COMPUTER-AIDED DESIGN OF PARALLEL-CONNECTED MILLIMETER-WAVE DIPLEXERS/MULTIPLEXERS

by

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ABSTRACT

This paper describes the analysis and design of a novel integrated millimeter wave diplexer. The structure is simple and can easily be extended to a multiplexer configuration. The diplexer is composed of ladder-shaped E-plane metal insert filters which are fabricated on a single metallic sheet and embedded in a split block housing.The theoretical design procedure is based on the generalized scattering matrix method which includes mutual parasitic loading effects between the filters as well as higher order mode interaction. Thus, no physical fine tuning of the component is necessary.

Introduction

Conventional design methods for diplexers/multiplexers follow in principle a two step procedure: First, the design of the band-select filters and second, physical fine tuning of the individual filter positions in the channels relative to their common input port. This procedure is time consuming and particulary difficult at millimeter wave frequencies. Furthermore, T-junctions are commonly used to connect the individual channels with the main input waveguide (i.e. [1], [7]). T-junctions, however, are typically narrow bandwidth devices and it is difficult to compensate for the rapid change of the junction reactance over frequency. Thus, this scheme is not suitable if wide-band channels are required. To alleviate this problem a low-cost W-band diplexer design has been published in [3] which utilizes a wide-band printed probe transition from waveguide to suspended-stripline. E-plane filters were printed on the same substrate to select the frequency bands and computer-aided design methods were used to account for the loading effect of the filters to ensure a good overall response.

However, the application range of this structure is limited to low-power signals and it is difficult to extend this structure with the same simplicity to a multiplexer network.

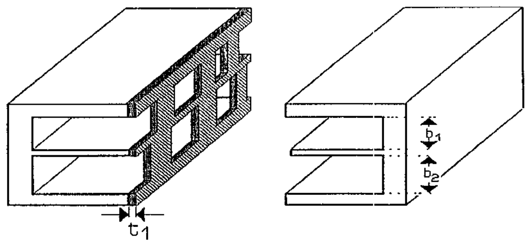

Fig.1a Perspective view of the diplexer arrangement with E-plane filters.

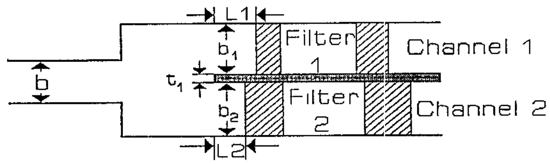

Fig.1b Side view of the diplexer connected to the input port by an abrupt E-plane step transition

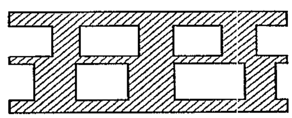

Fig.1c Dual ladder-shaped E-plane metal insert

Therefore, the present paper describes a new low-cost design for millimeter wave diplexers which can easily be extended to multiplexer applications. The principle structure was suggested in [5]. However, the design was based on equivalent network theory and did not include the transition between the standard waveguide and the power divider section to feed the individual channels.

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Diplexer Design

Fig.1 shows the principle configuration of the diplexer. The structure consists of an E-plane bifurcation and two E-plane metal insert filters. The advantage of this design approach is obvious. Firstly, in comparison to the E-or H-plane T-junction the E-plane bifurcation (or n-furcation for multiplexers) is a rather broadband transition for the TE10-mode. Secondly, since the output waveguides have common broadwalls, the ladder-shaped E-plane metal insert filters can be fabricated on a single metallic sheet. Thus, their positions in the channels with respect to the common port are automatically adjusted in a single manufacturing step. Finally, the present solution is not only suitable for high power signals but also provides extremely low insertion loss when needed due to the absence of a lossy substrate material in the filter structure [2].

To compensate for the mutual loading effect between the filters the lengths L1, L2 (Fig.1b) must be optimized as well as L to ensure a good match between the input waveguide and the output channels. Instead of using an abrupt step transition as shown in Fig.1, a tapered transition improves the return loss of the power divider up to an average of 30dB over 20 GHz in W-band. The optimum dimensions for the taper as well as for L1, L2 and L have been obtained by computer optimization which is an indispensible part of the design procedure. On the other hand the best optimization routine is useless if the analysis method is not accurate.

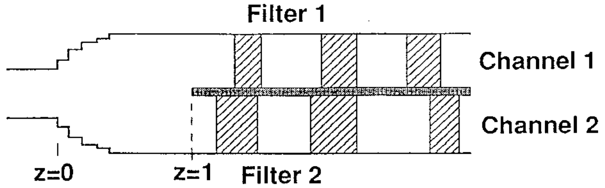


Fig.2a Diplexer connected to the input port by a tapered E-plane waveguide transition

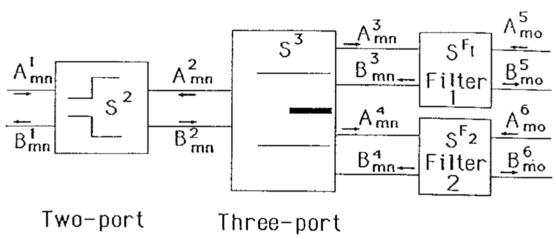


Fig.2b S-matrix representation of the various discontinuities involved in the structure of Fig.2a.

