# ASYMMETRIC MICROSTRIP DC BLOCKS WITH RIPPLED RESPONSE

D. Kajfez, S. Bokka, C. E. Smith

Department of Electrical Engineering
University of Mississippi
University, MS 38677

## ABSTRACT

A design procedure is described for dc blocks using asymmetric coupled microstrip which exhibit a rippled frequency response, while at the same time acting as impedance transformers. Lumped capacitances are added in order to compensate for the difference in propagation constants of two quasi TEM modes.

## Introduction

Microstrip dc blocks[1] are convenient building components because they permit the separation of the bias voltages while acting as bandpass filters for microwave signals. It would be convenient to design the microstrip dc blocks so that they also transform impedances, because they could then be used e.g. as output matching networks of typical microwave bipolar transistors.

Figure 1 shows a microstrip dc block, with the source and load attached. A symmetric dc block is the one for which the conductor widths are equal to each other ($w_1=w_2$), while an asymmetric dc block has unequal conductor widths ($w_1 \neq w_2$). When the source and load impedances are equal to each other ($R_S=R_L$), the symmetric dc block may be designed to produce a prescribed rippled response within the prescribed bandwidth[2]. Furthermore, it is possible to design the symmetric dc block for a prescribed ripple with slightly different source and load impedances[3]. The symmetric dc block can even be utilized for large impedance transformation ratios[4], if one is satisfied with a narrow-band, single-peaked response of the device.

This paper describes the procedure for designing the asymmetric dc blocks with rippled response, which also act as impedance transformers.

## Section of the asymmetric coupled microstrip

The two modes which may propagate on a coupled microstrip transmission line are neither odd nor even[5]. The character of the two modes is best described in terms of modal voltage eignevectors $|e_n\rangle$ and current eigenvectors $|i_n\rangle$, for the modes n=1 and n=2[5,6]. The voltage and current eigenvectors are related through the characteristic impedance matrix $\underset{\sim}{Z}_0$

$$|e_n\rangle = \underset{\sim}{Z}_0 \, |i_n\rangle \;\ldots\; (n=1 \text{ or } 2) \tag{1}$$

where $\underset{\sim}{Z}_0$ is a symmetric real matrix

$$\underset{\sim}{Z}_0 = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{12} & Z_{22} \end{bmatrix} \tag{2}$$

A section of a coupled asymmetric transmission line is a four-port which can be described in terms of generalized chain matrix equations

$$|V_I\rangle = \underset{\sim}{A} \, |V_{II}\rangle - j\underset{\sim}{B}|I_{II}\rangle \tag{3}$$

$$|I_I\rangle = j\underset{\sim}{C} \, |V_{II}\rangle - \underset{\sim}{D} \, |I_{II}\rangle \tag{4}$$

The real 2×2 matrices $\underset{\sim}{A}$ to $\underset{\sim}{D}$ are given in [5].

## Design equations

Once the circuit description of the asymmetric coupled section is available in the form of (3) and (4) it is possible to derive the transmission coefficient $S_{12}$ of the asymmetric dc block, and from there it may be possible to arrive at the set of design formulas for the device. The difficulties arise from the fact that the quasi TEM modes on the coupled section have different propagation constants. The obtained expressions for the transmission coefficients are simple enough for implementation on a computer, but they are too complicated for an analytical treatment. For that reason, a pure TEM case has first been analyzed, because it gives more manageable expressions. Under the pure TEM assumption, the asymmetric dc block has the following transmission coefficient

$$|S_{12}|^2 = \frac{1+\Omega^2}{a_0+a_2\Omega^2 a_4\Omega^4} \tag{5}$$

with $\Omega=\tan\beta_0\ell$, where both modes have the same propagation constant $\beta_0$. The **coefficients** $a_0$ to $a_4$ are the following

$$a_0 = \left[\frac{z_{12}^2 + 1}{2z_{12}}\right]^2 \tag{6}$$

$$a_2 = \frac{\left[z_{11}/\sqrt{\rho} + z_{22}\sqrt{\rho}\right]^2 - 2\Delta z\left[z_{12}^2 + 1\right]}{4z_{12}^2} \tag{7}$$

$$a_4 = \left[\frac{z_{11}z_{22} - z_{12}^2}{2z_{12}}\right]^2 \tag{8}$$

In the above, the normalized elements of the characteristic impedance matrix are denoted by the lower case symbols:

$$z_{ij} = Z_{ij}/\sqrt{R_S R_L} \tag{9}$$

The impedance transformation ratio is denoted by $\rho$

$$\rho = R_L/R_S \tag{10}$$

The desired rippled response is shown in Fig. 2, where $\Omega_c$ denotes the cutoff frequency in the transformed domain[2]

$$\Omega_c = \cot\left[\frac{\pi}{2}\left(1 - \frac{B_r}{2}\right)\right] \tag{11}$$

