Design of Sum-Difference Power Combiners with Second-Order Filtering Functions

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***Abstract*—A design approach for sum and difference power combiners with filtering capabilities is introduced. It is based on a symmetric configuration that can be straightforwardly analyzed using ABCD matrices of the even and odd networks. The resulting four-port scattering matrix is compared to the respective transmission functions of a second-order filter which relates prototype coefficients to inverter values. The method is demonstrated at the example of a 23.9 GHz substrate integrated waveguide (SIW) combiner with sum and difference ports. Good agreement between results obtained from CST and a mode-matching technique validates the design approach.**

Keywords*—*filter theory, power combiner, substrate integrated waveguide, waveguide filters

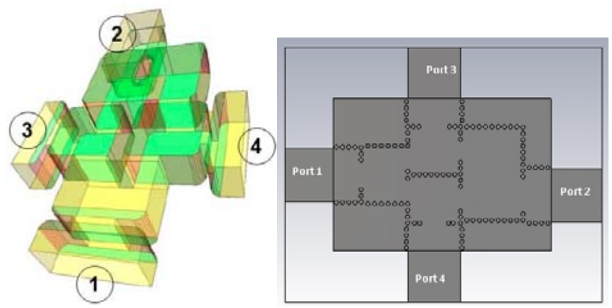
I. INTRODUCTION

The demand for ever smaller systems with increasing capabilities leaves the designing engineer with two basic options: First, to miniaturize individual components and subsystems and, second, combine functions of individual components. For feed systems in communication equipment, the first option usually implies a reduction in power handling capabilities, thus the second option is often employed.

Among the many individual components within feed systems, the combination of power dividers/combiners (or couplers) with filters appears to be a straightforward option. Therefore, a number of filtering dividers in different technologies have been proposed. In the lower GHz frequency range, microstrip [1] – [5] and LTCC [6] technologies are the obvious choice. Due to the increasing loss of microstrip circuitry with frequency, power dividers with filtering options operating in X-band have been designed using dielectric resonators within metal housings [7], conventional waveguide circuits [8], [9] or substrate integrated waveguide (SIW) components in single-layer [9] – [11] or dual-layer [12] arrangements.

This paper focuses on the design of power combiners with sum and difference ports and second-order filtering functions. Assuming a network with a symmetry plane, the four-port scattering parameters can be directly related to second-order filter prototypes by using ABCD matrices of the respective even- and odd-mode circuits.

In a pure waveguide environment, the symmetry is preserved if the difference port extends towards the E-plane while all other ports are situated in the H-plane. Such a structure is shown in Fig. 1a. For a SIW design, all ports must be located in the H-plane and, therefore, the location of port 2 in Fig. 1a must be moved to one side of a TE_{102}–mode resonator to provide the appropriate phase requirement for the difference port (Fig. 1b). However, that renders the circuit asymmetric, and optimization is employed to finalize the design. This is demonstrated at the example of a SIW combiner for K-band applications at 23.9 GHz.



(a) (b)

Fig. 1. Sum-difference power combiners with second-order filtering functions in waveguide (a) and SIW (b) technologies.

II. THEORY

If, for the circuits shown in Fig. 1, we assume that ports 3 and 4 are input ports, then those signals will be combined in phase at port 1 after each passing a second-order filter. Port 2 will receive the difference of the signals due to the iris couplings of the signals between the TE_{101} mode resonators and the TE_{102}–mode resonator with opposite phase (Fig. 1).

Taking the example of Fig. 1a, the circuit can be represented by the coupling scheme of Fig. 2 with initially different input/output (J_1, J_3) and coupling (J_{14}, J_{32}) inverters. Using ABCD matrices for the respective even and odd modes, we obtain for the even mode

$$\begin{pmatrix} A_e & B_e \\ C_e & D_e \end{pmatrix} = \begin{pmatrix} 0 & \dfrac{j}{J_1} \\ jJ_1 & 0 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ s + jB_1 + jJ_{14} & 1 \end{pmatrix} \begin{pmatrix} 0 & \dfrac{j}{J_{13}} \\ jJ_{13} & 0 \end{pmatrix} \times \begin{pmatrix} 1 & 0 \\ s + jB_3 + jJ_{32} & 1 \end{pmatrix} \begin{pmatrix} 0 & \dfrac{j}{J_3} \\ jJ_3 & 0 \end{pmatrix} \quad (1)$$

and for the odd mode

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$$\begin{pmatrix} A_o & B_o \\ C_o & D_o \end{pmatrix} = \begin{pmatrix} 0 & \dfrac{j}{J_1} \\ jJ_1 & 0 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ s + jB_1 - jJ_{14} & 1 \end{pmatrix} \begin{pmatrix} 0 & \dfrac{j}{J_{13}} \\ jJ_{13} & 0 \end{pmatrix} \times \begin{pmatrix} 1 & 0 \\ s + jB_3 - jJ_{32} & 1 \end{pmatrix} \begin{pmatrix} 0 & \dfrac{j}{J_2} \\ jJ_2 & 0 \end{pmatrix} \quad (2)$$

with s being the complex frequency.

