Compact Microwave Impedance Matching Using Patterned Conducting Planes

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Abstract ─ In this paper, a novel structure for impedance matching in microwave frequency range is introduced. The proposed structure utilizes one or more patterned conducting planes, which are placed transversely in a waveguide. The patterned conducting planes act as frequency selective surfaces for impedance matching purposes. The Green's function of one or more desired patterned planes, considering coupling effects between them, is obtained. Then magnetic field integral equations, in terms of unknown magnetic currents of patterned planes, are calculated and solved using method of moment. The suitable pattern of a patterned plane is obtained using optimization by genetic algorithm, so that the return loss of whole structure, in the desired frequency range vanishes.

The proposed structure, compared with traditional impedance matching structures, is very compact and wideband. It has been shown that its length, in comparison with another compact and wideband microwave impedance matcher, is fifty times shorter.

The usefulness of the proposed structure and its performance is verified in some examples, and the response of designed impedance matchers are compared with the results of simulations.

Index Terms ─ Frequency selective surfaces, genetic algorithm, impedance matching, magnetic field integral equation, patterned conducting planes.

I. INTRODUCTION

Impedance matching structures are very important parts of microwave systems. Impedance matching is necessary in the input of waveguide type power dividers and antennas [1-3]. The most important goal of impedance matching is maximum power delivery from the source to the load.

Some considerations in the design of impedance matching structures are complexity of implementation, adjustability, compactness and depending on the application, matching may be required over a band of frequencies, such that the bandwidth of the matching network is an important design parameter. Many approaches has been investigated for impedance matching in microwave devices such as using irises, rods, apertures, posts, and lumped microwave elements [1-4]; which could be used only for narrow band matching, i.e., a few percent. Other approaches are multi-section quarter-wave transformers, tapered homogeneous dielectrics filled waveguide [5,6], longitudinally inhomogeneous waveguides (LIWs) [7], and tapered or stepped transformer [8], whose length often is long; besides, construction of LIWs is not such practical.

In this article, we proposed patterned dielectric backed conducting planes as an impedance matcher, which is located transversely in a waveguide. Conducting plane of suitable pattern, supported by a dielectric layer acts as a frequency selective surface (FSS), which has many applications in microwave and antenna engineering. FSS improves the gain and return loss of antennas [9,10], it can also shift the phase of incident electromagnetic waves, and this characteristic could be used for beam steering of antenna [11], increases antenna gain, even for ultra-wide frequency band [12], transforms linear polarization to circular [13]. Also, it has been used as spatial microwave filter [14], and it has capability of switching between reflection and transmission for incident waves that was used for modifying the electromagnetic architecture of buildings [15].

Frequency response of a dielectric backed conducting plane such as its bandwidth, depends on the pattern of the plane, thickness and electric permittivity of the supported layer. So, using dielectric backed conducting planes in the structure of an impedance matcher, makes it possible to achieve compact

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impedance matching devices. In addition, according to the frequency selectivity of such planes, wide band or multiband impedance matcher can be designed. Design procedure is based on deriving magnetic field integral equations in terms of unknown magnetic currents of patterned conducting plane which can be solved using the method of moment. Suitable pattern for a desired specification can be extracted by genetic algorithm optimization.

II. DESIGN METHOD

Figure 1 (a) shows three dimensional view of the proposed impedance matcher, consisting of one dielectric backed patterned conducting plane which is located transversely in a waveguide. This structure is a two dimensional symmetric structure. Relative permittivity of dielectric layer is ε_r with thickness of d. Figure 1 (b) illustrates side view of the structure.

According to equivalence theorem, the equivalent magnetic current of the proposed structure in two regions has been shown in Fig. 1 (c). Using spectral domain immitance approach [16,17], one can derive magnetic type dyadic Green's functions in two side of conducting plane.

