COMPUTER AIDED DESIGN OF OPTIMUM E- OR H-PLANE N-FURCATED WAVEGUIDE POWER DIVIDERS

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ABSTRACT

Compact low-insertion-loss E- or H-plane n-furcated waveguide power dividers are designed with the method of field expansion into eigenmodes. As the finite septa thicknesses as well as the higher order mode interaction between all step discontinuities are rigorously taken into account, the full design potential including optimum stepped transformer configurations and their distance to the septa is utilized. Input reflection coefficients of better than -30dB between 26-40 GHz, or 29-38 GHz, are achieved by computer-optimized design examples with R320 waveguide E-plane bi-, or tri-furcation, respectively. The theory is verified by measurements. A very compact feed network example for a seven horn array demonstrates the usefulness of this design.

INTRODUCTION

Waveguide networks for satellite communication , [1]- [3], or radio link antennas [4] often require simple low-insertion-loss power dividers which are appropriate for compact topology and for uncomplicated high-precision fabrication techniques, [3], [4]. Although there are other types of waveguide elements capable of direct power splitting, e.g. E- and H-plane T-junctions [5], one of the advantages of the principle of septa in the E- or H-plane (Fig.1) is that it may achieve relatively broad-band power division with nearly identical phases in the n-furcated output ports. Moreover, low input VSWR for the feeding waveguide is simply provided by appropriate

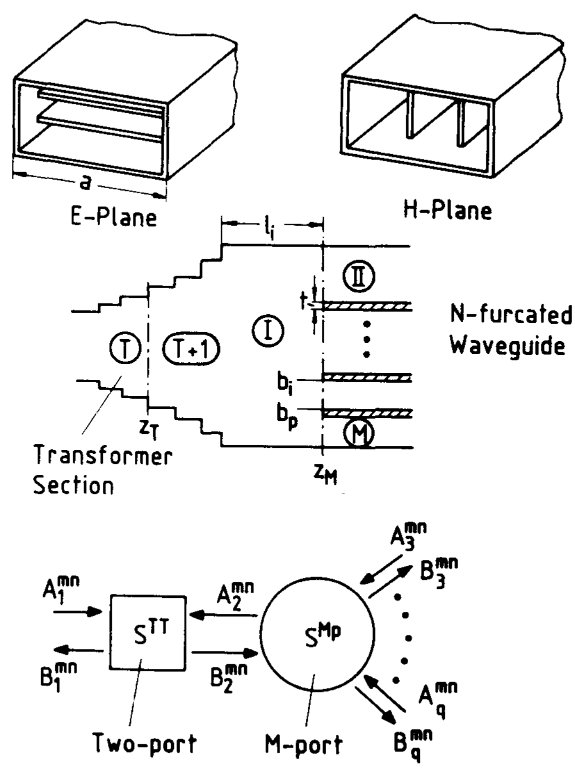

Fig. 1: E- or H-plane N-furcated power dividers with transformer sections

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transformer sections, whereby the E-Plane type is superior concerning the broad-band behaviour. A further advantage of the E-plane construction is that convenient milling or spark eroding techniques from a solid block allow one to produce compact waveguide channels of identical fundamental mode cutoff frequency. Such power dividers are well compatible with printed E-Plane components, such as filters and diplexers, which are of growing interest in recent millimeter-wave circuit design [6], [7].

Analysis methods based on equivalent network circuits for the bi-furcated waveguide are well-known [5]. The n-furcated H-plane waveguide with finite septa thicknesses has been solved more recently [8] with regard to applications for E-plane integrated metal insert filters with improved stopband behaviour. The n-furcated E-plane waveguide, and the combination of n-furcated H-or E-plane waveguides with appropriate transformer sections, however, have not yet been attacked rigorously with field theoretical methods. The purpose of this paper is to describe an appropriate computer-aided design method for such compensated n-furcated waveguide power dividers, to present some typical design examples, and to demonstrate the usefulness of the design theory by measurements and by a compact horn array feed network example using an optimized E-plane design.

