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POWER REFLECTANCE MEASUREMENTS OF METAL MESHES AND GRIDS USING A QUASI-OPTICAL RESONATOR

B. Kapilevich,[1] A. Gover,[2] and A. Eliran[2]
[1] Department of Electrical and Electronics Engineering
College of Judea and Samaria
Ariel, 44838, Israel
[2] Department of Electrical Engineering
Physical Electronics
Tel Aviv University
Tel Aviv, 69978, Israel

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ABSTRACT: *This paper presents a novel approach developed for the measurement of the power reflectance of metal meshes and wire grids using a quasi-optical resonator. It is based on the reconstruction of a grid's reflectance from the resonance curve of a quasi-optical resonator. The model of the resonator loaded by a mesh has been developed to provide direct reconstruction of its reflectance. A comparison of the measured results with data obtained in free-space measurements has shown good agreement. Examples of reflectance measurements of both conventional and twin-shaped meshes are presented to illustrate the technique suggested.* © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 45: 185–188, 2005; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.20765

Key words: *wire grids; metal meshes; quasi optical resonators; measurements*

1. INTRODUCTION

One-dimensional wire grids and 2D metal meshes are widely used as elements of quasi-optical devices: polarizing filters, deflectors, quasi-optical gratings, semi-transparent mirrors, different coupling elements, and so forth [1]. Numerous models have been developed to estimate the reflectance of these elements [2–6]. However, these models' accuracy becomes poor when the size of the mesh's cell or the distance between wires are comparable with wavelength. As a result, experimental estimation of reflectivity in such situations is mandatory. Basically, direct free-space measurements using reflection or transmission mode are widely employed [7]. However, multiple reflections from feeding elements and supporters are sources of decreasing measurement accuracy. Also, measurement of highly reflective structures in free space is not easy to do, since it requires a reflectometer with high resolution.

An alternative approach developed in this paper is based on the measurement of the resonance curve of a quasi-optical resonator loaded by a mesh or grid which may be considered as a coupling element. The coupling coefficient of the resonator with exciting guide is a function of mesh reflectance. The latter can be reconstructed from the measured resonance curve. To do this, a model of a quasi-optical resonator linked directly with input mesh reflectance needs to be developed. This model as well as the procedure applied for the reconstruction of reflectivity from the measured resonance curve, are described below.

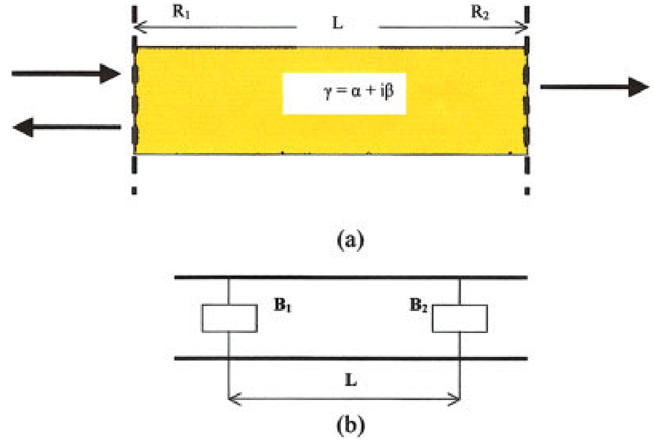

Figure 1 (a) Fabry–Perot resonator filled by a lossy medium; (b) microwave representation of a Fabry–Perot resonator. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com.]

2. THE MODEL DEVELOPED FOR RECONSTRUCTION OF POWER REFLECTANCE

Basic optical characteristics, such as attenuation constant of the resonator guide and power reflection coefficient of the meshes (grids), are introduced into the microwave-circuit model in order to provide a direct link between the measured data and reconstructed parameters. This method, briefly discussed below, is based on a transmission-matrix description [8]. Indeed, a Fabry–Perot resonator of length L can be described by its equivalent configuration shown in Figure 1(a). The lossy medium is placed between the two mirrors which are characterized by own power reflectivities R_1 and R_2. In our case, wire grids or metal meshes play the role of mirrors.

