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RIGOROUS MODAL-S-MATRIX DESIGN OF A NEW CLASS OF BROADBAND 180-DEGREE BRANCH GUIDE COUPLERS

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ABSTRACT

A new class of 180-degree branch guide couplers is introduced which combines the advantages of the ultra-broadband potential of conventional waveguide E-plane multiple-branch couplers with the low-insertion-loss qualities of E-plane stub-loaded phase shifters. Based on the modal scattering matrix method, the rigorous design takes into account both the finite branch heights and the higher order mode interaction at all step discontinuities. Computer-optimized five-branch three-stub prototypes, designed for 3 ± 0.2 dB coupling, for the waveguide Ku- (12 - 18 GHz) and Ka-bands (26 - 40 GHz), respectively, achieve typically a $180^\circ \pm 1^\circ$ differential phase shift at the output ports, within about 19 percent bandwidth, as well as more than 30 dB isolation and return loss. The theory is verified by available measured results.

INTRODUCTION

Branch guide coupling is a well-known technique to design 90-degree directional couplers with the potential of wide variation of coupling, directivity, input VSWR, and bandwidth values [1] - [5]. For many applications, however, such as for antenna beam forming networks, special test assemblies, or variable power dividers, [6] - [8], 180-degree couplers with input 'sum' and 'difference' ports are required where the corresponding incident signals divide equally between the output ports, being in phase, or 180° out of phase, respectively. The conventional magic-T or rat-race-hybrid techniques often used to provide that characteristic are too narrow-band for many purposes, [6] - [8]. This paper introduces a rigorous computer-aided design for novel broadband 180-degree couplers (Fig. 1) which integrate advantageously the broadband characteristics of multiple-branch couplers with those of stub-loaded phase shifters [9].

The design is based on the orthogonal expansion method which allows direct inclusion of higher order mode coupling, finite branch height, and discontinuity effects. Matching the fields at common interfaces yields the corresponding modal scattering matrix directly. An optimizing computer program varies the coupler and stub parameters until coupling, return loss, isolation and differential output phase shift correspond to predicted values. The evolution strategy method [9], i.e. a modified direct search where the parameters are varied statistically, is applied which requires no differentiation step, and, hence, helps to circumvent the problem of local minima.

The advantages of the waveguide E-plane stub loading principle are such that good compatibility with the E-plane branch coupling technique is achieved, that convenient milling and spark eroding techniques from a solid block allow compact and low-weight complete 180-degree coupler designs, that no severe problems may arise

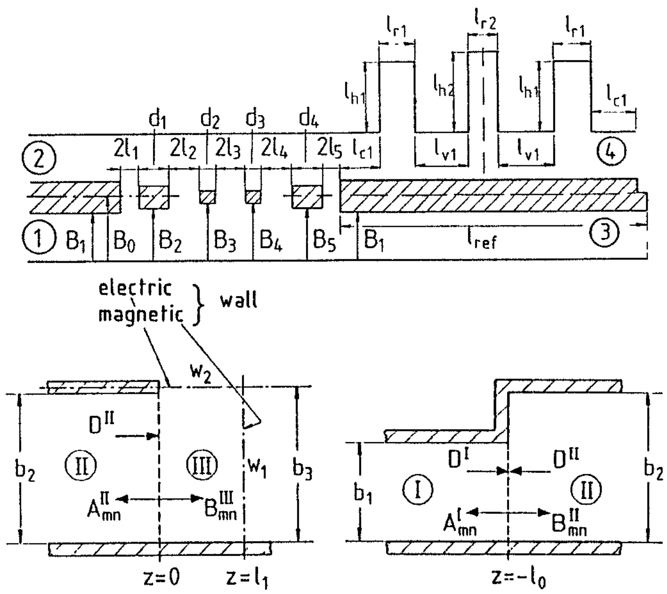

Fig. 1: Broadband 180-degree branch guide coupler
Coupler geometry and key-building block discontinuities

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to meet the power specifications as no dielectric or ferrite materials are necessary, and that good VSWR, low-insertion-loss and broad-band phase shift characteristics may be obtained by appropriate computer optimization of all relevant parameters. Moreover, the E-plane technique is highly compatible with printed E-plane components, such as diplexers [10], [11], and appropriate for millimeter-wave designs [4].

Computer-optimized examples for five-branch three-stub 180-degree coupler prototypes for the waveguide Ku- (12 - 18 GHz) and Ka-bands (26 - 40 GHz), respectively, may demonstrate the efficiency of the new design. The theory is experimentally verified by available measured results for the single components.