THEORY

For the computer-aided design the modal expansion technique is applied [8]-[10]. The theory is presented only for the E-plane type, as the H-plane n-furcated waveguide is already treated in [8]. Moreover, the theory is concentrated on the new aspects, such as general E-plane n-furcation and generalized modal- S-matrix description of multiport structures.

A TE_{10}-wave incident in the input port of an E-plane waveguide discontinuity excites longitudinal section TE^{x}_{mn}-waves [9], [10]. For each typical homogeneous subregion, denoted with ν=T,T+1 for the two-port transformer section, and ν=I,II,...M for the M-port n-furcated waveguide section (cf. Fig. 1), respectively, the fields [9], [10]

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

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:

$$\Pi^{\nu}_{hx} = \sum_{m=1}^{ME} \sum_{n=0}^{NE} (A^{mn}_{\nu} - B^{mn}_{\nu}) \cdot C^{mn}_{\nu} \cdot \sin(\frac{m\pi x}{a}) \cdot \cos(\frac{n\pi}{b_i - b_p}(y-b_p)) \quad , \qquad (2)$$

where A, and B, denote the normalized forward and backward wave amplitude coefficients, respectively, a is the waveguide width, b_i and b_p are the relative waveguide heights within each waveguide section (cf. Fig. 1), C is a normalization factor due to the transported power, [8]-[10], and ME and NE are the number of expansion modes considered; the z-dependence $\exp(\mp jk^{mn}_{z\nu} \cdot z)$ in forward and backward direction is understood.

By matching the tangential field components at the common interfaces at the two typical step discontinuities, at $z = z_T$, and $z = z_M$ (Fig. 1), respectively, the wave amplitude coefficients of (2) can be related to each other after multiplication with the appropriate orthogonal function, [8]-[10]. This yields the key building block two-port modal scattering matrix (S^{Tp}) for one transformer section step, and the key building block M-port modal scattering matrix (S^{Mp}) for the n-furcated waveguide

$$\begin{bmatrix} (B_T) \\ \\ (B_{T+1}) \end{bmatrix} = (S^{Tp}) \begin{bmatrix} (A_T) \\ \\ (A_{T+1}) \end{bmatrix} \quad ; \quad \begin{bmatrix} (B_I) \\ \vdots \\ (B_M) \end{bmatrix} = (S^{Mp}) \begin{bmatrix} (A_I) \\ \vdots \\ (A_M) \end{bmatrix} \quad . \qquad (3)$$

In order to preserve numerical accuracy, the direct combination of the involved scattering matrices at all step discontinuities and of the intermediate homogeneous waveguide sections is used. For a two-port series connection of the modal scattering matrices, the formulas, which may be used iteratively, are already given in [8], [9]. The connection of the total transformer section two-port modal scattering matrix (S^{TT}), including the intermediate homogeneous waveguide section of length l_i (cf. Fig. 1), with the the M-port modal scattering matrix (S^{Mp}), is found analogeously, since the direct scattering matrix formulation does not require any symmetry of ports or modes in contrast to the usual transmission matrix treatment. The relations for calculating the scattered wave vectors B of the mn modes considered at the ports 1, s, (s=3,4,...,q), cf. Fig. 1, are given by

$$B_1 = [S_{11}^{TT} + S_{12}^{TT} \cdot W \cdot S_{11}^{Mp} \cdot S_{21}^{TT}] \cdot A_1 + S_{12}^{TT} \cdot W \cdot K$$
$$B_s = S_{s-1,1}^{Mp} \cdot [S_{21}^{TT} + S_{22}^{TT} \cdot W \cdot S_{11}^{Mp} \cdot S_{21}^{TT}] \cdot A_1 + S_{s-1,1}^{Mp} \cdot S_{22}^{TT} \cdot W \cdot K + P, \qquad (4)$$

with $W=(U-S_{11}^{M} S_{22}^{TT})^{-1}$, $K=\sum_{r=2}^{q-1} S_{1,r}^{Mp} \cdot A_{r+1}$, $P=\sum_{p=2}^{q-1} S_{s-1,p}^{Mp} \cdot A_{p+1}$, and where the indices denote the related submatrices.