The lossy medium is characterized by a propagation constant $\gamma = \alpha + i\beta$, where α is the field attenuation constant and $\beta = 2\pi/\lambda$ is the wave number of the selected mode. The same resonator can be represented by microwave parameters to simulate the behavior of a mirror. We can replace them by the two reactive elements B_1 and B_2, as shown in Figure 1(b). Now a Fabry–Perot resonator can be analyzed in terms of a transmission ABCD matrix as follows:

$$T_{R1} = \begin{bmatrix} 1 & 0 \\ B_1 & 1 \end{bmatrix},\ T_{R2} = \begin{bmatrix} 1 & 0 \\ B_2 & 1 \end{bmatrix},\ T_L = \begin{bmatrix} \cosh(\gamma L) & \sinh(\gamma L) \\ \sinh(\gamma L) & \cosh(\gamma L) \end{bmatrix}, \tag{1}$$

where T_{R1} and T_{R2} are transmission matrices of the reactive elements and T_L is a transmission matrix of the lossy line of length L. By doing standard multiplication of these matrices, we can calculate the resulting transmission matrix T_{res} of the Fabry–Perot resonator as follows:

$$T_{res} = [T_{R1}][T_L][T_{R2}]. \tag{2}$$

The A, B, C, D elements of T_{res} are given by

$A = \cosh(\gamma L) + B_2 \sinh(\gamma L)$, $B_1 = \sinh(\gamma L)$,

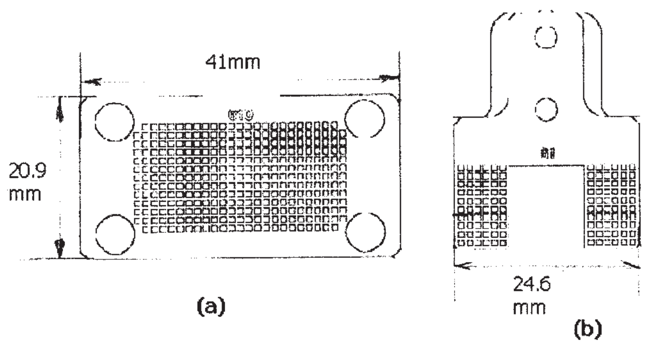

Figure 2 The configurations of measured meshes: (a) rectangular and (b) twin-shaped forms

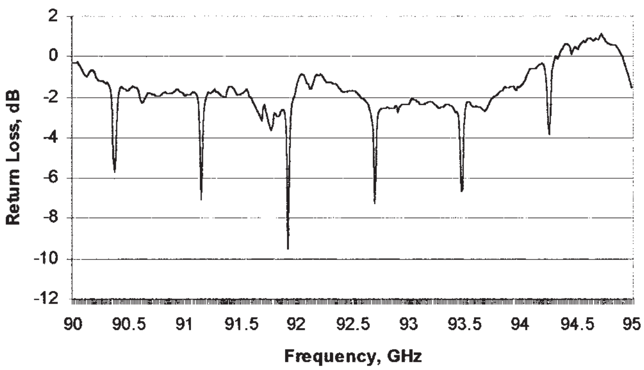

Figure 3 Typical frequency response of the quasi-optical resonator with metal mesh at its input

$$C = \sinh(\gamma L) + B_1\cosh(\gamma L) + B_2[B_1\sinh(\gamma L) + \cosh(\gamma L)],$$

$$D = B_1\sinh(\gamma L) + \cosh(\gamma L). \quad (3)$$

The parameters B_1 and B_2 are directly linked with power reflectance R_p using the following expression:

