3.07	- 3.07	2.53	2.35
0.0325	0.035	0.20	- 3.46	- 2.63	- 2.64	2.55	2.35
0.04	0.0375	0.215	- 3.76	- 2.40	- 2.40	2.55	2.35
0.045	0.0375	0.225	- 3.96	- 2.26	- 2.26	2.56	2.35
0.05	0.0375	0.2325	- 4.16	- 2.14	- 2.14	2.56	2.35
0.06	0.04	0.245	- 4.50	- 1.94	- 1.94	2.57	2.35
0.07	0.04	0.255	- 4.84	- 1.77	- 1.77	2.57	2.35

*Theoretical performance versus frequency is shown in Fig. 2.

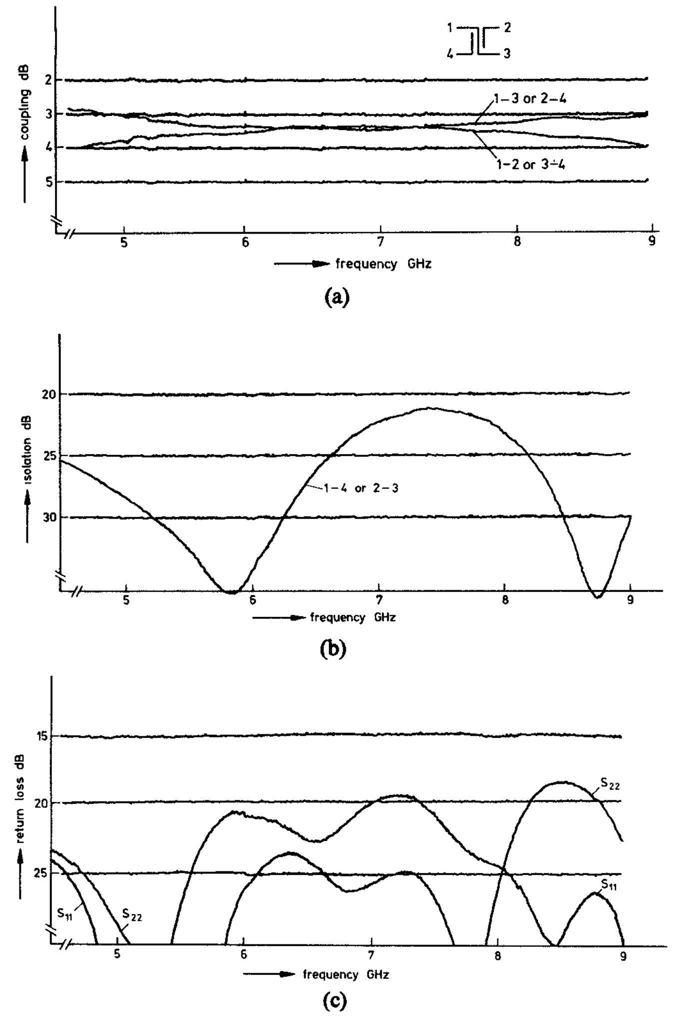

Fig. 3. Measured performance of a three-strip 3-dB coupler with the layout as given in Fig. 1.

isolation and return loss were lower than -25 dB over the frequency band. The calculated phase difference between S_{21} and S_{31} was between 89.9 and 90.15 degrees over an octave band. An example of the calculated performance of a 3-dB coupler is given in Fig. 2.

III. Experimental Results

Several couplers have been built for the range 4.5 to 9.0 GHz. The measured results for the coupling and isolation as well as the input return loss of a practical device are shown in Fig. 3. Over an octave frequency band, the measured return loss was better than 18 dB, the measured isolation was better than 21 dB and the coupling shows the typical behavior of a transmission line coupler.

The substrate thickness was 0.51 mm, the substrate material used was Al_2O_3 and the diameter of the bond wires was 10 μm. All dimensions were obtained from the theoretical analysis outlined above.

The losses include the connecting lines to the coupler on a substrate 1 in by 1 in and the coaxial-to-microstrip transitions. Another layout had a gap width of $s \approx 13$ μm. In this case the coupling was stronger, corresponding to an ideal coupling of 2.5 dB at the midband frequency. The thickness of the gold layer was 4 μm.

To a first approximation, the gap width s only influences the coupling (Table I). The theory is based on infinitely thin conductors, whereas in practice the thickness of the gold layer is by no means negligible compared with s. Therefore, in order to obtain a reproducible coupling, not only the gap width s has to be controlled carefully, but also the thickness and the edge definition of the gold layers.

Acknowledgment

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References

[1] J. Lange, "Interdigitated strip-line quadrature hybrid," in *1969 G-MTT Int. Microwave Symp.* pp. 10–13.
[2] W. P. Ou, "Design equations for an interdigitated directional coupler," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp. 253–255, Feb. 1975.
[3] D. D. Paolino, "Design more accurate interdigitated couplers," *Microwaves*, pp. 34–38, May 1976.
[4] R. Waugh and D. La Combe, "Unfolding the Lange Coupler," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 777–779, Nov. 1972.
[5] J. A. Malherbe, "Interdigital directional coupler with an odd or even number of lines and unequal characteristic impedances," *Elec. Lett.*, vol. 12, pp. 454–465, Sept. 1976.
[6] V. K. Tripathi, "On the analysis of symmetrical three-line microstrip circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 25, pp. 726–729, Sept. 1977.
[7] B. L. Lennartsson, "A Network analogue method for computing the TEM characteristics of planar transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 586–591, Sept. 1972.
[8] ——, "Capacitance matrix calculations for planar transmission lines imbedded in composite dielectric materials," Technical Report No. 49 (1974), Chalmers University of Technology, Göteborg, Sweden.
[9] U. Schulz, "Improvements and simplifications of network analogue capacitance matrix calculations, private communication (to be published).