A computer program was written using the preceding relations and utilizing the evolution strategy method, cf. [8], [9], for optimizing the geometrical parameters of the power dividers for given specifications. Sufficient asymptotic behaviour has been obtained by consideration of m=1 n=18 TE_{mn}^{x}-modes for the E-plane type, and m=15 TE_{m0}-modes for the H-plane type.

RESULTS

E-plane bi-furcated power divider examples are presented in Figs. 2. A very broad-band design with standard R320 waveguide input and

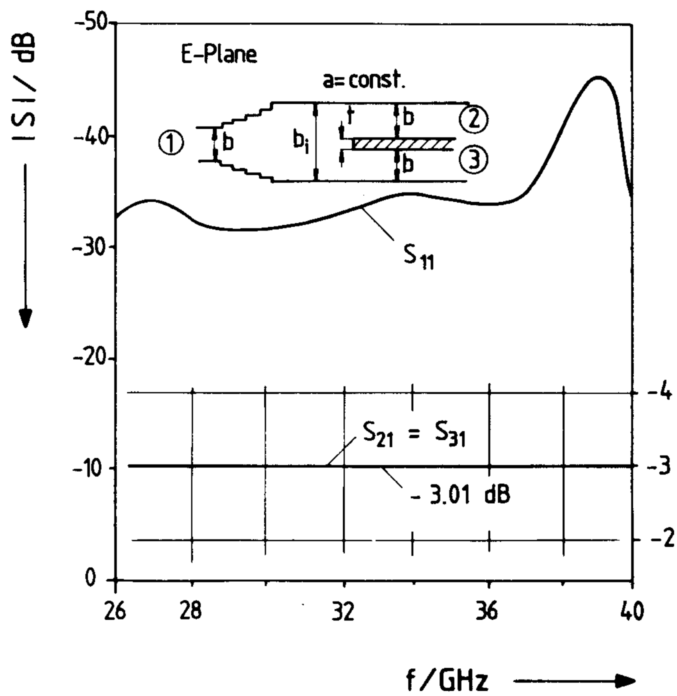

Fig. 2a: E-plane bi-furcation power divider with five-step transformer, R320 waveguide. Broad-band design

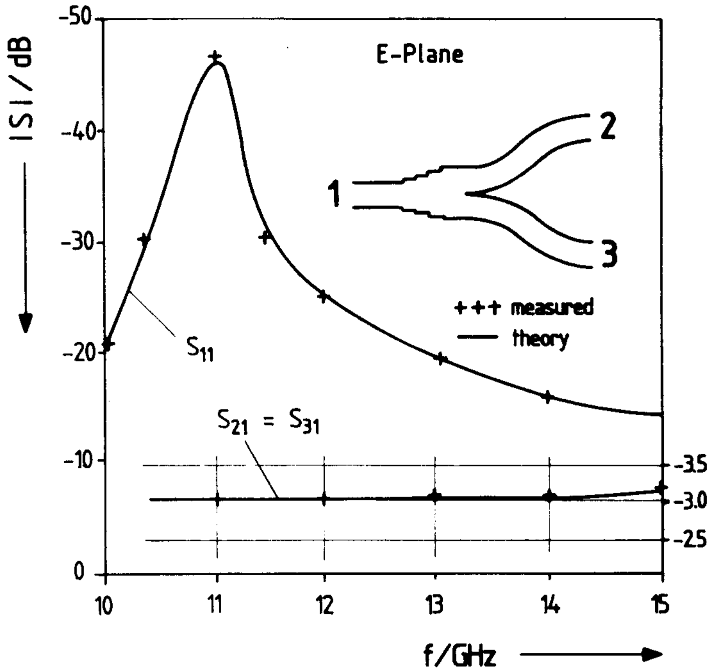

Fig. 2b: E-plane bi-furcation power divider with five-step transformer, R120 waveguide. Narrow-band design. (Measured results by courtesy of Antenna Dpt. MBB/Erno, Munich, W. Germany)

output port dimensions achieves less than -30dB input reflection between 26-40 GHz (Fig. 2a). A narrow-band design for about 11 GHz with R120 waveguide dimensions has been fabricated in the antenna departmant of MBB/Erno, Munich, by utilizing a computer-aided milling technique. The very good agreement between theory and measurements (Fig. 2b) shows that the design theory presented enables the direct hight precision manufacturing without the necessity for additional 'trial-and-error' adjustment methods. An H-plane bi-furcated design example (Fig. 3) demonstrates that the maximum input reflection is considerably lower, although a more complicated impedance transformer section profile was chosen.