$$B(R_p) := \sqrt{\frac{R_p}{(1 - R_p)}}. \quad (4)$$

Now we can derive the formula for an input reflection $\Gamma(R_p, \alpha, f)$ of the resonator considered as a function of power reflectivity R_p, attenuation constant α, and frequency f using the well-known expression [8]:

$$\Gamma(R_p, \alpha, f) = (A + B - C - D)/(A + B + C + D). \quad (5)$$

From a practical point of view, the measurement of reflectivity can be done more easily with a one-port resonator configuration that corresponds to the following conditions: $B_1 = \pm iB$, $B_2 = \infty$ (short-circuit line). Substituting Eqs. (3) and (4) into Eq. (5), we can write the analytical expression to be used for the reconstruction of power reflectance of the mesh under test from the measured resonator frequency response, namely:

$$\Gamma(R_p, \alpha, f) = \frac{\left[1 - i2\sqrt{\frac{R_P}{1 - R_P}}\right]\tanh[(\alpha + i\beta)L] - 1}{\left[1 + i2\sqrt{\frac{R_P}{1 - R_P}}\right]\tanh[(\alpha + i\beta)L] + 1}. \quad (6)$$

It should be pointed out that Eq. (6) is valid for any loss and can be used for direct reconstruction of power reflectivity from the measured $\Gamma(R_p, \alpha, f)$ curve.

3. THE RECONSTRUCTION PROCEDURE

To built up the procedure for reconstructing R_p and α, the two measurements of Γ must be done at some specified frequencies—$\Gamma(R_p, \alpha, f_0)$ and $\Gamma(R_p, \alpha, f_1)$—where f_0 is the resonance frequency and f_1 is an arbitrary frequency located just near resonance. Assuming that the measured values are A_0 and A_1, respectively, the following system of two nonlinear equations can be written for unknown R_p, and α:

$$\Gamma(R_p, \alpha, f_0) = A_0, \qquad \Gamma(R_p, \alpha, f_1) = A_1. \quad (7)$$

The system of Eq. (7) can be solved numerically using different techniques. The results of the reconstruction given below were obtained with the following standard algorithms available from the MathCad 2001 library: Conjugate Gradient, Quasi-Newton and Levenberg–Marquardt.

When solving equations numerically, it is necessary to define values from which the solver has to start the search for a solution of Eq. (7). To create a set of guess values, some preliminary information concerning the behavior of reflection coefficient $\Gamma(R_p, \alpha, f)$ in the space $\{R_p, \alpha\}$ is required near the resonance frequency. This can be obtained, for instance, by using a proper approximate analytical model for the resonator and grid [9]. It should be pointed out that the resonance curve becomes asymmetrical when the parasitic modes are excited. In this case, measurements of the return loss must be done on both sides of the resonance curve (below and above the resonance frequency). The value of frequency detuning $\Delta f = f_1 - f_2$ depends on the asymmetry of the resonance curve. A trade-off needs to be used to determine almost the same resonator parameters for both detuned frequencies—f_1 and f_2. The average value of the power reflectance estimated for these detuned frequencies is used below as a final result of the measurement discussed. Then, the average values of R_p and α are substituted in Eq. (6) in order to make a comparison with the measured resonance curve. This procedure has been

TABLE 1 Summary of Resonator Parameters Used for Reconstructing Power Reflectivity from Measured Resonance Curve

f_0 [GHz]	f_1 [GHz]	f_2 [GHz]	RL_0 [dB]	RL_1 [dB]	RL_2 [dB]	α_{av} [1/m]	$R_{p\text{-}av}$	$R_{p\text{-}fs}$
106.7895	106.7845	106.793	−10.63	−8.454	−8.542	0.139	0.956	0.944
102.8725	102.87	102.876	−9.366	−7.762	−7.663	0.151	0.954	0.942
96.6205	96.617	96.624	−8.916	−7.652	−7.432	0.190	0.937	0.975
91.1525	91.149	91.158	−9.48	−8.355	−8.16	0.193	0.945	0.982
86.3165	86.308	86.324	−15.58	−7.376	−7.092	0.122	0.938	0.979
79.219	79.216	79.222	−10.8	−9.52	−9.597	0.148	0.954	0.987

f_0, f_1, and f_2 are resonance and detuned frequencies; RL_0, RL_1 and RL_2 are measured return losses at the above frequencies; α_{av} is averaged attenuation constant after reconstruction; $R_{p\text{-}av}$ is averaged power reflectivity after reconstruction; $R_{p\text{-}fs}$ is measured power reflectivity in free space.

