

Modeling of Spurious Coupling Between Modes in Metal Packages and Embedded Circuits

T. Bolz* S. Held† M. Neinhüs† A. Beyer† and K. Solbach†

*IMST GmbH

47475 Kamp-Lintfort, Germany, Email: bolz@imst.de

†University of Duisburg-Essen

47057 Duisburg, Germany, Email: hft@uni-due.de

Abstract—The consideration of electromagnetic interaction between spurious modes in metal packages and its enclosed microwave circuit (MMIC) in a circuit simulator is an important topic in circuit design. In this paper a novel method is described to determine mutual coupling networks for incorporation of this parasitic coupling into a SPICE based circuit simulator. For simple package shapes the coupling networks can be determined analytically, whereas full-wave simulations are needed for more complex ones. To shorten the duration, time-domain simulators need for analyzing a high-Q structure, Prony's method is applied to extract circuit parameters from short time sequences.

Index Terms—Electromagnetic coupling, Coupling circuits, Coupled mode analysis

I. INTRODUCTION

Microwave circuit design at 10 GHz and beyond, nowadays is accomplished by circuit simulation software. The whole circuit is being divided into passive structures and active electronic devices and then analyzed separately. The components are characterized in terms of matrix representations (S-, Y- or Z-matrix), equivalent circuits or mathematical equations. Complex passive structures are normally analyzed by full-wave simulators. Then all components are recomposed and analyzed using linear microwave theory. The advantage of using a circuit simulator for analyzing MMICs is the speed with which parameter studies can be performed.

Parasitic effects, i.e. radiation, mutual coupling between adjacent elements, shielding issues, become more dominant with increasing frequency, but are not considered in the circuit models. Therefore full-wave simulators are more appropriate for analyzing the passive structure of the circuit. The disadvantage associated with full-wave analysis is the long simulation time for large structures and many parameter variations.

Normally MMICs are embedded in a metal enclosure to protect them from external fields and to avoid radiation. Surface currents and fringing fields of the microstrip structure can excite package modes, which will influence the proper operation of the MMIC. To reduce unnecessary technology cycles at development time, it is important to include these effects in the circuit simulator.

This paper should provide the microwave designer with physical insight into coupling effects between cavity modes and the embedded microwave circuit. Furthermore a procedure is introduced, which includes these coupling effects in terms of a coupling network into the circuit simulator.

II. THEORY

The proposed method is designed on the basis of the works [1], [2], [3]. In [1], the interaction of cavity modes to electrical surface current densities is described in terms of an impedance coupling network. An analytical description of the network is missing and the method is only valid for cuboid cavities.

In [2], the coupling is described by equivalent magnetic current line sources along the periphery of the microstrip structure in conjunction with the modal expansion approach [4] for a metal box. The inhomogeneous material distribution is approximated by an effective permittivity.

An extension of [2] is proposed in [3], where a full-wave simulator is used to determine a matrix, describing the mutual coupling between the magnetic current density and modes in cavities with more complex cross-sections. Our method differs from the techniques presented in [1], [2], [3] in terms of analytical circuit parameter determination and application of the full-wave solver. First an analytical derivation of the coupling networks and associated coupling matrices for simple cross section geometries and multi layer structures is described. Based on this analytical approach, coupling networks for cavities with arbitrary cross section geometry can be derived from only one full wave simulation in the time-domain (e.g. TLM or FDTD).

The full-wave analysis of high-Q structures is generally a time consuming task, because the observation quantity decays very slowly. Hence, Prony's method [6] is applied to yield short simulation times, even for highly resonant cavities.

A. Coupling Networks

In contrast to [2], the transmission line model depicted in Fig. 1 is used to analytically derive the coupling networks. For this purpose, the transversal fields of the TE_{mn} -

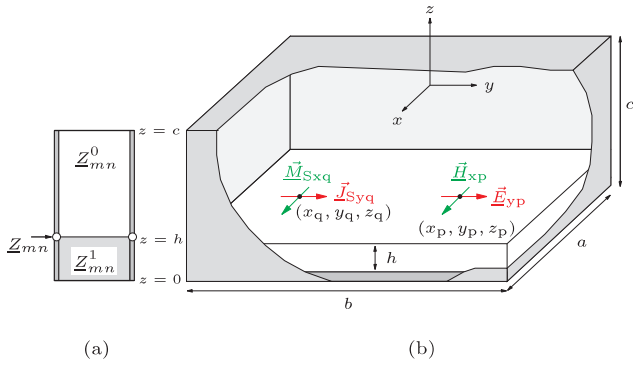


Fig. 1. (a) Transmission line model for determination of Z_{mn} and Y_{mn} . (b) Electric and magnetic current sources and excited fields.