As it has been shown in Fig. 1 (c), the electric and magnetic fields in region of z<0 is due to induced magnetic current on the aperture parts of patterned plane and incident wave, which is TE_{10} mode, while fields in region of z>0 is only due to induced magnetic current on the aperture parts of patterned plane.

Then magnetic fields, in two side of the conducting plane, in terms of their spectral Green's function and magnetic current, were obtained. Then boundary condition of transverse magnetic fields across the aperture parts of the patterned plane must be satisfied. A coupled set of magnetic field integral equations is obtained by enforcing continuity of tangential component of magnetic fields at z=0.

Expressing the magnetic fields, in term of their spectral magnetic type Green's function, makes it possible to use the method of moment with traditional *x*- and *y*-directed piece-wise linear or triangle basis and testing functions. So, integral equations are converted into linear equations.

To derive linear equations, unknown magnetic currents can be considered as follows:

$$\mathbf{M}_x = \sum_{l=1}^{C(R-1)} \mathbf{C}_x B_l^x , \qquad (1)$$

$$\mathbf{M}_y = \sum_{l=1}^{R(C-1)} \mathbf{C}_y B_l^y , \qquad (2)$$

where B_l^x , and B_l^y are the *x*-directed and y-directed triangular sub-domain basis functions for $\mathbf{M}_x$ and $\mathbf{M}_y$ representation, and R and C are the number of subsections at x direction and y direction, respectively.

By applying Galerkin's method, the following matrix equations are obtained:

$$\mathbf{Z}_1\mathbf{C}_x + \mathbf{Z}_2\mathbf{C}_y = \mathbf{U}_x^{inc}, \qquad (3)$$

$$\mathbf{Z}_3\mathbf{C}_x + \mathbf{Z}_4\mathbf{C}_y = \mathbf{U}_y^{inc}. \qquad (4)$$

Each of above matrices has been given in the appendix. $\mathbf{C}_x$ and $\mathbf{C}_y$ are obtained by solving equations, and then x-directed and y-directed magnetic currents can be achieved.

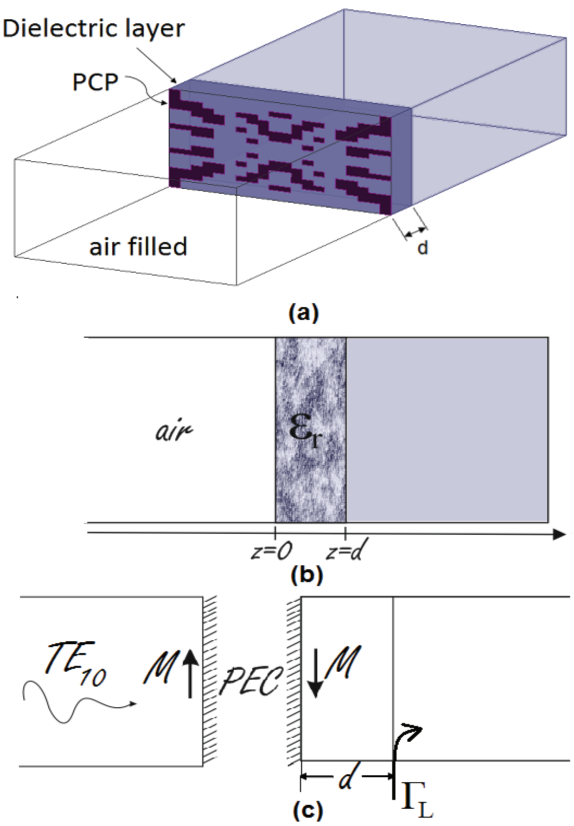

Fig. 1. The structure of proposed impedance matcher: (a) three dimensional view, (b) side view, and (c) equivalent structure of Fig. 1 (a).

