ANALYSIS AND OPTIMIZATION OF E-PLANE DIRECTIONAL COUPLERS

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ABSTRACT

A procedure for the simulation of E-plane coupled slot hybrids is proposed. The component is divided into a cascade of uniformly coupled subsections which are individually analyzed with a spectral domain program. The validity of the approach is confirmed by measurements on several couplers.

INTRODUCTION

Directional couplers, and in particular hybrid couplers, are a basic component of various communication systems: Balanced mixers, image-rejection mixers, QPSK modulators, etc... In E-plane technology, several configurations have been investigated for the realization of hybrid couplers. The printed-probe coupler introduced by Meier [1] is one of them. It consists of an array of printed probes on a dielectric substrate mounted in the E-plane common to two adjacent waveguides. The transfer of energy from one waveguide to the other is controlled by varying the size and position of the printed probes. The resulting couplers turn out to have a fairly limited usable bandwidth.

Another approach, the branchline hybrid coupler which has been realized in two slightly different ways in [2] and [3], is essentially similar to deRonde's microstrip coupler [4]. Although it demonstrates wideband behavior, this approach necessitates double-sided metallization of the substrate or the use of a bond wire which complicates the fabrication process.

A third alternative, the coupled slot hybrid, consists of two adjacent slots sharing the same housing. The transfer of energy from one slot to the other is controlled by their width, spacing and length. This technique appears advantageous because it combines wideband performances, and simplicity of fabrication. The structure also produces forward coupling, which allows easier connections between the components of an integrated system than backward coupling. The main inconvenient, however, is that this type of coupler is difficult to analyze rigorously. The reason for that is the non-uniform coupling produced by the curved slots connecting the main coupling section to the ports. The only design procedure reported so far which systematically accounts for the excess coupling in the non-uniform regions is that by Callsen *et al* [5]. However, this approach concentrates on the forward coupling and neglects the reflections produced by the non-uniform sections. All other reported realizations were optimized using an empirical approach based on trial and error [6,7].

This contribution alleviates this problem by presenting a simple procedure that allows the numerical simulation of non-uniform coupled slot hybrids. The procedure has been tested with a number of circuits, and a good agreement was obtained with the measured results. This indicates that the proposed method is suitable for the design of these circuits.

MODEL OF FINLINE COUPLED-SLOT HYBRIDS

Figure 1 shows a typical finline coupled slot hybrid. It consists of a main coupling section with two parallel slots connected at each port to single slots. In order to minimize parasitic reflections, the single slots form a smooth transition to the main section, and tapered rounded elbows are used at the junctions. The regions of convergence of the single slots will be referred to as the non-uniform regions, producing a so-called excess coupling.

The two coupled slots form a coupled transmission-line arrangement, supporting two normal modes of propagation. In the symmetric cases (where both lines or slots have individually the same properties) these modes degenerate to the even and odd modes, having equal voltage amplitude on the two lines with, respectively, same and opposite polarity. The different propagation constants $\beta_{e,o}$ of the modes create a beating effect responsible for the coupling (periodical coupling). For uniformly coupled lines with similar even and odd mode impedance values ($Z_e \approx Z_o \approx Z_L$, where Z_L is the load impedance at each port), it can be shown that for an excitation at port 1, the ratio of the output voltages V_3 and V_4 (see figure 1) is given by:

$$\left|\frac{V_3}{V_4}\right| = \tan\left(\frac{\theta_e - \theta_o}{2}\right). \tag{1}$$

