Removing Package Effects From Microstrip Moment Method Calculations

by

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Abstract

The electromagnetic behavior of microstrip circuits often depends on the package that surrounds the circuit. This is especially true if the package is electrically large. This paper discusses the importance of a damped enclosure for modeling and operating MMIC circuits. It suggests a method of damping enclosures and outlines an efficient technique for applying the method of moments so as to determine the significance of the package effects on full wave simulations.

Introduction

In CAD of an MMIC, an electromagnetic simulator is often used to characterize smaller pieces of the over-all circuit. Usually the simulator analyzes the piece in an analysis enclosure which is considerably smaller than the final operating enclosure (the package). This is partially to enhance numerical efficiency since the scale size difference between the large operating enclosure and the small circuit often results in very long spectral summations. A large analysis enclosure also introduces a number of enclosure resonances which may not have any relation to the circuit's behavior in the, presumably different, operating enclosure. Finally, a large analysis enclosure may introduce a strong frequency dependence which makes an efficient frequency interpolation difficult to implement. However, the trend is to simulate larger and more complicated circuit pieces and thus eventually the analysis enclosure will become large enough that enclosure resonances must occur. Identifying their effect on the circuit will be important.

A properly designed package should have only a small effect on the circuit. In particular, the effect of resonant modes in either the analysis enclosure or the operating enclosure is reduced by lowering the Q of both enclosures[1]. This is commonly done in practical circuit packaging, but not in numerical modeling of such circuits. It makes the enclosure look like a sort of "anechoic" chamber to the electromagnetic fields inside and results in the circuit's operating characteristics being less dependent on position. It also reduces resonant coupling between circuit elements. It is difficult to know when the enclosure's Q is reduced sufficiently for a particular application.

In this paper we investigate using a lossy damping layer to reduce the influence of resonance effects on circuitry in analysis or operating enclosures. The generic structure in the interior of an enclosure is shown in Figure 1. It consists of a perfectly conducting ground plane, a dielectric layer supporting the microstrip circuit, a free space layer, a lossy substrate, and a perfectly conducting top cover. To investigate the package influence, we introduce vertical walls by using a new method of applying the spectral domain method of moments to enclosed microstrip circuits[1]. In the next section, the technique is briefly outlined. A more detailed description can be found elsewhere [2]. The characteristics of a lossy damping layer are then discussed in terms of a trade-off between enclosure Q and circuit Q. A doped silicon substrate is assumed for the damping layer since it is inexpensive, plentiful, and well characterized. Finally, we demonstrate the resulting reduced sensitivity of circuit performance to enclosure resonances.

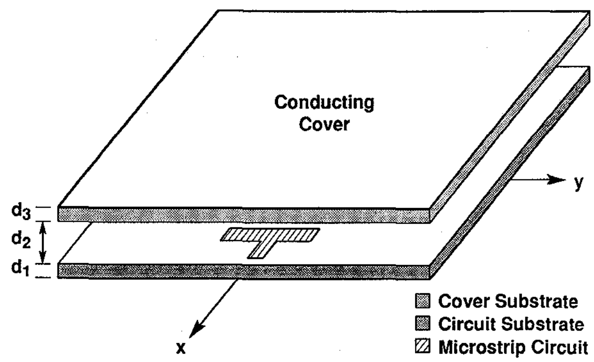

Figure 1. Layered structure in the interior of an MMIC enclosure. Cover substrate damps enclosure resonances.

Mutual Impedance Modification Based on Side Wall Images

A fundamental part of the moment method analysis of microstrip circuits involves computation of the mutual impedance between two surface currents. A representative mutual impedance between two x directed currents is for example,

$$Z_{ij}^{xx} = -\int\int dx\, dy\, J_{xi}(x,y)[E_x(x,y)]_j \quad (1a)$$

$$= -\int\int dx\, dy\, J_{xi}(x,y) \cdot \int\int dx'dy'\, G_{xx}(x,x',y,y')\, J_{xj}(x',y') \quad (1b)$$

where the Green's function, G_{xx}, can be written in terms of TM_z or TE_z parts according to,

$$G_{xx}(x,x',y,y') = \frac{\partial^2}{\partial x^2} G^{TM}(x,x',y,y') + \frac{\partial^2}{\partial y^2} G^{TE}(x,x',y,y') \quad (2)$$

For an open structure the TM or TE Green's functions, G^{TM} and G^{TE}, can be expressed in spectral form as

