in contrast to the electrostatic case [3] where the electric polarisability of a single perforation is larger than other cases. Fig. 3 shows the behaviour of $\psi(0)$ against the distance between adjacent apertures $L/(2a)$. Case (i) has a higher total magnetic polarisability than case (ii), especially as $L/(2a)$ decreases, while the square-like case (ii) has a higher total electric polarisability than case (i) [3].

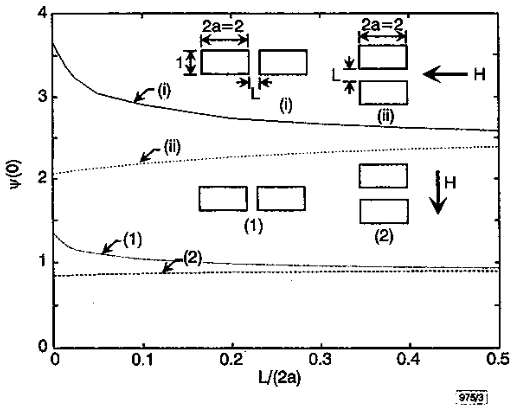

Fig. 3 *Total magnetic polarisability ψ(0) against L/2a*

Aperture thickness $d = 0$

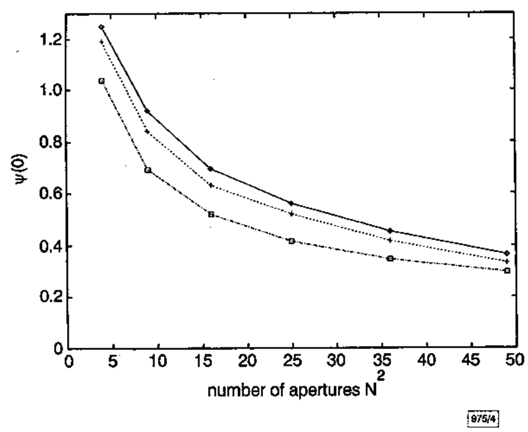

Fig. 4 *Total magnetic polarisability ψ(0) against number of apertures N^2*

Aperture thickness $d = 0$
◇ $L = 0.01$
+ $L = 0.03$
□ $L/(2a) = 0.5$

Note that the total longitudinal magnetic polarisability ψ_x of case (i) rises exponentially as $L/(2a) << 0.1$, implying that it behaves as a single slit. Note that ψ_x and ψy of the single aperture with ($2a = 4$, $2b = 1$) are 6.49 and 0.85, respectively [2]. Our computation shows that the coupling effect between adjacent apertures is almost negligible in the case of $L/(2a) > 0.5$. Fig. 4 shows the behaviour of the total magnetic polarisability $\psi(0)$ against the number of apertures (N^2) where the total area of apertures ($2aN \times 2aN$) is chosen to be 4 ($a = 1/N$). The geometry of the multiple apertures in Fig. 4 is the same as in Fig. 5 ($L = 0.01, 0.03$) of [3]. It should be noted that the curve represented by square symbols is approximated by $2.05/N$ which ignores the coupling effect between the apertures. As L decreases, $\psi(0)$ increases slightly due to an increase in the coupling effect.

Conclusion: The rigorous solution of the magnetic polarisabilities for multiple rectangular apertures in a thick conducting plane has been obtained in the form of a numerically efficient series. Numerical results have been shown to illustrate the behaviour of the magnetic polarisabilities for different aperture thickness, shape and distance. The salient feature of the magnetic polarisability differed from the electric case has been discussed.

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Circuit model for mode conversion in coplanar waveguide asymmetric series-impedances

M. Ribó, J. de la Cruz and L. Pradell

A new 'circuit model' for converting between even and odd modes in asymmetric series impedances in the ground plane of a coplanar waveguide is presented. The model is based on the separation of modes into two input and two output ports. In contrast to previous work, it enables a quantitative analysis of the energy exchange between modes to be obtained.

Introduction: Uniplanar guiding structures, such as coplanar waveguide (CPW) and slotline structures, are widely used in hybrid and monolithic (MMIC) microwave integrated circuits. The CPW is a multimode waveguide supporting two fundamental modes, the coplanar even mode (CEM), or coplanar mode, and the coplanar odd mode (COM), or slotline mode. Slotline supports only one fundamental (slotline) mode. In a CPW, series impedances can be easily connected as slotline stubs in the ground planes, allowing a number of circuit functions to be obtained, such as filters, resonators, and impedance-matching circuits [1], transitions [2 – 4] and baluns [5].