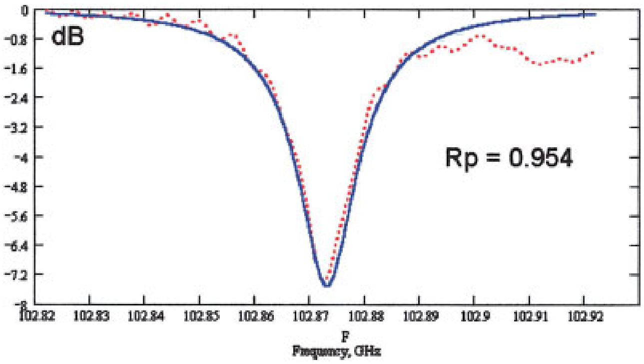

Figure 4 Measured (dotted line) and reconstructed (solid line) return loss (dB) for the mesh given in Table 1. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com.]

applied to the meshes of the square cells shown in Figure 2 with rectangular and twin-shaped configurations.

3. WIDEBAND MEASUREMENTS OF POWER REFLECTANCE

To measure reflectance in a wide frequency range, we used a set of fixed modes excited in a quasi-optical resonator. A short-circuited resonator ($R_2 = 0$), as shown in Figure 1, with length $L = 0.1893$ m and rectangular cross section 10.5×24.8 mm^2 was excited by a rectangular horn illuminating input mesh. The mesh's dimensions are 11×29.5 mm^2, the period is g_x, $g_y = 1.1$ mm, the width of the metal strips between apertures is $2a = 0.35$ mm, the material is stainless steel [see Fig. 2(a)]. An example of measured spectrum in the 90–95 GHz band is depicted in Figure 3. Several selected resonance modes were chosen within 75–110 GHz for measurement by HP-8757D network analyzer. Their resonance frequencies, as well as the other parameters needed for the reconstruction procedure, are given in Table 1.

The averaged attenuation constants and power reflectances given in Table 1 were substituted into Eq. (6) for comparison with the measured resonance curve. A typical example of such a comparison is depicted in Figure 4. Due to the excitation of higher-order modes, the resonance curve is asymmetrical at frequencies above resonance. However, the reconstruction procedure works well, even in the presence of asymmetry of the resonance curve. In the case discussed, we have obtained very close values of the reconstructed parameters for both sides of the resonance curve:

- $\alpha = 0.156$ [1/m] and $R_p = 0.952$ for frequency set f_0 and f_1;
- $\alpha = 0.147$ [1/m] and $R_p = 0.956$ for frequency set f_0 and f_2.

We can state that both the measured and constructed data are in good agreement, thus proving that the roots of nonlinear system (7) have been found correctly.

4. MEASUREMENTS OF THE POWER REFLECTANCE OF TWIN-SHAPED MESHES

Some quasi-optical devices employ beam-splitting based on the Talbot effect [10]. Figure 5 shows the structure of a split beam used to illuminate a twin-shaped mesh. Here, direct free-space measurement of reflectance is very difficult to carry out. However, if such a mesh is used as a coupling element of a quasi-optical resonator excited via a Talbot splitter, as shown in Figure 6, the