with $B_r$ denoting the relative bandwidth. For the standing

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wave ratio at the input to remain within a prescribed value S, the coefficient $a_0$ must be chosen as follows

$$a_0 = \frac{(S+1)^2}{4S} \qquad (12)$$

The remaining two coefficients are then set to be

$$a_4 = \frac{(S-1)^2}{4S(-1 + \sqrt{1+\Omega_c^2})^2} \qquad (13)$$

$$a_2 = a_0 - a_4\Omega_c^2 \qquad (14)$$

The normalized elements of the characteristic impedance matrix are then selected as follows

$$z_{11} = \sqrt{\rho}\ \sqrt{a_2\sqrt{S} + (S+1)\ \sqrt{a_4}} \qquad (15)$$

$$z_{22} = \frac{z_{11}\sqrt{S}}{\rho} \qquad (16)$$

$$z_{12} = \sqrt{S} \qquad (17)$$

Determination of microstrip dimensions

Voltage and current eigenvectors and the corresponding propagation constants of the quasi TEM modes are found from the solution of the electrostatic boundary value problem. The two conductors $w_1$ and $w_2$ are kept at potentials $\phi_1$ and $\phi_2$ with respect to the image plane, and the charge distributions on each of the conductors are computed. By using the two-dimensional Green's function and by enforcing the continuity of the normal electric flux density on the air-dielectric interface, a coupled pair of integral equations is obtained which is then solved by the moment method[7]. After the charge distributions on the individual conductors have been determined, it is possible to find the total charge on each conductor by simple summation of charges on elementary strips. The charge distribution is evaluated twice, first in presence of the dielectric material, then for the air dielectric. From these two results, the normalized modal eigenvectors $|e_n\rangle$ are computed. The characteristic impedance matrix is then[5]

$$\underset{\sim}{Z}_0 = \sum_{n=1}^{2} |e_n\rangle\langle e_n| \qquad (18)$$

By the procedure just described, the characteristic impedance matrix $\underset{\sim}{Z}_0$ of the asymmetric coupled microstrip can be evaluated for any given set of dimensions. There are three numbers which uniquely determine $\underset{\sim}{Z}_0$: s/h, $w_1/h$, and $w_1/w_2$. Therefore, a complete graphical representation of any element $Z_{ij}$ from the characteristic impedance matrix would require a three-dimensional diagram. This is, of course, impractical, and a simplified graphical determination of the microstrip dimensions has been found in the following manner.

The most demanding dimension for achieving the wide bandwidth is the spacing s. Therefore, the ratio s/h was fixed at the smallest feasible value, and the following two diagrams have been constructed: one representing $Z_{11}$ vs. $Z_{22}$ (Fig. 3), and the other representing $Z_{12}$ vs. $w_2/w_1$ (Fig. 4).

It is observed from (16) that the plot of $Z_{11}(Z_{22})$ is specified by a straight line. Such a straight line for $\rho$=2.7 and S=1.177 is shown in Fig. 3. Any point on this line will result in a transformer from $R_S$=50Ω to $R_L$=18.54Ω. The corresponding value $Z_{12}$ may be computed from (17) and (10):

$$Z_{12} = \sqrt{1.177 \cdot 50 \cdot 18.54} = 33.03\Omega$$

The horizontal line for 33.03Ω can now be drawn in Fig. 4 (dotted line). Along this line, one can read the pairs of numbers $w_1/h$ and $w_2/w_1$, which together give the required $Z_{12}$. These points should be mapped into Fig. 3, and then connected with a dotted line. The intersection of the solid line ($\rho$=const.) and the dotted line ($Z_{12}$=const.) gives then the required dimensions $w_1$ and $w_2$.

Practical realization

The first measured response of the asymmetric microstrip dc block, showed a larger reflection coefficient than computed, while the transmission coefficient displayed an aymmetric behavior with frequency. However, it was found experimentally that by adding a lumped capacitance $C_{II}$ on the low-impedance side of the coupler (see Fig. 1), the input standing wave ratio can be considerably improved. The addition of another capacitor $C_I$ on the high-impedance side of the dc block showed a less pronounced improvement. A computer program, based on the exact equations (3) and (4), has confirmed the deterioration of the response due to the inequality of propagation constants, the result being shown in Fig. 5 by the dotted line. By adding two capacitances $C_I$ and $C_{II}$ in the computer model of the circuit, the response can be improved. The best values of $C_I$ and $C_{II}$ have been eventually obtained by a numerical optimization. The computed response after optimization becomes almost identical with the ideal (pure TEM) response, as shown by the solid line in Fig. 5.

The capacitance values obtained by optimization are very small indeed: for a dc block designed to operate at center frequency 2.85GHz, the values are $C_I$=0.01pF and $C_{II}$=0.04pF. Only $C_{II}$ was actually mounted on the experimental model, and the measured transmission coefficient is now symmetric as seen in Fig. 6.