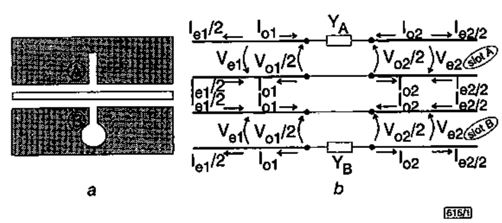

Fig. 1 *Asymmetric series impedances in ground planes of CPW*

a Typical situation
b Slot voltages and currents at admittance plane

Although asymmetric slotline stubs in the ground planes (Fig. 1*a*) are widely used, only a few models have been proposed to explain the energy exchange between the CEM and COM that takes place due to the asymmetry. In [1], a model was presented for the case where the slotline stubs are symmetric; owing to the symmetry, no mode conversion takes place, and only the transmission of the CEM between the input and the output port can be modelled. In [5], a technique was presented for using asymmetric slotline stubs to perform a coplanar-to-slotline transition;

although a model was presented, it only explains the mode conversion from the CEM to the slotline mode. In [2 – 4], models were presented for a slotline-to-CPW transition with a slotline stub in the CPW ground plane and airbridges at the CPW sections; owing to the airbridges, the COM mode is not generated and the models cannot be used in a general case. A generic CPW discontinuity has been modelled as an equivalent four-port 'black-box' [6], each port corresponding to a mode (either CEM or COM); since this method requires measurements, it is not easily applicable to CAD.

In this Letter, a new four-port 'circuit model' for the energy exchange between the CEM and the COM at asymmetric series impedances in the CPW ground plane is proposed. Since the circuit separates the CEM and the COM into different ports, it overcomes the limitations of previous models. Namely, it allows the CPW series impedances to be excited with either mode (CEM or COM), providing a quantitative analysis of mode conversion, and enables an analysis to be made of the series impedances in the ground planes of a CPW when the CPW is loaded with (coplanar) structures presenting different responses to each mode. Moreover, the model is easily implemented in microwave CAD. Validation of the model is achieved through its application to a 25GHz slotline series resonator in the CPW ground plane.

Model derivation: Consider the structure shown in Fig. 1*b*, which comprises a symmetrical CPW section with two series admittances, Y_A and Y_B (the input admittances of slotline stubs A, B, in Fig. 1*a*), connected along the CPW ground planes. Either CPW slot can be seen as a transmission line propagating both the CEM and COM. We denote the left and right sides of the CPW slots by subscripts i = 1, 2 respectively. The slot voltages and currents at the admittance plane, V_{Ai}, V_{Bi}, I_{Ai}, I_{Bi} (i = 1, 2) are written as a function of the CEM and COM voltages and currents V_{ei}, V_{oi}, I_{ei}, I_{oi} (i = 1, 2) as follows:

$$V_{Ai} = (V_{oi}/2) - V_{ei} \qquad V_{Bi} = (V_{oi}/2) + V_{ei} \qquad (1)$$

$$I_{Ai} = I_{oi} - (I_{ei}/2) \qquad I_{Bi} = I_{oi} + (I_{ei}/2) \qquad (2)$$

Substituting eqns. 1 and 2 into the boundary conditions (Kirchoff laws) $I_{A1} = -I_{A2}$, $I_{B1} = -I_{B2}$, $V_{A1} - V_{A2} = I_{A1}/Y_A$, $V_{B1} - V_{B2} = I_{B1}/Y_B$, imposed by the admittances Y_A and Y_B, the following equation system is obtained:

$$\begin{bmatrix} I_{e1} \\ I_{o1} \\ I_{e2} \\ I_{o2} \end{bmatrix} = \frac{1}{4} \begin{bmatrix} 4(Y_A+Y_B) & 2(Y_B-Y_A) & -4(Y_A+Y_B) & -2(Y_B-Y_A) \\ 2(Y_B-Y_A) & (Y_A+Y_B) & -2(Y_B-Y_A) & -(Y_A+Y_B) \\ -4(Y_A+Y_B) & -2(Y_B-Y_A) & 4(Y_A+Y_B) & 2(Y_B-Y_A) \\ -2(Y_B-Y_A) & -(Y_A+Y_B) & 2(Y_B-Y_A) & (Y_A+Y_B) \end{bmatrix} \begin{bmatrix} V_{e1} \\ V_{o1} \\ V_{e2} \\ V_{o2} \end{bmatrix} \qquad (3)$$

The 4 × 4 Y-matrix in eqn. 3 leads to the circuit model proposed in Fig. 2, composed by admittances Y_A and Y_B connected to ideal transformers. The new model relevant features are:

(i) The two-port circuit (one multimode CPW input port and one multimode CPW output port) is modelled as a four-port device that separates the CEM and COM contributions into different singlemode ports; i.e. one monomode CPW input port (1) and one monomode CPW output port (3) only propagating the CEM (with characteristic impedance Z_{0e}), and one monomode CPW input port (2) and one monomode CPW output port (4) only propagating the COM (with characteristic impedance Z_{0o}).