THEORY

In solving for the modal S-matrix representation of the 180-degree coupler (Fig. 1), we only require the field theory solution for two waveguide discontinuities which represent the two key building blocks: 1) step discontinuity at the symmetric coupling slot section [4] (where, due to the application of the even and odd normal mode principle [12], merely a one-quarter section has to be taken into account with four boundary cases: electric and/or magnetic wall at w_1 and/or w_2, respectively), 2) step discontinuity at the change in waveguide height [9]. Note that for the corresponding inverse discontinuities only the port designations of the related modal scattering matrix need to be interchanged. The total scattering matrix of the structure under consideration is formulated by suitable direct combination of the individual modal scattering matrices (S^C), and (S^H), of the coupling section and the step in waveguide height, respectively, by an iteration process already described in [13], and by including appropriately the known scattering matrices (S^W) of a homogeneous waveguide section, for adequate consideration of the distances between the individual discontinuities (Fig. 1). This procedure preserves numerical accuracy, avoids instabilities, and requires no symmetry of modes [13].

A TE_{10} wave incident in port 1 (Fig. 1) excites longitudinal section TE^x_{1n}-waves [4], [9] at all step discontinuities. As in [4], therefore, for the homogeneous subregions ν = I, II, III (Fig.1) the fields

$$\vec{E}^{(\nu)} = -j\omega\mu\,\nabla\times\vec{\Pi}^{(\nu)}_{hx}, \quad \vec{H}^{(\nu)} = \nabla\times\nabla\times\vec{\Pi}^{(\nu)}_{hx} \qquad (1)$$

are derived from the x-component of the magnetic Hertzian vector potential $\vec{\Pi}_h$, which is assumed to be a sum of suitable eigenmodes satisfying the vector Helmholtz equation and the boundary conditions:

$$\Pi^{(\nu)}_{hx} = \frac{2}{\sqrt{ab^{(\nu)}}} \cdot \frac{1}{\sqrt{k^2-k_x^2}} \cdot \sin(k_x x) \cdot \sum_{n=0}^{N} \left[\frac{1}{\sqrt{Z^{(\nu)}_{Fn}}} \cdot \frac{1}{\beta^{(\nu)}_n} \cdot \frac{1}{\sqrt{1+\delta_{0n}}} \cdot \cos(k^{(\nu)}_{yn} y)\right] \cdot \left[A^{(\nu)}_n e^{-j\beta^{(\nu)}_n z} - B^{(\nu)}_n e^{+j\beta^{(\nu)}_n z}\right], \qquad (2)$$

where a = waveguide width, $b^{(\nu)}$ = waveguide height in the subregion ν, $k^2=\omega^2\mu\epsilon$, $k_x=\frac{\pi}{a}$, $Z^{(\nu)}_{Fn}=\omega\mu/\beta^{(\nu)}_n$, $k^{(\nu)}_y=n\pi/b^{(\nu)}$, δ_{0n} = Kronecker delta, $k_c^2=k_x^2+k_y^{(\nu)2}$

$$\beta^{(\nu)}_n = \begin{cases} \sqrt{\omega^2\mu\epsilon - (k_x^2 + k_y^{(\nu)2})} & \omega^2\mu\epsilon \ge k_c^2 \\ -j\sqrt{(k_x^2 + k_y^{(\nu)2}) - \omega^2\mu\epsilon} & \omega^2\mu\epsilon < k_c^2 \end{cases}.$$

The eigenmodes in (2) with the still unknown amplitude coefficients A_n and B_n are suitably normalized, so that the power carried by a given wave is 1W for a wave amplitude coefficient of $1\sqrt{W}$ [4].

By matching the tangential field components at the common interface between the subregions (Fig. 1), and utilizing the orthogonal property of the modes [4], [9] the amplitude coefficients of (2) are related to each other in the form of the desired modal scattering matrix (S^H), of the discontinuity change in waveguide height [9], and (S^{Ci}_{1p}), of the input reflection coefficient of the one-port equivalent of the one-quarter section III for the four boundary cases i = 1, 2, 3, 4 (magnetic wall at w_1, electric wall at w_2, etc., cf. [4])

$$S^{Ci1q}_{1p1n} = \sqrt{\frac{|k^{II}_{z1n}|}{|k_{z1q}|}} \cdot A^{II1q}_{1n}\Big|_{Ci} \qquad (3)$$

where A^{IImq}_{mn} is the amplitude coefficient of the TE^x_{1n}-wave reflected in the subregion II excited by a TE^x_{1q}-wave incident with the amplitude $D^{II}=1$. The four-port modal S-matrix (S^C) of the symmetric total coupling section is then calculated by the even and odd mode relations given in [12].

For computer optimization, the expansion into ten eigenmodes at each step discontinuity and four eigenmodes along each intermediate homogeneous waveguide section has turned out to yield sufficient asymptotic behaviour of the coefficients of the total scattering matrix of the structure. The final design data are provided by expansion into twenty eigenmodes.