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$$G^{TV}(x,x',y,y') = \frac{1}{2\pi} \int \int dk_x\, dk_y \cdot \frac{Q_{TV}(\beta)}{\beta^2} e^{jk_x(x-x')} e^{jk_y(y-y')} \qquad (3)$$

where V is M or E, $\beta^2 = k_x^2 + k_y^2$. $Q_{TV}(\beta)$ is listed in reference [4]. The impedance described by (1) includes all sources of coupling between the i'th and j'th currents. For an open structure, these sources may include reactive fields, radiation, and surface waves. In the covered structures discussed in this paper the only sources of coupling are reactive fields and the TM_0 parallel plate wave. Although expression (3) is written as though no sidewalls are present (k_x,k_y continuous), it can also be used to derive the effects of side walls. In what follows, we show that the method of moments mutual impedance between two currents in an enclosure can be calculated by first calculating the mutual impedance in a laterally open structure and then adding a modifying term which includes the effect of the enclosure side walls. The modifying term is easily computed and makes use of the approximation described next.

In many cases of interest it is desirable to have the mutual impedance between two current elements that are not very closely spaced. It then becomes convenient to rewrite (3) in terms of sums of TM and TE parallel plate waves as derived in what follows. Changing the integral in (3) to polar form yields

$$G^{TV}(\rho) = \frac{1}{2\pi} \int_{-\infty}^{\infty} d\beta \frac{Q_{TV}(\beta)}{2\beta} H_0^{(2)}(\beta\rho) \qquad (4)$$

where $\rho^2 = |x-x'|^2 + |y-y'|^2$. Q_{TM} and Q_{TE} have sets of poles which correspond to the TM and TE parallel plate mode propagation constants, β_n^{TM} and β_n^{TE}.

By closing a contour integration around the lower half of the complex β plane, equation (4) becomes

$$G^{TV}(\rho) = -j \sum_n \mathrm{Res}\left[Q_{TV}(\beta_n^{TV})\right] H_0^{(2)}(\beta_n^{TV}\rho)/2\beta_n^{TV} \qquad (5)$$

where $\mathrm{Res}[Q(\beta_n)]$ denotes the residue of Q at β_n.

In the MMIC environment with layering such as in Figure 1, usually only the lowest order parallel plate mode, the TM_0 mode, is propagating while all the other modes are evanescent. Thus, for relatively large ρ, only the TM_0 term in (5) is important and

$$E_x(x,y) \approx \frac{-j}{2\beta_0^{TM}} \mathrm{Res}\left[Q_{TM}(\beta_0^{TM})\right] \cdot \frac{\partial^2}{\partial x^2} \int dx'dy'\; J_x(x',y')\; H_0^{(2)}(\beta_0^{TM} |\vec{\rho} - \vec{\rho}\,'|) \qquad (6)$$

For smaller ρ the reactive fields of the evanescent modes are important and more terms in (5) must be included.

Expression (6) can be used to evaluate equation (1) in cases where the i'th and j'th currents are well separated. [1],[3]. It is generally the case that the subsectional basis functions (and even groups of basis functions) used in moment method solutions are small enough that $\Delta x\, \beta_0^{TM} << 1$ where it is assumed that Δx is the scale length of the basis function. As a result equation (1) can be written

$$Z_{ij}^{xx} \approx \frac{-j}{2\beta_0^{TM}} \mathrm{Res}\left[Q_{TM}(\beta_0^{TM})\right] \cdot \frac{\partial^2}{\partial x_i^2} H_0^{(2)}(\beta_0^{TM} |\vec{\rho}_i - \vec{\rho}_j|) \langle J_{xi}\rangle \langle J_{xj}\rangle \qquad (7)$$

where $\langle J_{xi}\rangle = \int \int dx\, dy\, J_{xi}(x,y)$ and $\vec{\rho}_k$ locates the center of the k'th current.

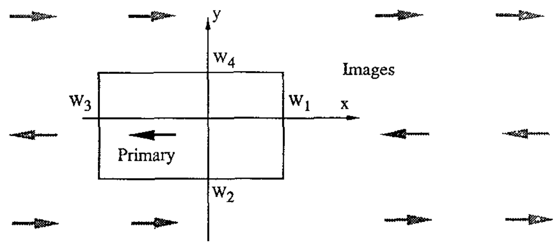

Figure 2. Schematic drawing of enclosed current and a few of the infinite set of image currents.