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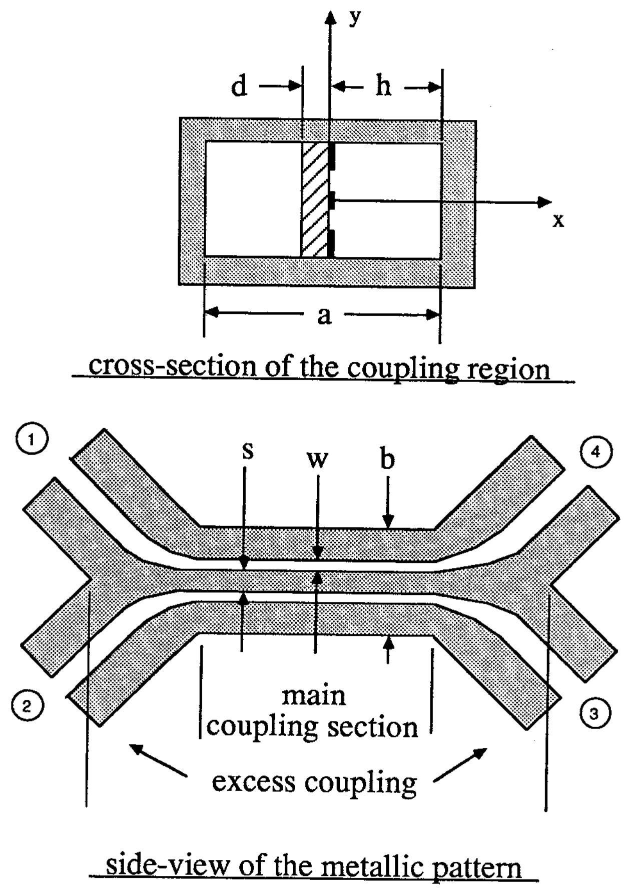

Figure 1: Typical finline coupled slot hybrid: cross-section and side view.

$\theta_{e,o} = \beta_{e,o}\, l$ are respectively the even and odd mode electrical length of the lines.

In the previous attempts to design hybrid couplers using this result, only the main coupling section had been considered, yielding insufficient accuracy [6,8]. The excess coupling had been neglected because it is non-uniform in nature and therefore difficult to evaluate theoretically. Callsen *et al* [5] took into account, to some extent, the excess coupling by the use of the *effective coupling length* concept. That is, they approximated the coupling length of the structure by the summation of the coupling lengths of a number i of subsections, each considered uniform:

$$\left|\frac{V_3}{V_4}\right| = \tan \sum_i \left(\frac{\theta_{ei} - \theta_{oi}}{2}\right). \tag{2}$$

This concept neglects the variation of impedance along the lines and, unfortunately, no results are presented to show how well their model agrees with the experiment. The present approach uses a similar decomposition in conjunction with simple circuit theory, to include the effect of the varying impedances in the analysis.

As a first step, the known physical layout is decomposed longitudinally into a number of subsections. Each of these is then considered as an independent uniformly coupled slot stucture and analyzed with the spectral domain method (see following section). An equivalent coupled transmission-line circuit is obtained with its even and odd parameters. The equivalent circuits are then connected in cascade and analyzed with a circuit analysis program to obtain the overall response. The load impedance at each port corresponds to the characteristic impedance of the single slots. The information related to the reflections produced along the slots is preserved by consideration of the equivalent line impedances and, consequently, the model can predict relatively well both the transmission and reflection coefficients of the coupler.

SPECTRAL DOMAIN PROGRAM

The spectral domain method was used to compute the even and odd parameters of the uniform coupled slot subsections. This method is very efficient when certain qualitative features, such as the edge condition, are incorporated in the choice of the basis functions for the electric field in the slots. Accordingly, the following one-term expansions have been used for the longitudinal and transverse components [9]:

$$\begin{aligned} E_{\text{long}} &= \frac{2t}{w}\sqrt{1-\left(\frac{2t}{w}\right)^2} \\ E_{\text{trans}} &= \left(1-\left(\frac{2t}{w}\right)^2\right)^{-1/2} \end{aligned} \tag{3}$$

where $t = y - (s + w)/2$.

The results of the program have been checked against those published by Schmidt *et al* in [10], who used up to five edge-corrected sinusoidal basis functions for each component of the field. For the dominant even and odd modes, an excellent agreement for the propagation constant and the impedance has been obtained in all cases. It thus appears that the basis functions chosen here are adequate for most practical applications.

SIMULATION OF THREE COUPLERS

The simulation procedure has been applied to three circuits described in [11]. Their topology is that of figure 1. The single slots make a 45 degree angle with the uniform coupling section. The housing dimensions for the input lines and the main coupling section are a = 7.112 mm and b = 3.556 mm (WR-28). The substrate has a thickness d = 0.254 mm with ϵ_r = 2.2 (RT/Duroid). The distance between the metallic pattern and the housing is h = 3.556 mm. The input finlines have a slotwidth of 0.381 mm. At each port, a smooth taper is used to provide compatibily with waveguide test equipment.