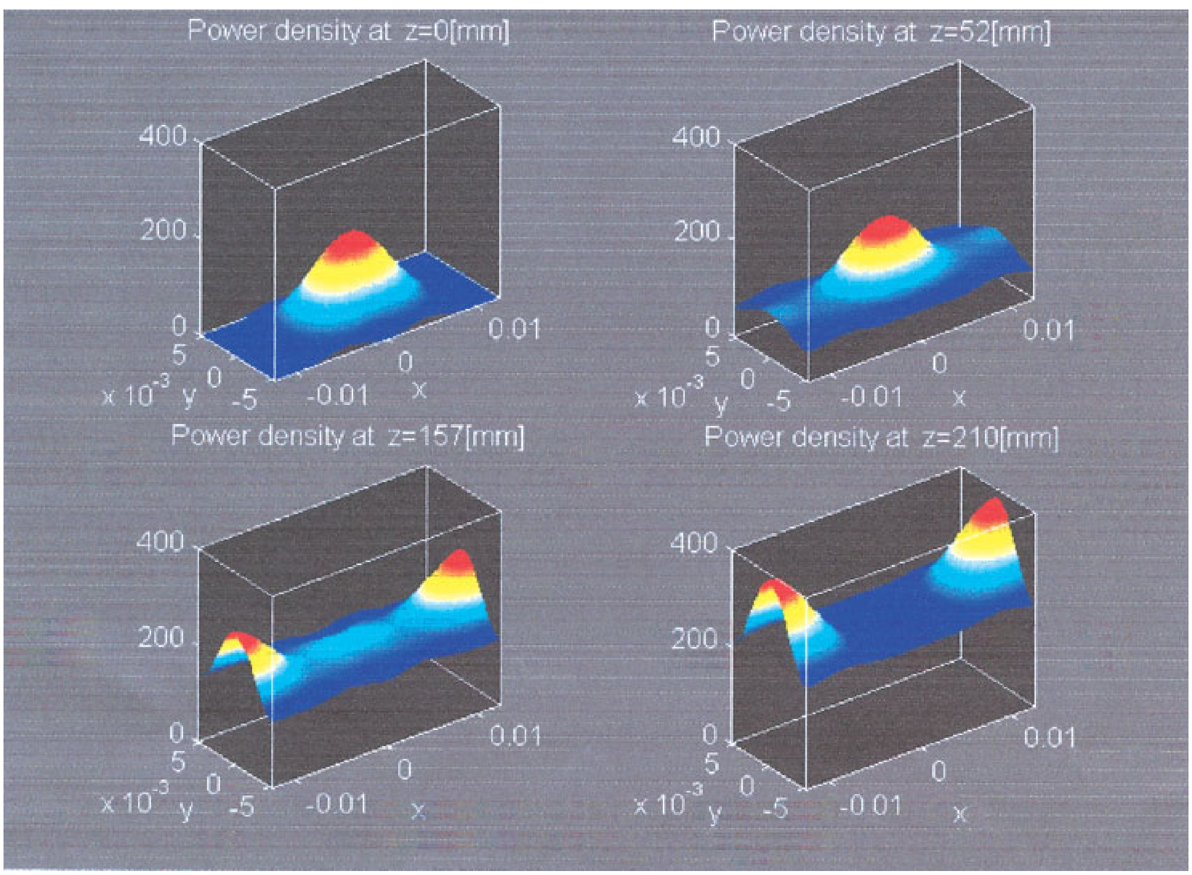

Figure 5 Illustration of beam splitting used in the resonator with twin-shaped mesh. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com.]