The amplitude of reflected wave at z=0, is calculated by the following equation:

$$A^- = \frac{1}{P}\int_v \mathbf{H}_{inc}^+ .\mathbf{M}_x dv , \qquad (5)$$

where P is the total incident power, and $\mathbf{H}_{inc}$ is the magnetic field of dominant mode. So, reflection coefficient can be obtained from the following equation:

$$S_{11} = \frac{-\left|\mathbf{H}_{inc}^x\right| + A^-}{\left|\mathbf{H}_{inc}^x\right|}. \qquad (6)$$

It is aimed to achieve a specific pattern for patterned conducting plane, so that a desired reflection coefficient for the impedance matcher, in a given frequency range, can be obtained. This procedure is performed with MATLAB GA tool.

For this purpose, the cross-section area of waveguide is divided into $R \times C$ subsections. Non-metalized parts and metalized parts are presented in terms of 1 s and 0 s in the GA, respectively. Also, GA population type is set to bit string, consisting of 0 and 1 binary digits. Also, those are produced with uniform creation function.

Here, based on the symmetry excitation requirements (incident plane wave), it is assumed that the pattern of the patterned conducting plane is symmetric respect to horizontal and vertical axes of the waveguide cross-section. Consequently, GA operated only on one quarter region of cross-section [14].

The designing method is based on the optimization of a suitable fitness function so that, return loss of the structure can be adjusted to the desired return loss in the frequency range. The following fitness function is defined here:

$$fitness = \sqrt{\frac{1}{M}\sum_{m=1}^{M}\left|20\log\left|S_{11}\left(f_m\right)\right|-(\text{desired RL})\right|^2}, \quad (7)$$

where $S_{11}(f_m)$ are reflection coefficient of the structure at frequency samples, calculated with mentioned manner and M is the number of frequency samples in the frequency range.

III. EXAMPLES AND RESULTS

A. Example I

In the first example, an impedance matching network in a frequency range from 8 to 12 GHz (X-band) using a waveguide with dimensions of 0.9 and 0.4 (in), WR-90, and one dielectric backed patterned conducting plane with the relative electric permittivity of 3.55 and thickness of 0.4 mm is designed. Three dimensional view of this structure is shown in Fig. 1 (a).

We consider the right medium of the network has been connected to a frequency dependent load impedance, whose reflection coefficient is varying from 0.48×exp (1.91i) to 0.28×exp (1.45i) in the frequency range. We also considered 10×10 subsections for the quarter of conductive plane and M=17 frequency samples with equal distance of 250 MHz.

Figure 2 illustrates the designed geometry of conductive plane. Figure 3 compares the return loss of the designed impedance matching structure with the simulated one, which is obtained by HFSS. It is obvious that there is a good agreement between two curves in the frequency range.

Also, an impedance matcher using a waveguide filled by inhomogeneous dielectrics for the same load is presented in [7]. Figure 4 compares the return loss of the proposed impedance matcher in this example and that in [7]. This impedance matcher is fifty times as short as that impedance matcher; besides, the constructing of inhomogeneous dielectrics is not such practical, while constructing of the proposed structure is quite practical.

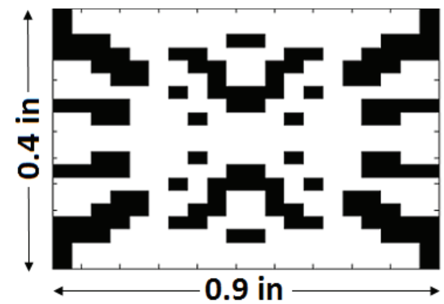

Fig. 2. The designed geometry of patterned conducting plane of impedance matching structure.

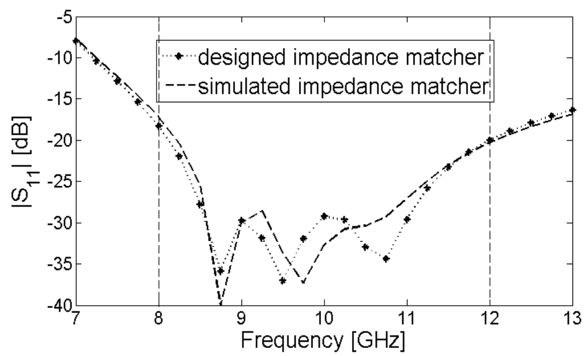

Fig. 3. The return loss of designed, and simulated impedance matching structure.