Now consider the mutual impedance between two enclosed currents. Figure 2 shows a top view of Figure 1 with a primary source current and the infinite set of images introduced when perfectly conducting side walls are located at x = W_1 and W_3 and at y = W_2 and W_4. The electric field at a point can be divided into two components, $E_x = E_x^{pri} + E_x^{img}$. The first component is the field resulting from the primary source current in a laterally open structure and the other results from the infinite set of image sources which occur due to the presence of side walls. The field point is often close to, or on, the primary source current, thus all the reactive terms in equation (5) (or equation (4)) are often necessary in calculating the field from the primary source current. In most cases however, the image sources are far enough from the field point that only the TM_0 mode radiating from each image current has any bearing on E_x^{img}. This field is thus simpler to compute. The simplicity is helpful since, in a large enclosure, the image contribution carries much of the frequency dependence of E_x and often must be computed at many points in the bandwidth of interest.

Splitting the field into two parts leads to splitting the mutual impedance into two parts,

$$Z_{ij}^{xx} = -\int \int dx\, dy\, f_i(x) g_i(y) \left[E_x^{pri} + E_x^{img}\right] \qquad (8a)$$

$$= Z_{ij}^{pri} + Z_{ij}^{img} \qquad (8b)$$

where

$$Z_{ij}^{xxpri} = -\int \int dxdy\, f_i(x) g_i(y) \int \int dx'dy'\, f_j(x') g_j(y') G_{xx} \qquad (8c)$$

$$Z_{ij}^{xximg} \approx -\langle J_{xi}\rangle \langle J_{xj}\rangle \left(\frac{-j}{2\beta_0^{TM}}\right) \mathrm{Res}\left[Q_{TM}(\beta_0^{TM})\right] \bullet \frac{\partial^2}{\partial x_i^2} \sum_{\alpha,\beta} s_\beta H_0^{(2)}\left(\beta_0^{TM} \sqrt{(x_i - x_\alpha)^2 + (y_i - y_\beta)^2}\right) \qquad (8d)$$

where (8d) is calculated using equation (7). The double sum in (8d) sums over the infinite grid of image sources . The variable s_β equals + or - 1 depending on the appropriate image sign. Summing (8d) as shown is very inefficient ; however, it can be shown that writing this summation in a spectral form results in a single sum which converges within ten or fifteen terms. Space limitations preclude a description of this form here, but details can be found in reference [2]. The mutual impedance between the primary currents, Z_{ij}^{pri}, is calculated by inserting (2) and (3) into (8c) exactly as if the currents were located in a laterally open structure. When the currents are located close to one another a spectral integration is necessary. Z_{ij}^{pri} is slowly varying with respect to frequency and thus its value over a broad range of frequencies can be evaluated by computing only a few points in a band and then interpolating to find Z_{ij}^{pri} at all other frequencies. In addition, Z_{ij}^{pri} depends only on the relative location of i,j and this leads to a number of useful redundancies when a uniform grid is used. The image contribution, Z_{ij}^{img}, does not depend upon the details of the i'th or j'th current element, only on their average values. It can be calculated very quickly, but is very frequency dependent especially if the enclosure is large, has many resonances, and is high Q.

The concept described in the previous section separates the current from its enclosure in as far as calculating the moment method matrix is concerned. It works very well for covered circuits located in the interior of the package, not too near a side wall. This is the situation for most of the circuitry on a large integrated circuit. However, if a gap generator on a side wall is to be used to excite a circuit, a modification in procedure must be made before the enclosure and the circuit can be separated. This is due to the fact that the currents excited by the generator are on or near one side wall. Thus their lowest order images in the side wall are close enough that all the image reactive fields contribute to the field evaluated at a near interior point. The second order images in the near side wall and the images in the other three side walls are still far enough away so that only the TM_0 wave communicates their contribution. Calculating the mutual impedance between two laterally enclosed near wall currents is again viewed as testing the field due to a primary current source and all of its images. As before, these sources are divided into two groups except that the near wall image source is included in the full spectral calculation of Z_{ij}^{pri} and removed from the calculation of Z_{ij}^{img}.

Combining the impedance calculations for both of the aforementioned groups results in the method of moments matrix for the circuit in an enclosure. Figure 3 shows the real part of the input impedance seen by a gap generator which drives a stub of length 0.775mm and a width of 0.2mm extending from the W_1 wall. The gap generator source is located at x = 0, y = 0 and the enclosure walls are located at $W_1 = 0$, $W_2 = -5.0$ mm, $W_3 = -20.0$ mm, $W_4 = 7.0$ mm. The layering in this example is specified to be; $d_1 = 0.1$ mm, $d_2 = 0.3$ mm, $d_3 = 0.1$ mm, $\varepsilon_{r1} = 12.7$, $\varepsilon_{r2} = 1.0$, and $\varepsilon_{r3} = 12(1-j)$. Applying the MoM using only Z_{ij}^{pri} (including the near image) yields an impedance response corresponding to a configuration without walls 2, 3, and 4. In contrast, applying the MoM using $Z_{ij}^{pri} + Z_{ij}^{img}$ yields the fully enclosed impedance response. The enclosure introduces several variations in the impedance which are not present without the enclosure. These are due to package resonances -- about seven of them in the frequency range shown in the figure. The results using the conventional technique [4] are also shown for comparison. Excellent agreement is apparent.