The computer-aided design is carried out by an optimizing program [9] applying the evolution strategy method, i.e. a modified direct search method, which varies the input parameters statistically until the desired 180-degree coupler characteristic for a given bandwidth are obtained. For desired waveguide housing dimensions a, b, as well as number of coupling and stub sections, the parameters to be optimized are the dimensions of the coupling branches and of the stub sections, the lengths of the intermediate waveguide sections, and the length of the reference waveguide section. The number of frequency sample points was chosen to be ten.

RESULTS

Fig. 2 shows the calculated and measured scattering parameters of an optimized 3dB-5-branch coupler for the Ku-band (12 - 18 GHz, R140 waveguide housings: 15.799mm × 7.899mm). Good agreement between theory and measurements may be stated. The same is true for the two-stub phase shifter in Fig. 3, for the waveguide R120-band (10 - 15 GHz, 19.050mm × 9.525mm). The realized component [9] shows a measured insertion loss of only about 0.1 dB, and a phase error of about $+2.5^{\circ}/-1^{\circ}$, for about 17% bandwidth.

The scattering parameters in decibels as well as the differential phase shift between the output ports 4 and 3 (input port 1) as a function of frequency are shown in Fig. 4 for a computer-optimized 180-degree 5-branch-3-stub coupler design example in the Ku-band. 30 dB return loss and isolation are achieved, together with 3dB ± 0.2dB coupling, between 13.3 and 16.4 GHz; $180^{\circ} \pm 1^{\circ}$ differential phase shift between the output ports is maintained between about 13 and 16 GHz.

Fig. 5 shows the results of a computer-optimized design example for the waveguide Ka-band (26 - 40 GHz, R320 waveguide housings, 7.112mm × 3.556mm). $180^{\circ} \pm 1^{\circ}$ differential phase characteristic, together with 30dB return loss and isolation, is achieved within about 19 percent bandwidth.

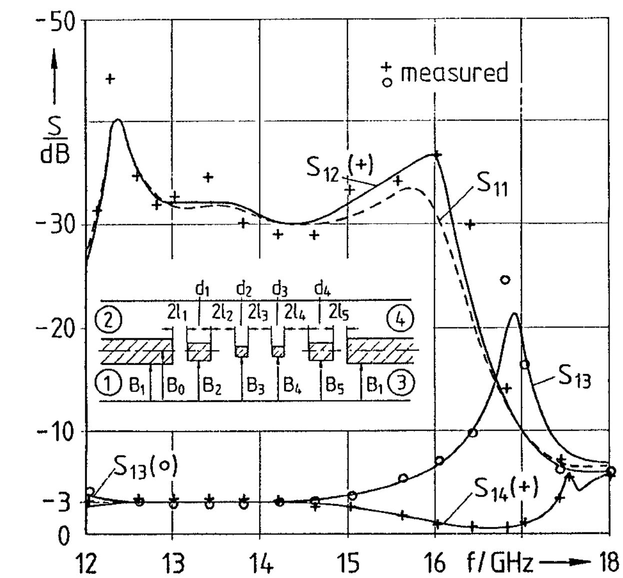

Fig. 2: Computer-optimized Ku-band 90-degree 3dB-5-branch coupler
Calculated and measured scattering parameters (R140 waveguide housing: 15.799mm × 7.899mm).

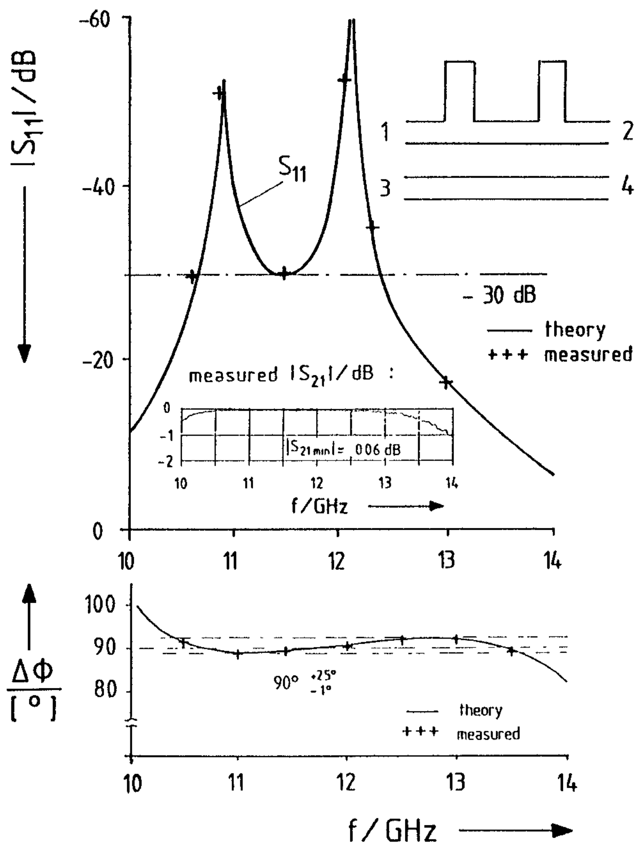

Fig. 3: Computer-optimized 90-degree two-stub phase shifter
Calculated and measured [9] input reflection coefficient. Waveguide R120-band (10 - 15 GHz, 19.050mm × 9.525mm).