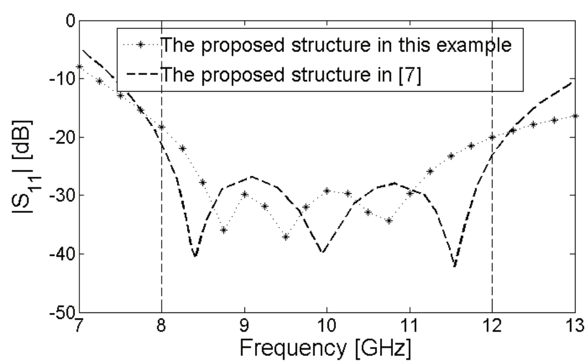

Fig. 4. The return loss of the proposed impedance matcher in this example versus the proposed one in [7].

B. Example II

In the second example, impedance matching network for a frequency dependent load impedance, whose reflection coefficient is varying from 0.64×exp (1.97i) to 0.43×exp (1.4i) in X-band, is proposed. All specification of the network is the same as previous example, except the pattern of the conducting plane. We considered 8×12 subsections for the quarter of conductive plane and M=27 frequency samples with equal distance of 150 MHz.

Figure 5 illustrates the designed geometry of conductive plane. Figure 6 compares the return loss of the designed impedance matching network obtained by MATLAB with the simulated one obtained by HFSS. As it shows, this approach is effective to obtain a good impedance matching condition in desired frequency range.

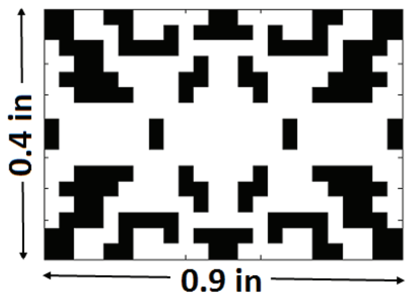

Fig. 5. The designed geometry of patterned conducting plane of impedance matching structure.

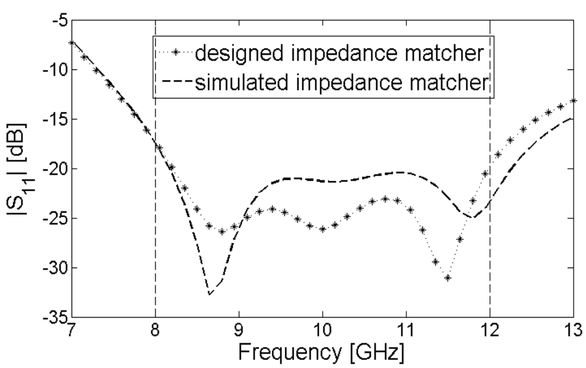

Fig. 6. The return loss of designed and simulated impedance matching structure.

C. Example III

In the third example, impedance matching network for a frequency dependent load impedance, whose reflection coefficient is varying from 0.43×exp (3.14i) to 0.3×exp (3.14i) in X-band, using the same WR-90 is designed. For given reflection coefficient, we have to utilize more than one patterned conducting plane to realize impedance matching structure. In this example, three patterned conducting plane is considered; also, the first and the last planes are the same. The space between them is assumed to be 2.5 mm, and all are selected with relative permittivity of 3.55 and thickness of 0.4 mm. So the total length of impedance matching structure is 6.2 mm. Its three dimensional view is shown in Fig. 7. In this design procedure, coupled magnetic field integral equations are used, which has been obtained in [14]. It is assumed that the number of subsections is 8×12, also, M=17 frequencies with equal distance of 250 MHz, is considered in the optimization process. Also, Fig. 8 illustrates the designed geometry of patterned conducting planes.