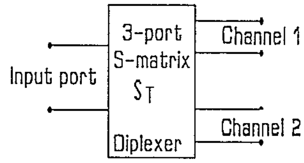


Fig.2c Overall 3-port S-matrix of the diplexer.

Therefore, we used a rigorous mode matching approach to calculate the generalized scattering matrix of the entire component and then we optimized the dimensional parameters for best circuit performance. The individual steps are outlined in Fig.2: First, we find the optimum dimensions for the E-plane band-select filters and the corresponding s-matrices. This can be done using the method described in [2]. Second we optimize the transformer section and combine its s-matrix with that of the E-plane bifurcation. Finally, the resulting three-port s-matrix is combined with the individual filter s-matrices, and the parameters L1, L2 and L are optimized to yield the best overall diplexer response. The same procedure applies to the design of multiplexers.

Analysis

For the mode matching method the structure is decomposed into two classes of discontinuities:

1) The E-plane bifurcation which includes the abrupt step in waveguide heigth (Fig.1).

2) The H-plane bifurcation in the filter sections.

The H-plane bifurcation was treated in detail in [2] in conjunction with E-plane metal insert filters. E-and H-plane n-furcations have in common that the field components at the discontinuities can be derived from the x component of the magnetic Hertz potential. However, instead of only three field components (Ey, Hx,Hz) at the H-plane discontinuity, an incident TE10-mode will excite five field components (Ey,Ez,Hy,Hx,Hz) at the E-plane discontinuity. The generalized scattering matrix in this case is then derived from the matching condition for Ey and Hy or Ey and Hx field components.

$$Ey^{\nu} = j\omega\mu \sum_{n=0}^{N}\sum_{m=1}^{M} C^{\nu}_{nm} \sin(kx^{\nu}_{m}x)\cos(ky^{\nu}_{n}y) \cdot (A^{\nu}_{nm} + B^{\nu}_{nm}) e^{\mp jkz^{\nu}_{nm}z} \quad (1)$$

$$Hx^\nu = \sum_{n=0}^{N}\sum_{m-1}^{M}(ko^2-(kz^\nu_{nm})^2)C^\nu_{nm}\sin(kx^\nu_m x) \cdot$$

$$\cdot \cos(ky^\nu_n y)(A^\nu_{nm}+B^\nu_{nm})e^{\mp jk^\nu_{nm}z} \qquad (2)$$

The abbreviations Amn and Bmn in (1) and (2) denote the incident and reflected wave amplitudes and Cmn is a factor to normalize the wave amplitudes to 1 W.

From the matching conditions at the two principle interface planes, z=0 and z=1 (Fig.2a), one obtains the two-port scattering matrix for the abrupt step transition

$$\begin{bmatrix} B^1_{mn} \\ B^2_{mn} \end{bmatrix} = S^2 \begin{bmatrix} A^1_{mn} \\ A^2_{mn} \end{bmatrix} \qquad (3)$$

and the three-port s-matrix for the E-plane bifurcation

$$\begin{bmatrix} A^2_{mn} \\ A^3_{mn} \\ A^4_{mn} \end{bmatrix} = S^3 \begin{bmatrix} B^2_{mn} \\ B^3_{mn} \\ B^4_{mn} \end{bmatrix} \qquad (4)$$

The algorithm to cascade two-port s-matrices necessary in the taper section has been given in [7]. Cascading the three-port s-matrix with the filter two-ports in each channel yields the following algorithm: At first we combine the three-port with the two-port of the filter in channel 1

$$\begin{aligned}
S^0_{11} &= S^3_{11} + S^3_{12} * S^{F_1}_{11} * W * S^3_{21} \\
S^0_{21} &= S^{F_1}_{21} * W * S^3_{21} \\
S^0_{31} &= S^3_{31} + S^3_{32} * S^{F_1}_{11} * W * S^3_{21} \\
S^0_{12} &= \left[S^3_{12} * S^{F_1}_{11} * W * S^3_{22} + S^3_{12}\right] * S^{F_1}_{21} \\
S^0_{13} &= S^3_{13} + S^3_{12} * S^{F_1}_{11} * W * S^3_{23} \\
S^0_{22} &= S^{F_1}_{22} + S^{F_1}_{21} * W * S^3_{22} * S^{F_1}_{12} \\
S^0_{23} &= S^{F_1}_{21} * W * S^3_{23} \\
S^0_{32} &= \left[S^3_{32} * S^{F_1}_{11} * W * S^3_{22} + S^3_{32}\right] * S^{F_1}_{12} \\
S^0_{33} &= S^3_{33} + S^3_{32} * S^{F_1}_{11} * W * S^3_{23} \\
W^0 &= \left[E - S^3_{22} * S^{F_1}_{11}\right]^{-1}
\end{aligned}$$