CONCLUSION

Novel broad-band low-insertion loss 180-degree branch guide couplers are designed which integrate advantageously the characteristics of the multiple-branch coupling technique with the principle of E-plane stub loading. This leads to compact structures well appropriate for composed components, such as antenna-beamforming networks or printed E-plane components. The method of computer-optimization, which is based on field expansion into normalized incident and scattered waves, takes the higher-order mode coupling effects rigorously into account, and yields directly the overall scattering matrix. The theory is verified by measurements.

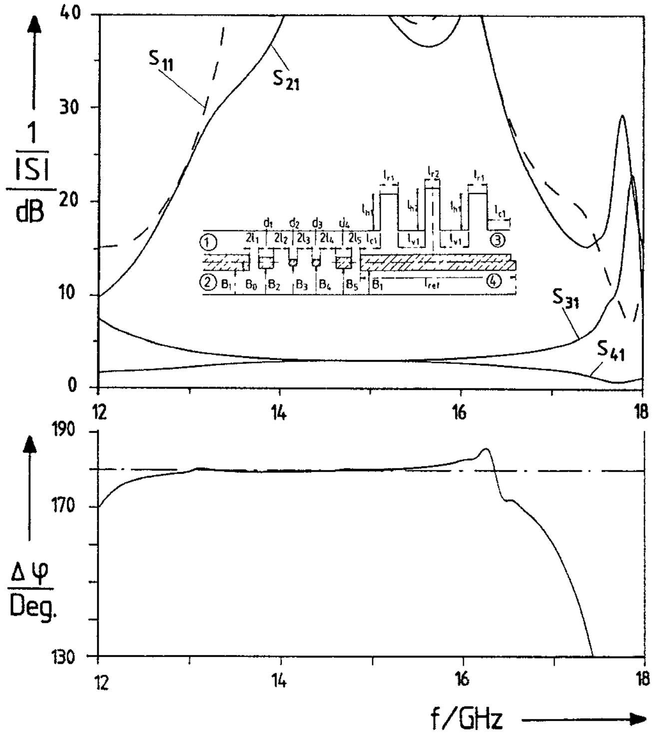

Fig. 4: Computer-optimized 180-degree 5-branch-3-stub coupler design example in the Ku-band

Scattering parameters as well as the differential phase shift between the output ports 4 and 3 (input port 1) as a function of frequency.

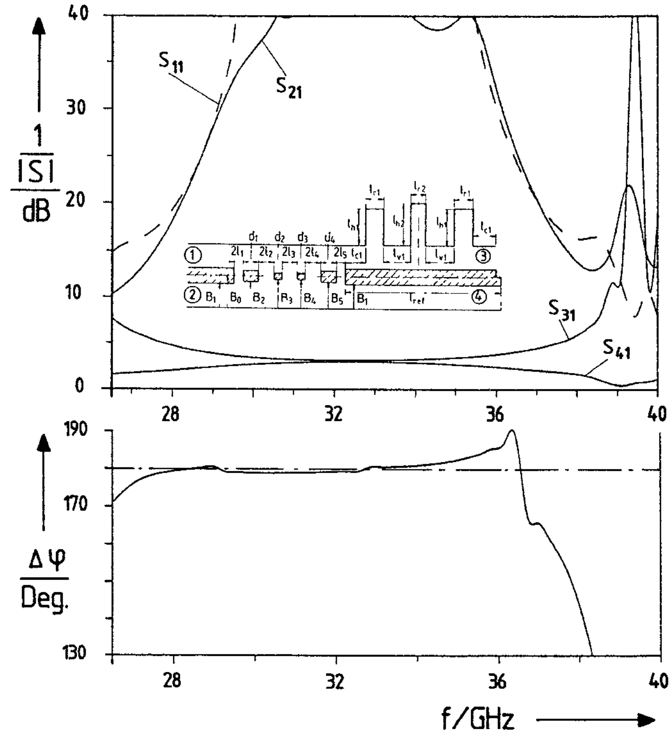

Fig 5: Computer-optimized 180-degree 5-branch-3-stub coupler design example in the Ka-band

Scattering parameters as well as the differential phase shift between the output ports 4 and 3 (input port 1) as a function of frequency (R320 waveguide housing, 7.112mm × 3.556mm).

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