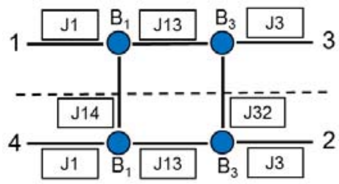

Fig. 2. Equivalent circuit of power combiner in Fig. 1a.

By converting $ABCD_{e,o}$ to scattering parameters $S_{e,o}$, the overall four-port scattering matrix of the sum-difference power combiner is obtained as

$$S = \begin{pmatrix} \dfrac{S_{11}^e + S_{11}^o}{2} & \dfrac{S_{13}^e + S_{13}^o}{2} & \dfrac{S_{13}^e - S_{13}^o}{2} & \dfrac{S_{11}^e - S_{11}^o}{2} \\ \dfrac{S_{13}^e + S_{13}^o}{2} & \dfrac{S_{33}^e + S_{33}^o}{2} & \dfrac{S_{11}^e - S_{11}^o}{2} & \dfrac{S_{13}^e - S_{13}^o}{2} \\ \dfrac{S_{13}^e - S_{13}^o}{2} & \dfrac{S_{11}^e - S_{11}^o}{2} & \dfrac{S_{33}^e - S_{33}^o}{2} & \dfrac{S_{13}^e + S_{13}^o}{2} \\ \dfrac{S_{11}^e - S_{11}^o}{2} & \dfrac{S_{13}^e - S_{13}^o}{2} & \dfrac{S_{13}^e + S_{13}^o}{2} & \dfrac{S_{11}^e - S_{11}^o}{2} \end{pmatrix} \quad (3)$$

The denominators of the scattering parameters of the even and odd modes become identical when the symmetry relations $J_1 = J_3$, $B_1 = B_3$ and $J_{32} = -J_{14}$ are enforced. Each entry in the scattering matrix is then reduced to a second-order response. Note that the last condition imposes the phase reversal required to obtain the difference of port 3 and 4 signals at port 2. With these restrictions, the scattering parameters are obtained as

$$S_{11} = S_{22} = S_{33} = S_{44} = \frac{s^2 + 2jsB_1 - B_1^2 + J_{14}^2 + J_{13}^2 - J_1^4}{s^2 + 2s(jB_1 + J_1^2) + 2jJ_{13}^2 B_1 - B_1^2 + J_{14}^2 + J_{13}^2 + J_1^4} \quad (4)$$

$$S_{13} = S_{31} = S_{24} = S_{42} = \frac{2jJ_1^2 J_{13}}{s^2 + 2s(jB_1 + J_1^2) + 2jJ_{13}^2 B_1 - B_1^2 + J_{14}^2 + J_{13}^2 + J_1^4} \quad (5)$$

$$S_{14} = S_{41} = -S_{23} = -S_{32} = \frac{2jJ_1^2 J_{14}}{s^2 + 2s(jB_1 + J_1^2) + 2jJ_{13}^2 B_1 - B_1^2 + J_{14}^2 + J_{13}^2 + J_1^4} \quad (6)$$

$$S_{12} = S_{21} = S_{34} = S_{43} = 0 \quad (7)$$

A direct comparison of the polynomials in s with those of a second-order filter function will yield the values of the coupling coefficients (inverters). Once they are determined, the design of the basic building block of Fig. 2 follows standard techniques of coupled-resonator filters, e.g. [13]. As an example, Fig. 3 shows the response of a two-resonator block for a midband frequency of 23.9 GHz and 22 dB return loss.

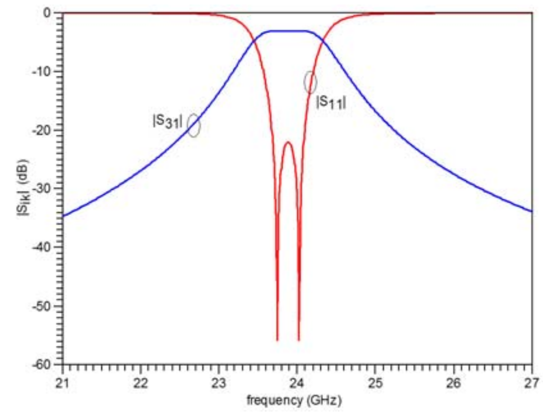

Fig. 3. Response of the basic building block at 23.9 GHz with 22 dB return loss.

III. Results

A single-plane symmetric four-port all-metal waveguide circuit that satisfies the conditions outlined in Section II has been designed and prototyped for a bandwidth of 150 MHz, centered at 11 GHz with a return loss of 26 dB. Very good agreement between theory and measurement is observed. The reader is referred to [9] for details.