In a second step the overall s-matrix of the diplexer is obtained by combining S with the filter in the second channel using the same algorithm again but replacing S^3_{nm} by S^0_{nm} and S^{F1}_{nm} by S^{F2}_{nm}. For multiplexers the procedure must be repeated according to the number of channels involved.

There are only TE$^x_{1m}$ -modes in the vicinity of an E-plane discontinuity. Nevertheless, we consider in equ.'s (1) and (2) the more general case of TE$^x_{nm}$ -modes. This is so because the s-matrix of the E-plane bifurcation must be cascaded with the s-matrix of the channel filters which contain only TE$^x_{no}$ -modes. To cascade both types of matrices we must consider only those scattering elements in S (Fig.2b) which correspond to B1o, B3o, B5o (etc.) wave amplitudes. This results in diagonal matrices for S11, S12 etc. Since TE$^x_{no}$ -modes of higher order than 2 to 3 (odd modes) are negligible a short distance from the first filter discontinuity, the size of the matrices to be cascaded is considerably reduced accelerating the computer program significantly. The E-plane discontinuities are sufficiently described by taking into account 8 to 12 even higher order TE$^x_{1m}$ -modes. This is shown for the E-plane power divider and the five step tapered transition in Fig.3.

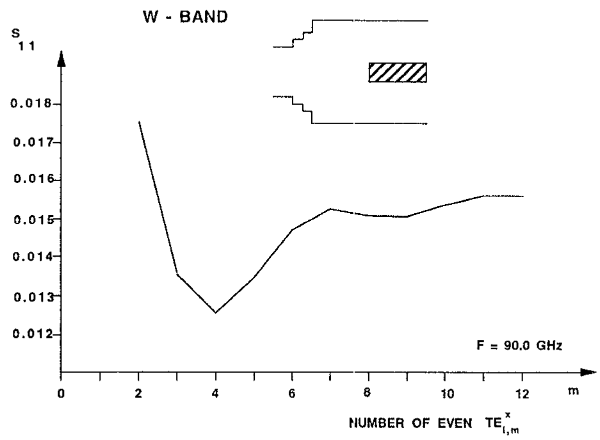


Fig.3 Convergence analysis of a five step E-plane taper with bifucation in W-band(a=2.54,b/a=0.5) d1=0.0195, d2=0.109, d3=0.288,d4=0.538,d5=0.68,s1=1.24,s2=1.174, s3=1.083,s4=1.135 L=1.82 (dimensions in mm)

Results

Fig.4 shows the frequency response of the same taper over a 20 GHz bandwidth in W-band. The tapered transition was necessary since the abrupt step transition between the input standard waveguide and the bifurcated power divider section was rather large (2b). The corresponding return loss was below 10dB. With the optimized taper the return loss remains above 30dB in the average.

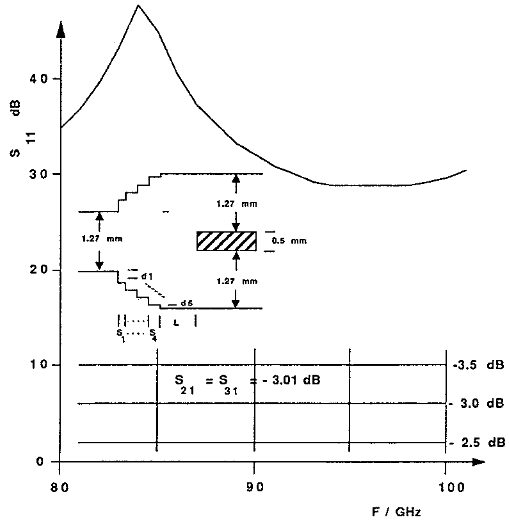


Fig.4 Frequency response of the tapered power divider given in Fig.3

Fig.5 shows the response of noncontiguous diplexers in Ka- and W-band with a guard band between 29 and 32 GHz and 94.5 and 95.5 GHz, respectively. Both diplexers are fed by tapered transitions. The predesigned bandpass filters show excellent performance as stand-alone components. In