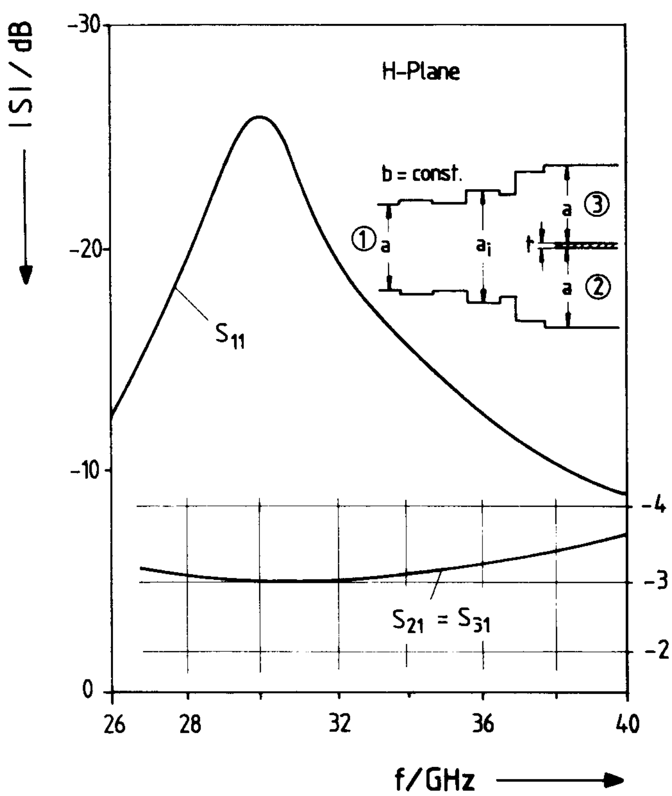

Fig. 3: H-plane bi-furcation power divider, R320 waveguide

An E-plane tri-furcated power divider example for R320 waveguides is presented in Fig. 4. The input reflection coefficient is better than -30dB between 29-38 GHz, together with (+0.2/-0.1)dB variation of the transmission coefficient to the output ports between 29-35 GHz, and (+0.5/-0.2)dB between 35-38 GHz.