(ii) Therefore, planar structures that present different responses to each mode (such as slotline-to-CPW transitions and airbridges) can be easily connected by loading the four ports in Fig. 2 with appropriate impedances. Whenever a mode is cut-off, its effect is also easily included by terminating the appropriate ports in Fig. 2 with reactive impedances.

(iii) It allows a quantitative analysis of the energy transfer from one mode to the other; in particular, whenever $Y_A = Y_B$, it can be seen (from eqn. 3 or Fig. 2) that there is no energy conversion from the CEM to COM and vice versa. Conversely, whenever $Y_A \neq Y_B$, a mode conversion takes place.

(iv) Since the model contains circuit elements, it can be used to analyse or design any application with series impedances in the ground planes, and can be easily implemented in microwave CAD.

Model application and experimental validation: The proposed model has been applied to a 25GHz slotline series resonator, fabricated on CuClad 217, ε_r = 2.17, thickness = 0.254mm (see Fig. 3*a*) and placed in the *E*-plane of a WR-42 rectangular waveguide. The circuit consists of a centred slotline series resonator (length, ℓ_2 = 1.9mm; slot width, 0.15mm) placed in one of the ground planes of a CPW section (length ℓ_1 = 33.53mm; slot width, 0.175; central strip width, 0.15mm). The circuit also includes two slotline-to-CPW transitions at its input and output ports, respectively, in order to excite the circuit with the COM modes (ports 2 and 4 in Fig. 2). The slotline-to-CPW transitions add a small parasitic shunt susceptance for the COM (or slotline mode), but they act as an open circuit for the CEM. Consequently, according to the model proposed (Fig. 2), the circuit is equivalent to a COM transmission line (length ℓ_1) loaded at its centre with two CEM open-ended shunt stubs (length $\ell_2/2$) and two slotline short-ended series stubs (length ℓ_2). Fig. 3*a* and *b* compare its simulated S-parameters, obtained from the model, with the measured S-parameters, showing an excellent agreement (the resonance frequency is exactly predicted), and demonstrating the model's validity and applicability.

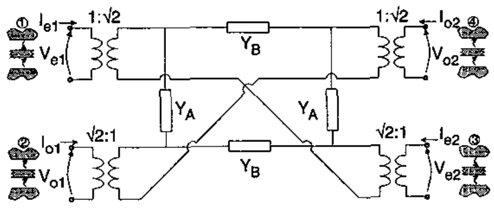

Fig. 2 *Circuit model for asymmetric series impedances in ground planes of CPW*

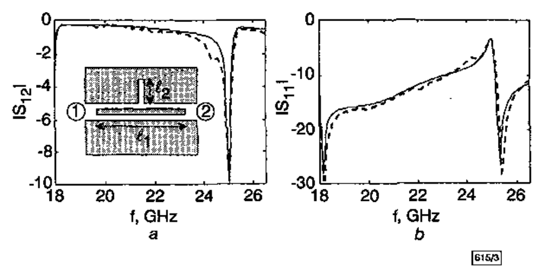

Fig. 3 *Measured and simulated parameters for asymmetric series impedances in ground plane*

——— simulations
– – – – measurements
(*a*) $|S_{21}|$ dB
(*b*) $|S_{11}|$ dB

Conclusions: A new 'circuit model' that allows a quantitative analysis of the energy exchange between even and odd modes at asymmetric series impedances in the ground planes of a CPW has been proposed, and applied to the simulation of a 25GHz microwave slotline resonator. The excellent agreement between simulated results and S-parameter measurements demonstrates the validity of the model and its usefulness as a CAD tool in the design and optimisation of hybrid /MMIC microwave CPW circuits.

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Foam technology for integration of millimetre-wave 3D functions

J.P. Harel, Ch. Person and J.Ph. Coupez

It is demonstrated that shielded foams can be efficiently used as millimetre-wave waveguides. Experimental examples are provided and discussed with respect to standard planar or waveguide techniques in the 26–40GHz frequency range. A 2% bandpass foam-waveguide filter is designed, with an insertion loss of ~ –3.2dB at 28.7GHz.

Introduction: Polymethacrylimide foam (PMI, simply called here 'foam'), is a closed-cell rigid plastic, presenting outstanding mechanical and electrical properties (especially in terms of density, weight, dielectric values and losses). Furthermore, this material can easily be machined or pre-formed by press-moulding or milling techniques, and therefore offers the possibility of creating any geometrical configurations with very simple and cheap procedures.