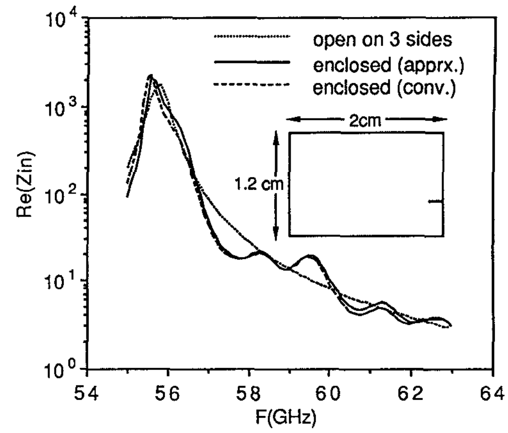

Figure 3. Input resistance of an open circuit stub protruding from a side wall.

Damped Package Considerations

A doped silicon layer is suggested as an alternative to the lossy film proposed in[5]. Doped silicon is very well characterized, inexpensive, and has often been used with GaAs MMICs without any contaminating influence. In addition, it presents a lossy volume to the enclosure instead of a lossy surface and as a result should be effective in enclosure Q reduction. Enclosures of interest to MMIC designers would typically have large lateral dimensions and small heights. In such cases resonances with differing mode numbers but similar resonant frequencies have roughly the same Q s. Figure 4 shows the package Q s at various resonant frequencies for various cover heights and substrate thicknesses. In all cases the doping level in the Si substrate is chosen such that the skin depth of the substrate is equal to the substrate thickness at the center of the band of interest, about 55 GHz in this case. A lower doping results in less damping while higher doping starts to make the substrate to look like a metal with an attendant increase in enclosure Q. As expected, the figure shows that when the lossy substrate takes up a higher percentage of enclosure volume, the enclosure Q s are smaller.

The trade-off involved with an increase in the relative volume of the lossy substrate is that losses are introduced in the circuit on the GaAs substrate below. To examine this effect we consider the admittance of a half wavelength linear resonator which is driven by a gap generator at the center and which resonates near 57 GHz. The Q of the resonator admittance

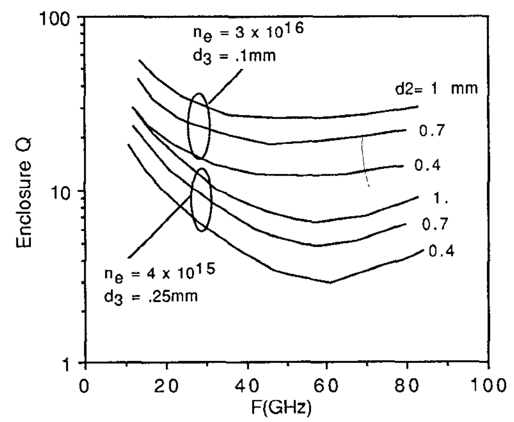

Figure 4. Enclosure Q s for a 2 cm x 1.2 cm enclosure with a .1mm GaAs circuit substrate and a doped Si cover substrate.

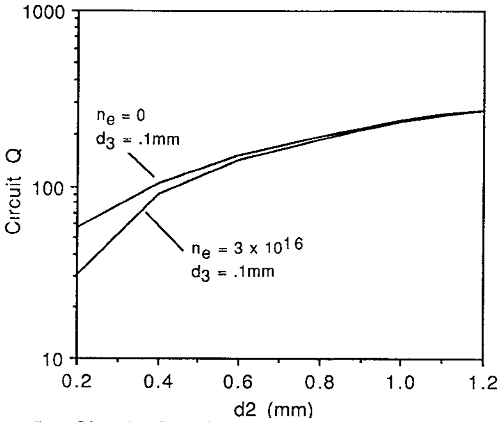

Figure 5. Circuit Q s for a .775 mm x .2mm center fed microstrip resonator on a .1mm GaAs substrate in a laterally open structure with a doped Si cover substrate.