The circuits have been obtained in the process of optimizing a unit for equal power split ($V_2 = V_3$) at 32 GHz. They were all mounted in the same housing and, therefore, had all the same length (main section = 11.1 mm). The optimization was performed by empirically varying the slot width w and spacing s of the main section from one circuit to the other. Their frequency responses are shown in figures 3 to 5.

In each circuit, the portion effectively contributing to the coupling extends up to the point where the two slots become separated by the enclosure. For circuits A and B (figures 3 and 4) the coupling region has been decomposed into 9 uniform subsections as shown in figure 2. The uniform coupling region constitutes the longest subsection. The bends at each end are segmented into three subsections of equal length, and one slightly longer where the coupling is known to be less important. The slot width, spacing, and the enclosure height for each subsection were taken as their average values over the length of the subsection. Similarly, circuit C (figures 5 and 6) has been simulated successively with 3 and 17 sections to compare the accuracy of the results.

COMPARISON OF THE CALCULATED AND EXPERIMENTAL RESULTS

The frequency responses of the models were calculated at a number of different frequencies because of the dispersive nature of finlines. They are compared in figures 3 to 6 with the experimental curves. The 9 subsection models for circuits A and B yield fairly accurate results for the S_{41} and S_{31} coefficients. The S_{21} and S_{11} responses also display a good agreement, considering that the waveguide to finline transitions (−20 to −25 dB return loss) have not been included in the model.

The comparison of the results for the 3 and 17 section models of circuit C show that the S_{31} and S_{41} responses are not very sensitive to the number of subsections. This can be explained by the fact that for slot profiles varying very smoothly (as in this case), the characteristics of the coupled slots can be averaged over a longer subsection without affecting significantly the total coupling length (see (2)). On the other hand, the two figures clearly reveal that the 17 section model is in much better agreement with the experiment than the other one regarding the S_{11} and S_{21} coefficients, as was expected.

It is interesting to see that for circuits B and C, the experimental S_{31} and S_{41} responses are more broadband than calculated. Part of the reason for this is a housing resonance occuring in the Y junction at a frequency of 38 GHz. This resonance perturbs the responses in such a way as to broaden the bandwidth. This effect is not noticeable for circuit A, the results being available only up to 32 GHz.

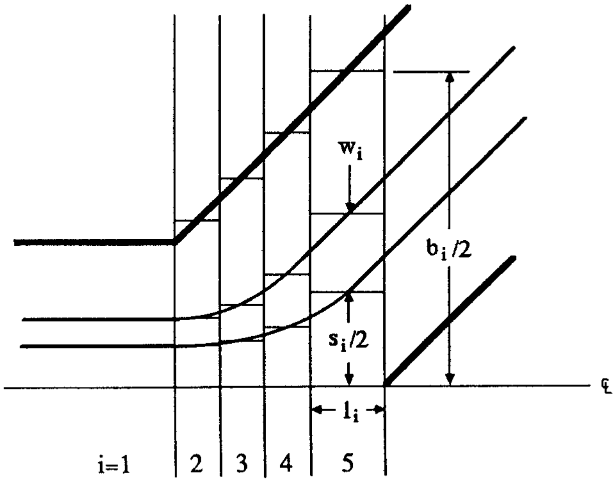

Figure 2: Details of the 9 subsection decomposition for circuits A and B. One quarter of the structure is shown.

The accuracy achieved by the simulations presented here shows the potential of the method for design purposes. This procedure also opens the door to the investigation of other configurations, for instance coupling between asymmetric slots or non-parallel slots.

CONCLUSION

A simulation procedure for non-uniform coupled slots has been reported here. It relies on the field analysis of a number of cascaded uniform subsections. It has been found that both the coupling level and the reflection responses were well predicted by this approach. Its simplicity allows its implementation into CAD programs with user-defined models, and it can be extended to almost any configuration involving coupled slots.