Even if the usual fields of application for foam are patch antennas and radomes, the use of such a material, with such a low density, low dielectric constant and low loss values ($\varepsilon_r = 1.08$ and $\tan\delta = 0.001$ measured at 60GHz), may be successfully used in a waveguide approach, as a substitute for air. Under this assumption, foam could consequently be used as a propagation media, and also as a mechanical support itself. In the following, we demonstrate that such electrical and mechanical advantages can be exploited to realise very low cost devices, such as waveguides and filters, for operation in the millimetre-wave domain.

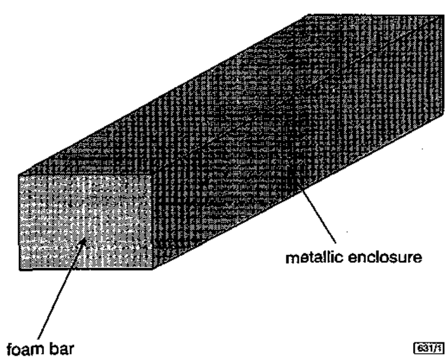

Fig. 1 *Shielded foam bar as waveguide structure*

Realisation and characterisation of foam waveguides: Waveguides are popular because of their low insertion loss, high power handling capabilities and high Q, compared to those of other electromagnetic propagation media such as planar structures (i.e. microstrip and coplanar lines). Unfortunately, this is not a low cost solution, mainly for millimetre-wave applications, due to size constraints.

One major problem in building a waveguide using a simple foam bar is the difficulty of realising the metallic enclosure of the guide. This metallic shielding should have a good electrical performance and should be easily manufactured, for cost reasons. To this end, several tests have been carried out in order to define a simple metallic shielding procedure, and a comparison made with the performance obtained using a standard WR 28 brass guide. All pieces were 100mm long.

Table 1: Measured insertion loss depending on shielding method

Shielding method	Averaged IL
	dB
Standard brass waveguide, air filled	0.25
Standard brass waveguide, foam filled	1.3
Foam bar shielded with thin sticked copper film	1.75
Foam bar painted with silver solution	1.9
Foam bar shielded by press-moulding techniques (copper film and foam sticked together by means of adhesive polypropylene dielectric film)	1.5

The corresponding insertion loss measurements are compared in Table 1. It can be seen that the use of foam implies the addition of a 1.05dB minimum loss per 100mm with respect to a conventional air-filled brass guide. In comparison, a 50Ω 100mm microstrip line realised on an $\varepsilon_r = 2.2$ substrate ($\tan\delta = 10^{-3}$) gives approximately a 2dB insertion loss.

The various waveguide structures presented here are somewhat better than a microstrip transmission line, even if the employed metallic shielding methods are still not optimised. Indeed, the insertion loss should significantly decrease through, for example, an electrochemical metallisation procedure which is an industrially mature technology. Consequently, we conclude that PMI foam can be considered to be a good substitute for standard millimetre-wave brass guides.

Applicationof foam-waveguide filters: Many applications in future millimetre-wave systems (WLL, LMDS,..) will require filters with very narrow frequency bandwidths (< 5%). Of course, those filters will have to have the lowest insertion loss possible and good return loss, which implies high quality factors Q_o. Consequently, only waveguide filters could be used to obtain the better performances required in terms of the available Q_o values in comparison with those possible using planar technology.

We have investigated the possibility of developing 3D bandpass filters with the proposed foam waveguide solution. The objectives were to design an easily reproducible and non-tunable filter to enable mass production, by means of standard press-moulding and etching techniques.

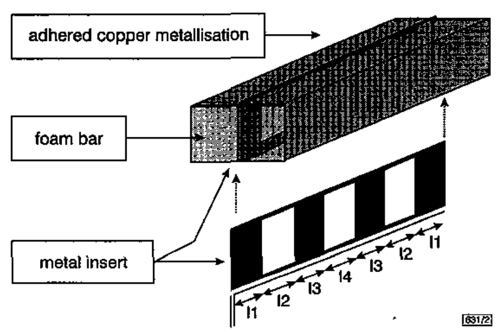

Fig. 2 *E-plane filter using shielded foam guide*

$l_1 = 1.009$mm, $l_2 = 4.778$mm, $l_3 = 3.87$mm and $l_4 = 4.796$mm

We have chosen the 'E-plane metal insert bandpass filter' topology, shown in Fig. 2.

Such a topology is quite easy to synthetise and quite insensitive to technological spread [1]. A third-order E-plane metal insert filter is presented in Fig. 2 which is designed to operate in the Ka band. The metal insert thickness was chosen to be $t = 0.254$mm; the device was patterned by photoetching techniques. A WR 28