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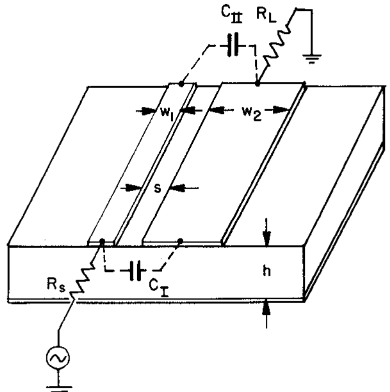


Figure 1. Asymmetric coupled microstrip.

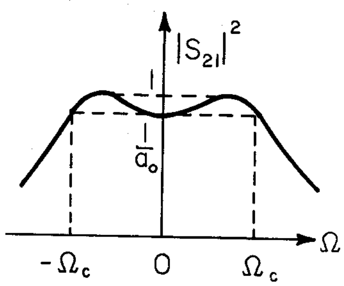


Figure 2. Rippled transmission coefficient.

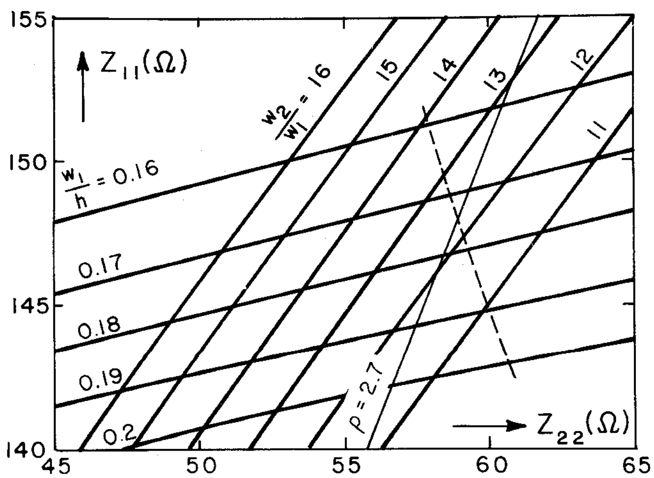


Figure 3. $Z_{11}$ vs. $Z_{22}$ for a substrate with $\varepsilon_r$=2.2 and conductor spacing s/h=0.16.

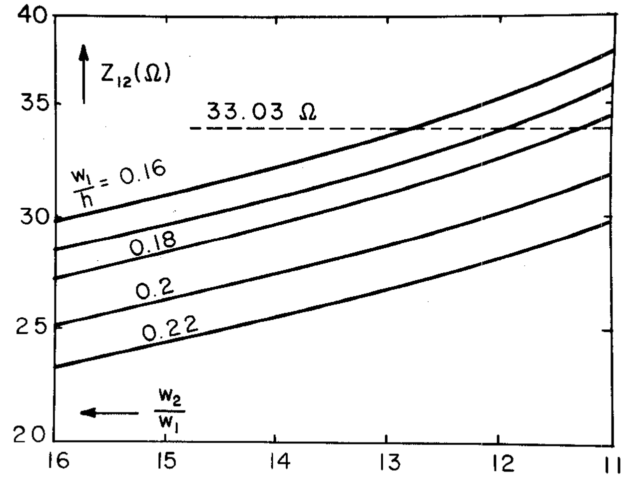


Figure 4. $Z_{12}$ vs. $w_1/w_2$ for the same substrate as in Figure 3.

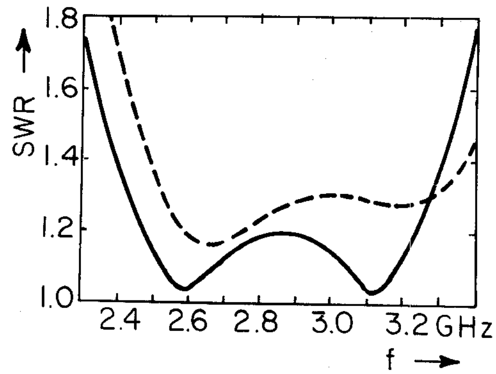


Figure 5. Dotted line: computed input standing-wave ratio without compensating capacitances. Solid line: same, with optimized compensating capacitances.

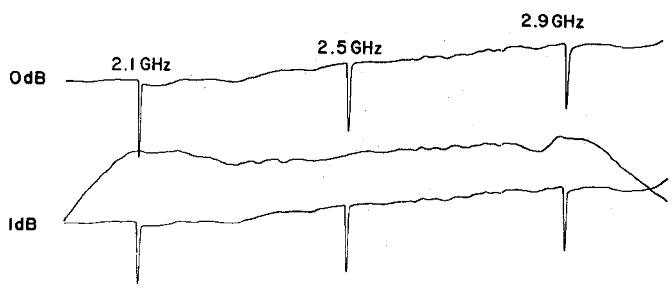


Figure 6. Measured transmission coefficient, $C_{II}$ added.