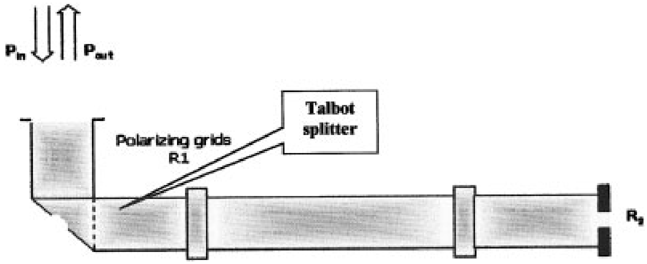

Figure 6 Configuration of the resonator with a grid inside the Talbot splitter

method proposed can be very successful. As an illustrative example, we consider a quasi-optical resonator with twin-shaped metal mesh [Fig. 2(b)] denoted as R_1, placed inside the Talbot reflector (see Fig. 6). The mesh is illuminated by a Gaussian beam splitter inside the Talbot splitter. Element R_2 is a short with a hole in the center to allow transport of an electron beam, since the resonator is part of a free-electron maser. The total resonator length determining its resonance frequencies is $L = 1.31$ m. The resonator is excited by a rectangular horn antenna. A scalar network analyzer HP-8757D measures the input reflection coefficient (return loss). Figure 7 shows the measured results (dotted line) and reconstructed data (solid line), based on the method suggested, for the grid dimensions considered in the previous section. The power reflectance of this grid is $R_P = 0.9414$ at the frequency 99.938 GHz, which is very close to the data reported in Table 1 for a rectangular grid with the same cell dimensions.

The stability of the solution of the system of the two nonlinear equations of Eq. (7) has also been investigated in the case discussed. The following range of guessed values, which provide the same solutions of Eq. (7), have been determined: $0.01 < \alpha < 0.5$ [1/m] and $0.8 < R_p < 0.99$ [%].

CONCLUSION

A model linking the reflectivity of metal meshes and grids and used as a coupling element of a quasi-optical resonator has been presented. Based on this model, the reflectance measurements of rectangular and twin-shaped meshes have been done by applying reconstruction algorithms. The measurements were based on an asymmetrical resonance curve and their validity was demonstrated.

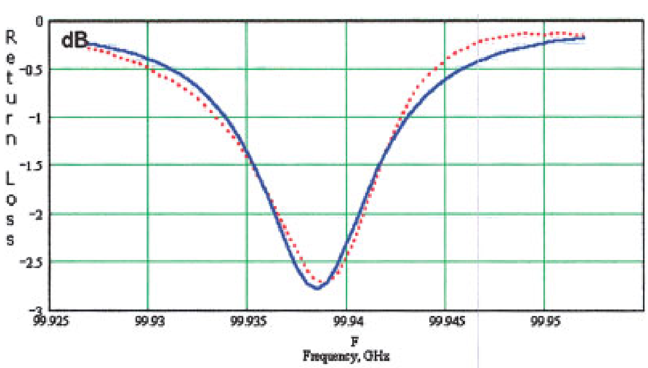

Figure 7 Measured results (dotted line) and reconstructed data (solid line) for the resonator with twin-shaped mesh. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com.]

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BAND-NOTCHED ULTRA-WIDEBAND CIRCULAR-DISK MONOPOLE ANTENNA WITH AN ARC-SHAPED SLOT

Kin-Lu Wong,[1] Yun-Wen Chi,[1] Chih-Ming Su,[1] and Fa-Shian Chang[2]

[1] Department of Electrical Engineering
National Sun Yat-Sen University
Kaohsiung 804, Taiwan
[2] Department of Electrical Engineering
Military Academy
Feng-Shan 830, Taiwan

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ABSTRACT: *By embedding a simple arc-shaped slot in a circular-disk monopole antenna, an ultra-wideband (UWB, 3.1–10.6 GHz) operation with a controlled notched frequency band can be obtained. The arc-shaped slot is placed close to the boundary of the circular-disk monopole and has a length of about one-half wavelength of the desired notched frequency. The proposed slotted circular-disk monopole antenna showing UWB operation with a notched frequency band for rejecting the 5.8-GHz WLAN band is demonstrated. The effects of the dimensions of the arc-shaped slot on the notched frequency band are also analyzed.* © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 45: 188–191, 2005; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.20766

Key words: *antennas; circular disk monopole antennas; ultra-wideband (UWB) monopole antennas; band-notched UWB antennas*