will be affected by the existence of the upper substrate. Microstrip conductor losses are not included. To divorce the circuit Q from the enclosure resonances, we examine the resonator admittance in a laterally open environment -- all four side walls removed. Two sources of loss contribute to reduced circuit Q, (1) reactive field coupling to the lossy substrate and (2) parallel plate waves which radiate energy away from the resonator. The latter are often the most important, even, as we shall see, in an enclosed environment. Figure 5 shows a plot of the circuit Q versus the distance d_2 as determined from the full wave simulation of the admittance response. For the undoped case the only Q reducing loss is due to radiation of parallel plate waves. When doping is introduced, circuit Q is further reduced due to reactive field interactions with the carriers. However, the reactive field loss is significant only for very small d_2. The curves shown apply to the 0.1mm Si layer, but the Q variation for a 0.25mm Si layer with $n_e = 4 \times 10^{15}$ is almost exactly the same.

It is important to note that the parallel plate radiation has an important effect when the substrate is laterally enclosed. If the enclosure is undamped, the radiated waves reflect off the sidewalls and, when a resonance occurs, recombine and reinforce each other at various points in the enclosure. Significant coupling occurs and the circuit is very package sensitive. In a properly damped enclosure, the radiated waves are damped as thev approach and reflect off the side walls and resonance reinforced coupling is much reduced.

Figure 4 indicates that increasing d_2 increases enclosure Q and thus has a deleterious effect on enclosure resonance damping. On the other hand, Figure 5 shows that increasing d_2 improves circuit Q. However, in most non superconducting circuits microstrip conductor losses will keep the resonator Q well below 200 at 50 GHz. With this in mind, Figure 5 indicates that there is no circuit loss advantage to increasing d_2 above about 1mm. Thus a d_2 of roughly 1mm is optimum for both the thicker and thinner Si substrates. Figure 4 shows that a lower enclosure Q results for the thicker substrate and thus this would result in the least sensitive enclosure.

Damped Enclosure Effect On Circuit Behavior

Figures 6 show resonator responses with and without a lateral enclosure for the two different lossy layer choices. The full wave side-wall image method described previously was used to insert and remove the side wall effect. The real part of the admittance seen by a gap generator located at the center of the resonator. The resonator has the dimensions described in the previous section. It is centered at {x,y} = {0,0} with the wall locations as described in the figure legends.

The results show that the thicker less lossy substrate (Figure 6b) results in an enclosure which has minimal effect on behavior of the circuit. The resonator gives the same response in one location as it does in the other or as it does when the substrate is not laterally enclosed. Figure 6a shows that the thinner more lossy substrate is less effective in eliminating the circuit's location dependence.

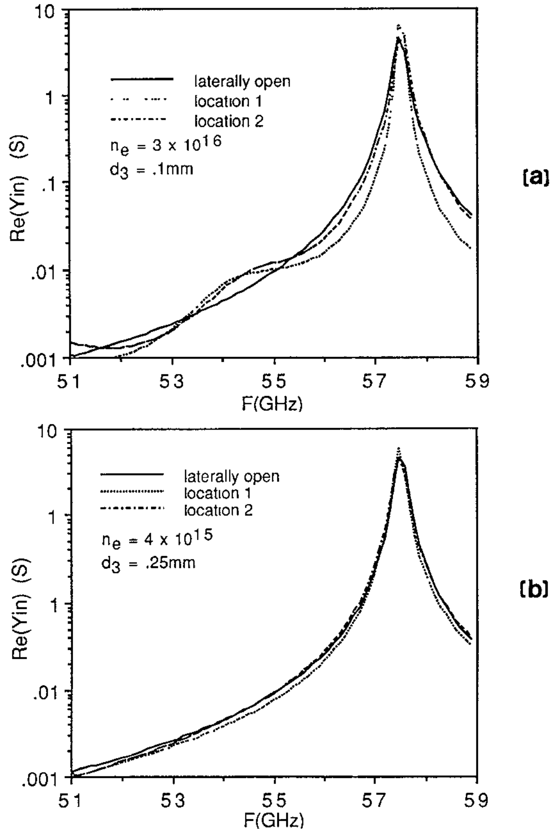

Figure 6. Input conductance for a resonator located at (1) W_1 = 6.0mm, W_2 = -3.3, W_3 = 14.0, W_4 = 8.7, or located at (2) W_1 = 3.0mm, W_2 = -5.3, W_3 = 17.0, W_4 = 6.7, or unenclosed. Plots (a) and (b) are for the thin and thick Si.

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