and TM_{mn} -modes at the surface of the layer, where the microstrip circuit is located, are determined first. Using the modal expansion approach for homogeneous waveguide structures [5], the electrical field (here as an example the y-component) at the observation point (x_p, y_p) for TE_{mn} - and TM_{mn} -modes can be expressed as (Fig. 1):

$$E_{y,mn}(x_p, y_p) = -Z_{mn} \cdot i \cdot dl_q \cdot t_{E_{y,mn}}(x_p, y_p) \cdot t_{E_{y,mn}}(x_q, y_q), \quad (1)$$

if the electric current density

$$\vec{J}_{S q} = i \cdot dl_q \cdot \delta(x - x_q) \cdot \delta(y - y_q) \cdot \vec{e}_y \quad (2)$$

is excited at the source point (x_q, y_q) . Using the duality principle, the magnetic field (here as an example the x-component) at the observation point (x_p, y_p) may be expressed as:

$$H_{x,mn}(x_p, y_p) = -Y_{mn} \cdot v \cdot dl_q \cdot t_{H_{x,mn}}(x_p, y_p) \cdot t_{H_{x,mn}}(x_q, y_q), \quad (3)$$

if the magnetic current density

$$\vec{M}_{S q} = v \cdot dl_q \cdot \delta(x - x_q) \cdot \delta(y - y_q) \cdot \vec{e}_x \quad (4)$$

is excited at source point (x_q, y_q) . Due to the excitation in Eq. 2 and 4, the field components E and H (Eq. 1 and 3) are called Green's Functions. Z_{mn} and Y_{mn} are the input impedances and input admittances, respectively, of the transmission line model describing the configuration of Fig. 1. The functions $t_{E_{y,mn}}$ and $t_{H_{x,mn}}$ describe the transversal field distributions of the TE_{mn} - and TM_{mn} -modes.

The electromagnetic coupling between a cavity mode and the electric surface current density on the metalization of the microwave structure is given by the impedance coupling matrix $Z_{c,mn}$:

$$Z_{c,mn} = \begin{pmatrix} z_{11,mn}^2 & z_{12,mn} & \cdots & z_{1N,mn} \\ z_{21,mn} & z_{22,mn}^2 & \cdots & z_{2N,mn} \\ \vdots & \vdots & \ddots & \vdots \\ z_{N1,mn} & z_{N2,mn} & \cdots & z_{NN,mn}^2 \end{pmatrix}. \quad (5)$$

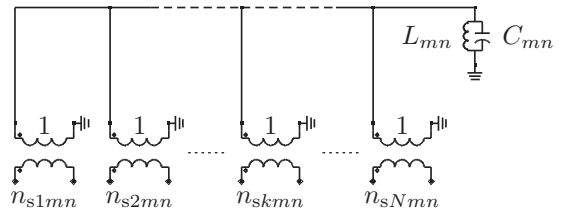


Fig. 2. Impedance coupling network representing Eq. 5.

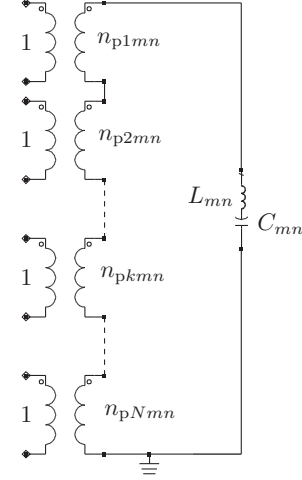


Fig. 3. Admittance coupling network representing Eq. 6

For the dual configuration, the coupling between a cavity mode and the magnetic surface current density corresponding to the fringing fields at the periphery of the microwave structure is given by the admittance coupling matrix $Y_{c,mn}$:

$$Y_{c,mn} = \begin{pmatrix} y_{11,mn}^2 & y_{12,mn} & \cdots & y_{1N,mn} \\ y_{21,mn} & y_{22,mn}^2 & \cdots & y_{2N,mn} \\ \vdots & \vdots & \ddots & \vdots \\ y_{N1,mn} & y_{N2,mn} & \cdots & y_{NN,mn}^2 \end{pmatrix}. \quad (6)$$

The matrix elements are determined using the reciprocity theorem:

$$z_{pq} = -\frac{1}{i_p i_q^*} \cdot \iint_A \vec{E}_t^* \cdot \vec{J}_{S p} dA \quad (7)$$

and

$$y_{pq} = -\frac{1}{v_p v_q^*} \cdot \iint_A \vec{H}_t^* \cdot \vec{M}_{S p} dA \quad (8)$$

with \vec{E}_t and \vec{H}_t , the transversal fields at the surface of the layer, where the microstrip circuit is located. Electrical equivalent circuits for one mode representing Eq. 5 and 6 are given in Fig. 2 and 3, respectively. In a further improvement with regard to [2], magnetic line currents are also considered at the input and output ports of the microstrip configuration. At multi mode operation, the impedance coupling networks are to be connected in series