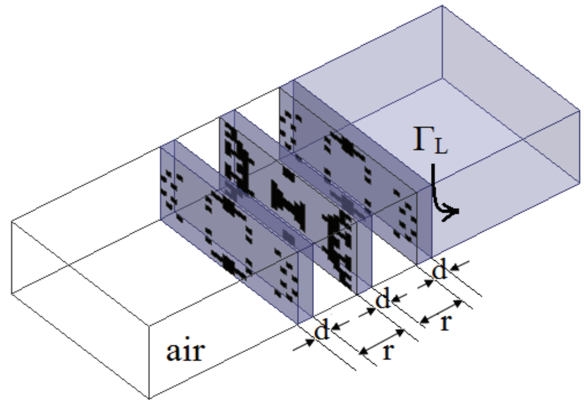

Fig. 7. Three dimensional view of impedance matching structure for Γ_L=1/3exp (jπ), based on 8-12 GHz optimization frequency range.

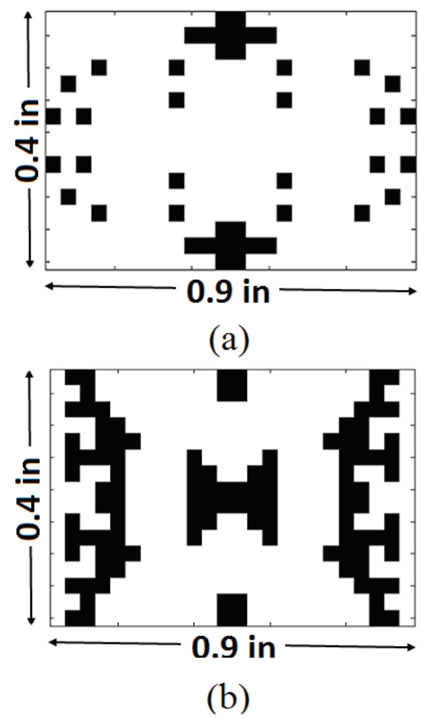

Fig. 8. The designed geometry of three FSSs: (a) geometry of first and third FSS, and (b) geometry of second FSS.

Figure 9 compares the return loss of the designed impedance matcher with the simulated one. It can be deduced that the designed structure has yielded a good impedance matching in the desired frequency band.

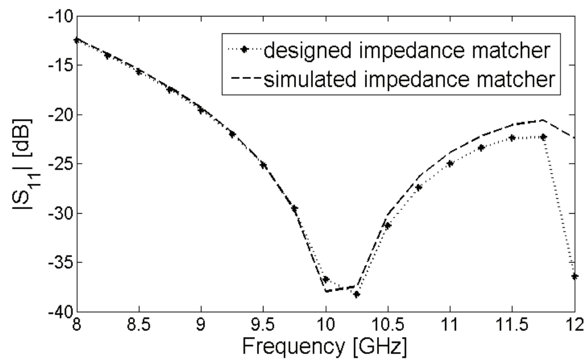

Fig. 9. The return loss of the designed impedance matcher, and that of the simulated one.

IV. CONCLUSION

In this paper, the idea of using dielectric backed patterned conducting planes in a waveguide for impedance matching was confirmed. A designing approach for finding suitable pattern of patterned conducting planes (PCPs) was introduced. The results of simulation show using only one dielectric backed PCP with inconsiderable thickness about 0.4 mm, adjust some loads to air filled waveguides. Matching condition was good about -25 dB return loss in a wide frequency band about 40 percent. Also, for other loads, which one dielectric backed PCP was not sufficient for their matching, we could use more than one dielectric backed PCPs. The usefulness and performance of the proposed structure was verified by testing three extended frequency dependent loads. The proposed method can be applied to any realizable and practical loads in desired frequency range, and for each given loads there is a specified patterned plane which can match the given load to the air-filled waveguide in the desired frequency range.