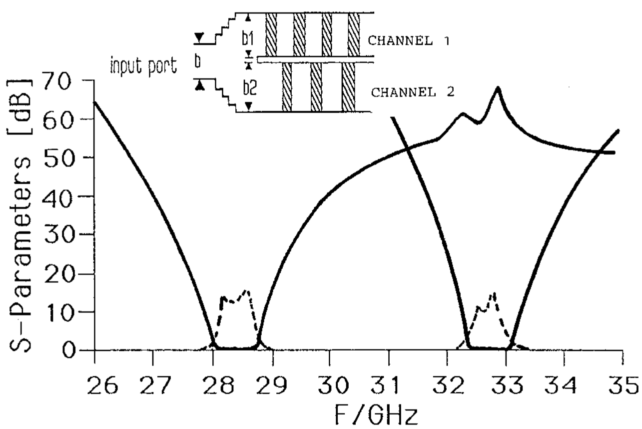
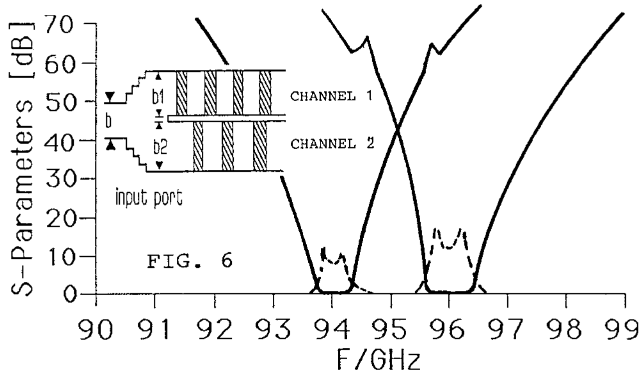


Fig.5 Frequency response of a noncontiguous Ka-band diplexer. Filter 1 resonators, 1.: 6.311mm; 2.: 6.397mm; coupling sections, 1.: 0.525mm; 2.: 2.539mm; Filter 2 resonators, 1.:4.372mm; 2.:4.4mm; coupling sections, 1.:0.647mm; 2.:3.446mm; 3.:4.056mm. Both filters are symmetrical about resonator 2 and coupling section 3, respectively. Metallization thickness t=127.um

Fig.6 Frequency response of a noncontiguous W-band diplexer. Filter 1 resonators, 1.:1.471mm; 2.:1.473mm; coupling sections 1.:0.586mm; 2.:1.829mm; 3.:1.98mm. Filter 2 resonators, 1.:1.555mm 2.:1.557mm; coupling sections, 1.:0.591mm; 2.:1.829; 3.:1.98mm. Both filters are symmetrical about 3.coupling section. t=127.um



the diplexer arrangement, however, their passband insertion loss increases somewhat (0.3 dB) and the return loss deteriorates down to 15 dB. The typical CPU time required to optimize a diplexer is approximately 30 min. on a PC (AT) with DSI20 coprocessor.

Conclusion

A parallel-connected diplexer configuration has been introduced which can easily be extended to a multiplexer component. Fabrication costs are reduced to a minimum due to the applicability of low-cost photolithographic techniques and the use of accurate CAD software, which makes fine tuning unnecessary. Furthermore, due to a computer time efficient but accurate numerical procedure the design software is operational on Personal Computers.

References

[1] Matthaei,G., L.Young and E.M.T. Jones,"Microwave Filters, Impedance-Matching Networks, and Coupling Structures", Artech House, Inc. 1980.

[2] Vahldieck, R., J. Bornemann, F. Arndt and D. Grauerholz," Optimized Waveguide E-Plane Metal Insert Filters for Millimeter Wave Applications", IEEE Trans., MTT-31, pp.65-69, January 1983.

[3] Shih, Y.-C., L.Q. Bui and T. Itoh, "Millimeter-Wave Diplexers with Printed Circuit Elements", IEEE Trans. MTT-33, pp.1465-1469, December 1985.

[4] Dittloff,J.,J.Bornemann and F.Arndt, "Computer Aided Design of Optimum E- or H-plane N-Furcated Waveguide Power Divider", Conference Proceedings 17th European Microwave Conference,1987, pp. 181-186.

[5] Reiter, G. and T. Kolumban," Diplexer Arrangement utilizing E-plane Filters", 15Proc. 15th Europ. Microwave Conference, pp.859-864, 1985.

[6] Marshall, T., J. Christopher and S. Kluger,"The Definicon 68020 Coprocessor", Part 1 and 2 published in Byte Magazin July and August 1986, pp.120-144 and pp.108-114.

[7] Vahldieck, R. and W.J.R. Hoefer,"Finline and Metal Insert Filters with improved Passband Separation and Increased Stopband Attenuation", IEEE Trans. Microwave Theory Tech., vol. MTT-33, No.12, Decemb. 1985

[8] Arndt, F., J. BOrnemann,D. Grauerholz," Waveguide E-plane Integrated-Circuit Diplexer", Electronics Letters, 4th July, 1985, Vol.21, No.14, pp.615-617