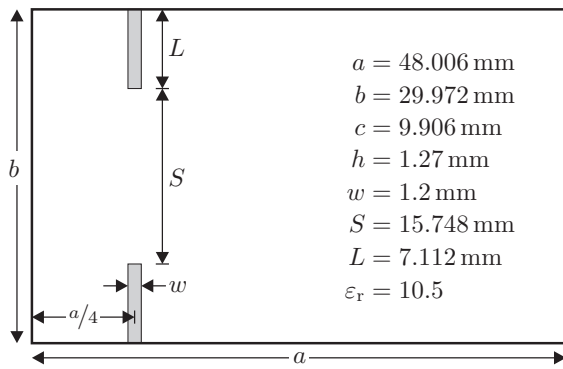


Fig. 4. Geometry of gap inside a rectangular cavity

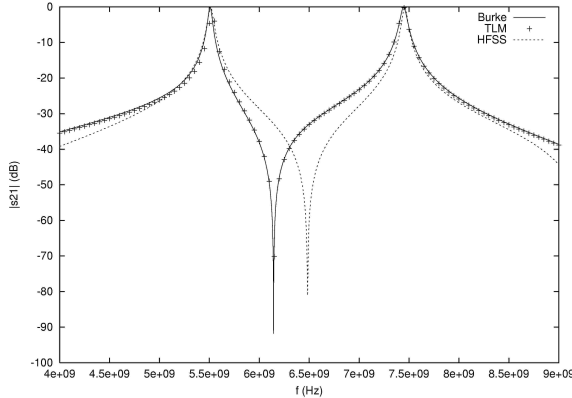


Fig. 5. S_{21} result of Fig. 4: Strong interaction with TE_{10} - and TM_{11} -mode

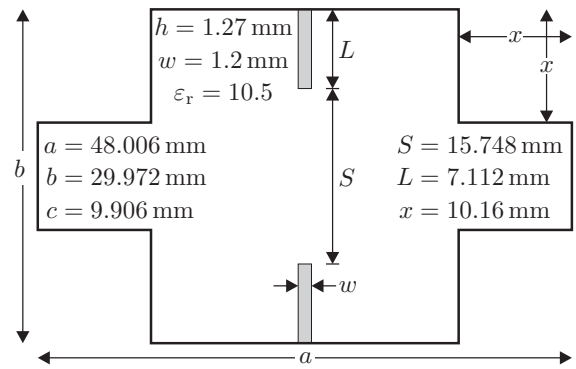


Fig. 6. Geometry of gap inside a cavity with complex cross section

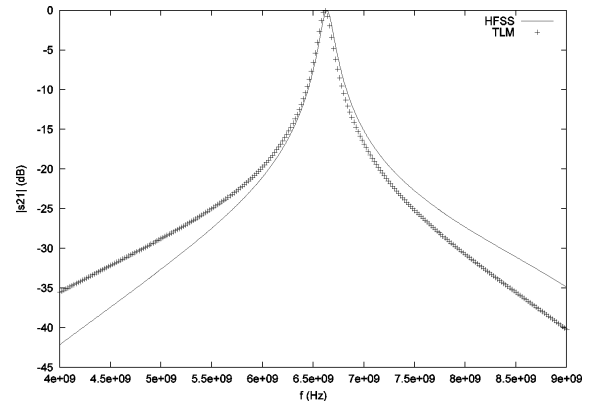


Fig. 7. S_{21} result of Fig. 6: Strong interaction with TM_{11} -mode

and the admittance coupling networks are to be connected in parallel.

III. RESULTS

A. Electric Current Modeling

For the examples shown in Fig. 4 and 6 the interaction of the electric current density on the metalization of the gap structure with cavity modes are shown by means of the transmission coefficient S_{21} for TE_{10} - and TM_{11} -modes in Fig. 5 and TM_{11} in Fig. 7, respectively. The corresponding coupling network for the structure in Fig. 4 was determined analytically, whereas the complex package (Fig. 6) is analyzed by a TLM simulator and post processed by applying singular value decomposition (SVD) to extract the number of resonant modes and Prony's Method to determine the coupling matrix $Z_{c,11}$ (Eq. 5). The derived coupling network accurately models the interaction of cavity modes and the microstrip structure, as can be seen from the transmission coefficient S_{21} in Fig. 5 and 7.