V. APPENDIX

In this appendix, we gave the matrices of third and fourth equation:Z

$$\left[\mathbf{Z}_1\right]_{kl} = \sum_{m=0}^{\infty}\sum_{n=0}^{\infty} \tilde{B}_k^x\, \tilde{B}_l^x \left(\tilde{G}^1{}_{xx} + \tilde{G}^2{}_{xx}\right), \tag{A1}$$

$$\left[\mathbf{Z}_2\right]_{kl} = \sum_{m=0}^{\infty}\sum_{n=0}^{\infty} \tilde{B}_k^x\, \tilde{B}_l^y \left(\tilde{G}^1{}_{xy} + \tilde{G}^2{}_{xy}\right), \tag{A2}$$

$$\left[\mathbf{Z}_3\right]_{kl} = \left[\mathbf{Z}_2\right]_{lk}, \tag{A3}$$

$$\left[\mathbf{Z}_4\right]_{kl} = \sum_{m=0}^{\infty}\sum_{n=0}^{\infty} \tilde{B}_k^y\, \tilde{B}_l^y \left(\tilde{G}^1{}_{yy} + \tilde{G}^2{}_{yy}\right), \tag{A4}$$

where $\tilde{B}_l^x$ and $\tilde{B}_l^y$ are the Fourier transform of x-directed and y-directed triangular sub-domain basis functions, respectively; and $\tilde{B}_k^x$ and $\tilde{B}_k^y$ are the Fourier transform of x-directed and y-directed triangular testing functions, respectively.

$$\mathbf{C}_x = \begin{bmatrix} c_{1,1}^x \\ c_{1,2}^x \\ . \\ . \\ c_{R-1,C}^x \end{bmatrix}, \tag{A5}$$

$$\mathbf{C}_y = \begin{bmatrix} c_{1,1}^y \\ c_{1,2}^y \\ . \\ . \\ c_{R,C-1}^y \end{bmatrix}, \tag{A6}$$

$$\left[\mathbf{U}_x^{inc}\right]_k = 2Y_a^{TE10}(ab)^{0.5}\, \tilde{B}_k^x\, (m=1, n=0), \tag{A7}$$

$$\left[\mathbf{U}_y^{inc}\right]_k = 0, \tag{A8}$$

$$\tilde{G}_{xx}^i = Y^{i,h}\sin^2\theta + Y^{i,e}\cos^2\theta, \tag{A9}$$

$$\tilde{G}_{xy}^i = (Y^{i,h} - Y^{i,e})\sin\theta\cos\theta, \tag{A10}$$

$$\tilde{G}_{yx}^i = \tilde{G}_{xy}^i, \tag{A11}$$

$$\tilde{G}_{yy}^i = Y^{i,h}\cos^2\theta + Y^{i,e}\sin^2\theta, \tag{A12}$$

$$Y^{1e,h} = Y_a^{TM,TE}, \tag{A13}$$

$$Y^{2e,h} = Y_d^{TM,TE}\, \frac{Y_L^{TM,TE} + Y_d^{TM,TE}\tanh(\gamma_d d)}{Y_d^{TM,TE} + Y_L^{TM,TE}\tanh(\gamma_d d)}, \tag{A14}$$

$$\Gamma_L = (Y_a^{TE_{10}} - Y_L^{TE_{10}}) / (Y_a^{TE_{10}} + Y_L^{TE_{10}}), \tag{A15}$$

where Y_a, Y_L and Y_d are wave admittance, in air, load and dielectric layer, respectively.

$$\sin\theta = \frac{k_x}{\sqrt{k_x^2 + k_y^2}}, \tag{A16}$$

$$\cos\theta = \frac{k_y}{\sqrt{k_x^2 + k_y^2}}, \tag{A17}$$

$$k_x = \frac{m\pi}{a}, k_y = \frac{n\pi}{b}. \tag{A18}$$

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