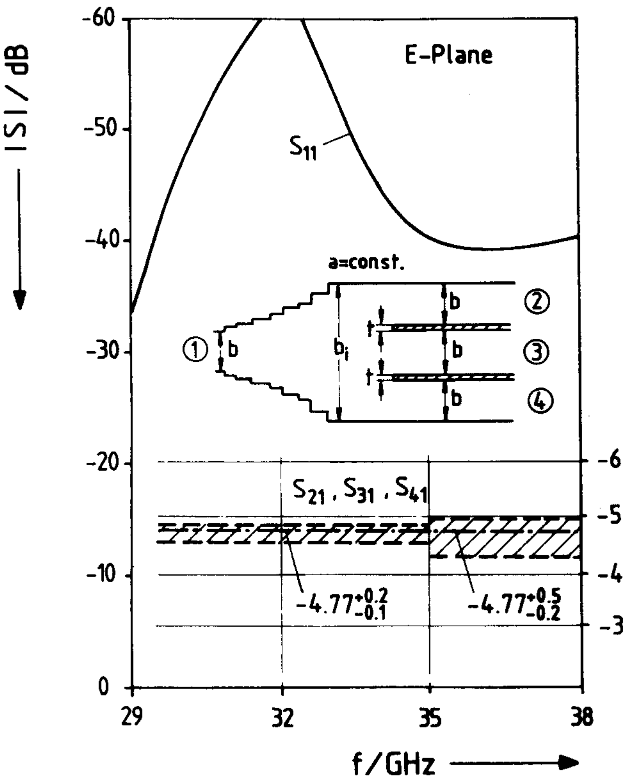

Fig. 4: E-plane tri-furcation power divider

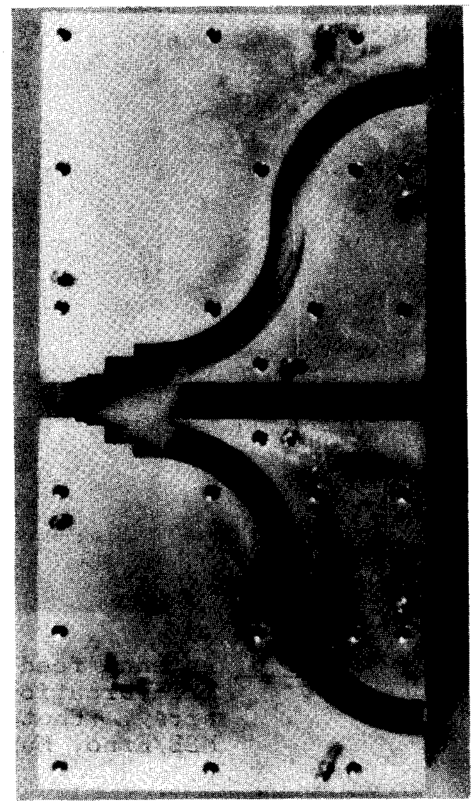

Fig. 5: E-plane tri-furcation power divider milled from a solid block, R120 waveguide. (Photograph by courtesy of Antenna Dpt. MBB/Erno, Munich, W. Germany)

Fig. 5 shows the photograph of an E-plane -4.77dB power divider example for R120 input and output waveguides, which was milled from a solid block to produce waveguide channels of identical a-dimension milling depth, and was spark eroded to size in the more critical areas of corners and steps. A top plate may be fitted to complete the assembly. The computer-aided design presented and the application of the appropriate fabrication technique may achieve very compact feed network assemblies, as is demonstrated in Fig. 6 at the example of a seven element horn array feeeding network for about 11 GHz. The overall feed network package and the multi-beam feed array have been fabricated in the antenna department of MBB/Erno, Munich.

Acknowledgment

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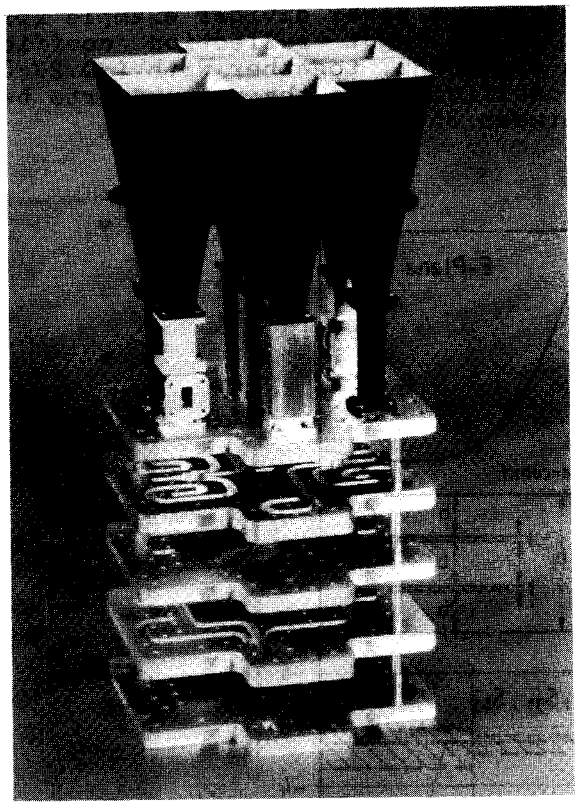

Fig. 6: Antenna feed network package utilizing E-plane tri-furcation power dividers. Seven element horn array. (Photograph by courtesy of Antenna Dpt. MBB/Erno, Munich, W. Germany)

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