B. Magnetic Current Modeling

The example shown in Fig. 8 is calculated using the magnetic current modeling method. The microstrip structure is substituted by a planar multiport network model

[2]. At the position of the ports along the periphery of the microstrip lines, equivalent magnetic current densities are introduced, from which the coupling network in Fig. 3 is derived.

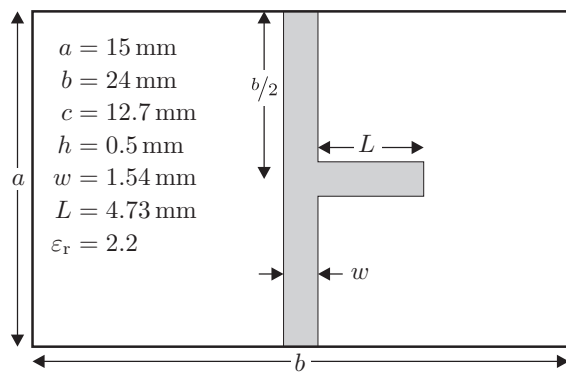
The transmission coefficient S_{21} for a single open stub circuit is depicted in Fig. 9. Again, the comparison with a full-wave simulation shows a good agreement between the different calculation schemes. Furthermore, the magnetic current approach is able to handle arbitrary shaped microstrip structures by using the segmentation and desegmentation techniques, or the contour integral method [7].

IV. CONCLUSION

It has been shown that spurious mode coupling within a metal package can be conveniently modeled by simple equivalent circuits. For simple cavity geometries, they can be evaluated analytically, whereas for complex ones a full-wave simulation is needed.

Several examples were presented and their results for the electric current model and the dual method, the magnetic current model, agree well with full-wave simulations.

The advantage of the proposed method is the inclusion of coupling effects into the circuit simulator. Thus



[7] T. Okoshi and T. Miyoshi, "The planar circuit - an approach to microwave integrated circuitry," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 20, pp. 245-252, Apr. 1972.

Fig. 8. Geometry of open stub centered inside a rectangular cavity

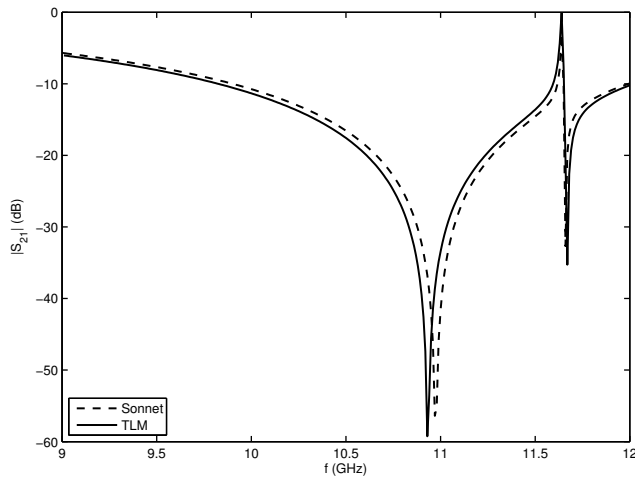


Fig. 9. S_{21} result of Fig. 8: Strong interaction with TM_{11} -mode

parameter studies can be performed more quickly than by using a full-wave solver. In addition complex cross section packages (Fig. 6) can easily be evaluated by the proposed method.

REFERENCES

- [1] J. J. Burke and R. W. Jackson, "A simple circuit model for resonant mode coupling in packages MMICs," in *Microwave Symposium Digest, 1991*, *IEEE MTT-S International*, Boston, MA, June 1991, pp. 1221-1224.
- [2] A. Sabban and K. C. Gupta, "Effects of packages on parasitic coupling among microstrip discontinuities in MMICs," *International Journal Of Microwave Millimeter-Wave Computer-Aided Engineering*, vol. 1, no. 4, pp. 403-411, 1991.
- [3] H. Cebi and K. C. Gupta, "Effect of package shape on spurious coupling among microstrip discontinuities," in *Microwave Symposium Digest, 1996*, *IEEE MTT-S International*, vol. 3, San Francisco, CA, June 1996, pp. 1823-1826.
- [4] R. F. Harrington, *Time-Harmonic Electromagnetic Fields*. McGraw-Hill, Inc., 1961.
- [5] R. E. Collin, *Field Theory of Guided Waves, 2nd Edition*. Wiley-IEEE Press, 1990.
- [6] W. L. Ko and R. Mittra, "A combination of FD-TD and prony's methods for analyzing microwave integrated circuits," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 39, pp. 2176-2181, Dec. 1991.