Here the design in Fig. 3 is performed in SIW technology (inset of Fig. 4) using a combination of a mode-matching code (MMT) and CST Microwave Studio. The substrate is RT/duroid 6002 with ε_r=2.94 and h=508 μm, all via hole diameters are 1/64" (0.397 mm), the via pitch at the ports is 0.6 mm, and the via center-to-center port width is 5.4 mm with an equivalent waveguide width of 5.096 mm [14]. Since the symmetry assumed in Section II is violated, port 2 (Fig. 4) is initially centered (symmetric) and then moved downwards to obtain the difference output port. With port 2 at that location, a full optimization of all apertures and resonator lengths is required to achieve the desired behavior of the circuit.

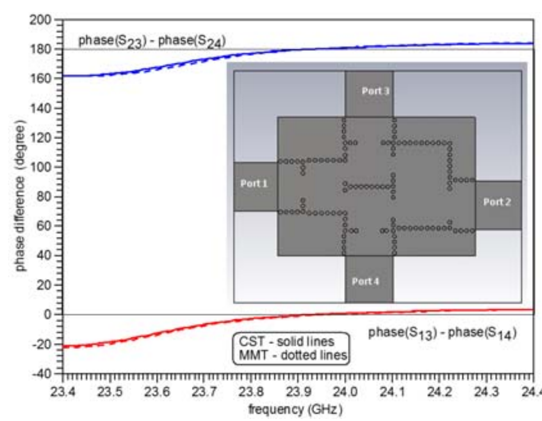

Fig. 4. Phase differences between ports 3 and 4 to port 1 (sum) and port 2 (difference).

Some of the most important design aspects are the phase relationships. Fig. 4 demonstrates that the 0° and 180° sum and difference conditions, respectively, are well met within the passband of the combiner.

All scattering parameters are displayed in Fig. 5. Good agreement between CST and MMT results is observed. The return loss is better than 22 dB and isolation values are better than 20 dB.

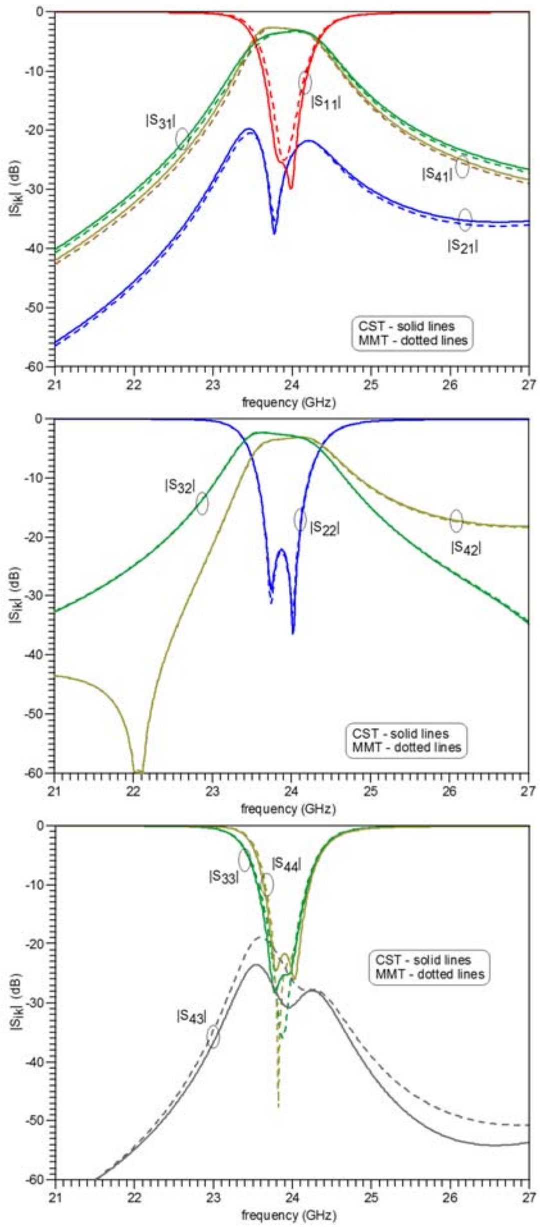

Fig. 5. Performance comparison between CST and MMT for the 23.9 GHz sum-difference SIW power combiner.

IV. Conclusions

The symmetric equivalent circuit permits the design of sum-difference combiners with filtering capabilities. Through a direct comparison between the derived scattering matrix and scattering parameters of second-order filter functions, the related filter inverter values are obtained. Implementation in rectangular waveguide technology maintains structural symmetry due to the fact that the difference port extends towards the third dimension. In a planar SIW circuit, this port has to be connected to a TE_{102}–mode resonator and the entire structure has to be optimized due to violation of symmetry. The method is demonstrated for a 23.9 GHz SIW sum-difference combiner with 22 dB return loss. Good agreement between CST and MMT validates the design approach. The basic building block can be extended to achieve higher filter degrees and/or to provide multi-port combiner networks.

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