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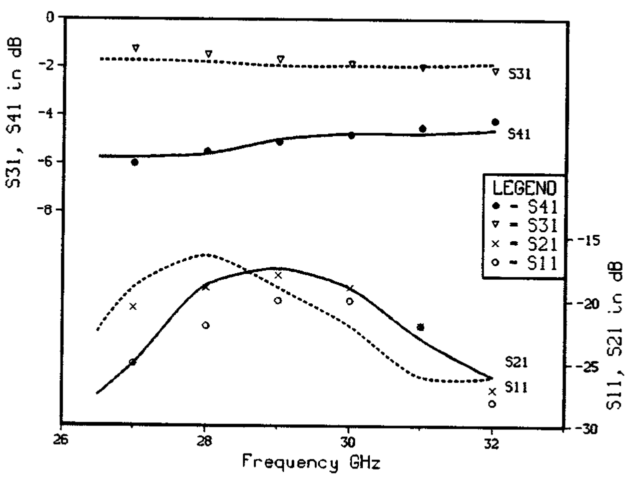

Figure 3: Circuit A, s = 1.397 mm, w = 0.381 mm: 9 section model. Experimental (continuous curves) and calculated (discrete points) responses.

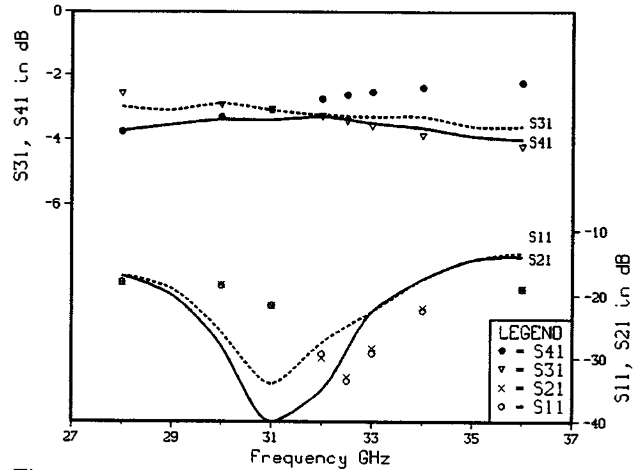

Figure 4: Circuit B, s = 1.905 mm, w = 0.254 mm: 9 section model. Experimental (continuous curves) and calculated (discrete points) responses.

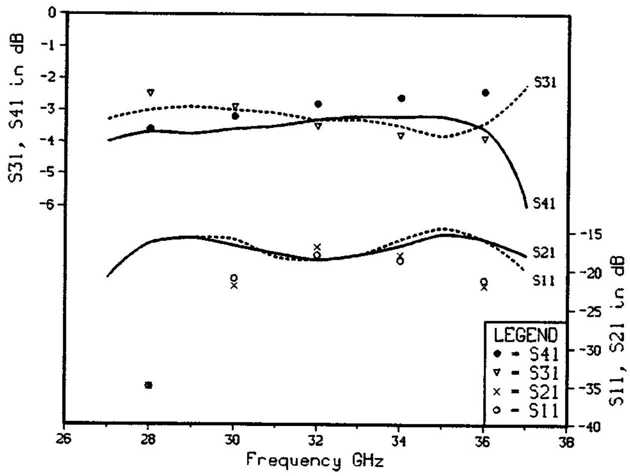

Figure 5: Circuit C, s = 1.40 mm, w = 0.178 mm: 3 section model. Experimental (continuous curves) and calculated (discrete points) responses.

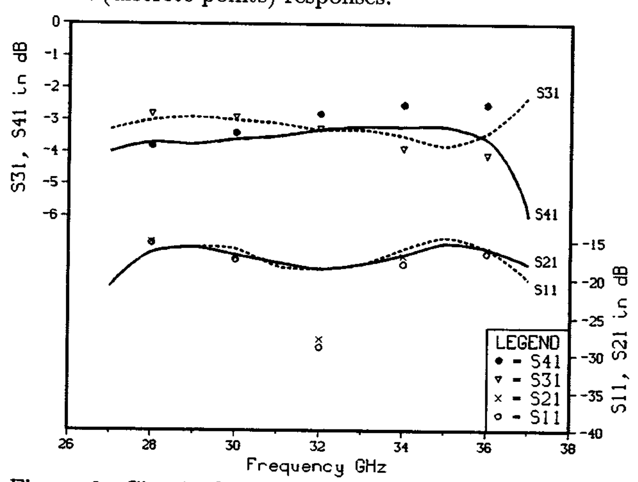

Figure 6: Circuit C, s = 1.40 mm, w = 0.178 mm: 17 section model. Experimental (continuous curves) and calculated (discrete points) responses.