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REDUCTION OF PARASITIC COUPLING IN PACKAGED MMIC'S

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Abstract

Electromagnetic coupling between circuit elements within a package is often significant, or even catastrophic, at frequencies near a package resonance. In future, large, highly integrated, millimeter-wave MMICs this resonant coupling will be difficult to avoid. The addition of lossy materials to the enclosure will reduce the coupling, but not eliminate it. In this paper, two different methods of introducing this loss will be compared. Results indicate that a further reduction in the coupling of power to the resonant modes is possible by repositioning the circuit within the enclosure.

Introduction

In the design of MMIC circuits, the effect of enclosing the circuit in a package is usually assumed to be negligible. This has not been a very serious problem in current designs because circuit packages have not been electrically large. As the level of integration and/or the operating frequency increases, the electrical size of the enclosure will also increase. For sufficiently large enclosures, package resonances are possible. If the system operates at frequencies near one of these resonances, catastrophic coupling can occur between different elements of a circuit. A common technique for reducing the coupling to a resonant mode is to place a microwave absorber on the cover. Williams [1] suggests using a dielectric substrate coated with a resistive film as an inexpensive alternative to the microwave absorber. In either case, the presence of loss has a very significant effect on circuit behavior at frequencies near an enclosure resonance.

A number of full-wave techniques for modeling microstrip circuits in an enclosure have been published [2,3,4]. However, none of these specifically study coupling of power to resonant modes. Armstrong and Cooper [5] experimentally investigated the use of microwave absorbing materials to suppress the coupling to resonant modes. Jansen and Wiemer [6] developed circuit models for a few microstrip discontinuities in an enclosure, but their use when a lossy material is present has not been verified.

In this paper, a moment method formulation is presented which models microstrip circuits in a lossy enclosure using rooftop currents. A Green's function is developed that includes the effect of lossy material on the cover. The coupling of power to a resonant mode as a function of circuit location in an enclosure is investigated. The absorbing layer will be compared to the resistive film with respect to their effectiveness in reducing resonance effects. It is shown that the coupling of power to a resonant mode can be reduced further by repositioning certain circuit features within the enclosure.

Theory

A rigorous application of the method of moments is used to determine the relative amount of power coupled from a microstrip circuit in an enclosure to a resonant mode. Figure 1 shows the geometry of the MMIC package under consideration. There are three dielectric layers of thickness d_1, d_2, and d_3. The lower substrate will support the MMIC circuitry. The upper substrate, in general, is a layer of lossy material (microwave absorber) coated with a thin resistive film. The surface impedance of the resistive film is designated by Z_s. The walls of the enclosure are assumed to be perfect conductors.

Green's Function

A Green's function has been developed that relates the tangential electric field at the $z=d_1$ surface to the Fourier series coefficient of the surface current on the same surface,

$$E_u(x,y) = \sum_{n=0}^{\infty}\sum_{m=0}^{\infty} \frac{4\varepsilon_n\varepsilon_m}{ab}\{Q_{ux}(k_{xn},k_{ym})J_x(k_{xn},k_{ym}) + Q_{uy}(k_{xn},k_{ym})J_y(k_{xn},k_{ym})\}T_u(x,y) \qquad (1)$$

where

$$J_u(k_{xn},k_{ym}) = \iint J_u(x,y)T_u(x,y)dxdy \qquad (2)$$

$$T_x(x,y) = \cos(k_{xn}x)\sin(k_{ym}y) \quad , \quad T_y(x,y) = \sin(k_{xn}x)\cos(k_{ym}y)$$

$$\varepsilon_k = \begin{cases} 0.5 & k=0 \\ 1.0 & k\neq 0 \end{cases} \quad , \quad u = x \text{ or } y$$

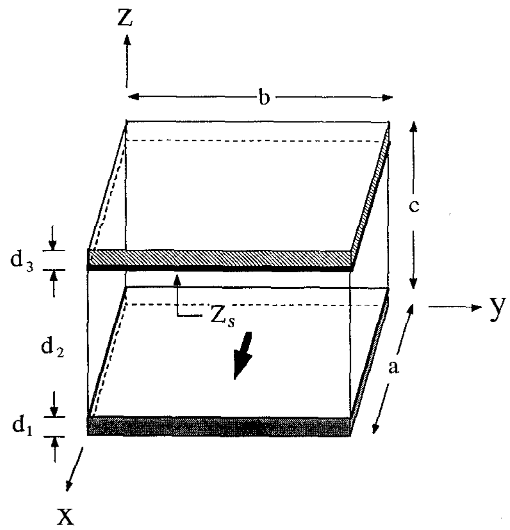

Fig. 1 Geometry used in the derivation of the Green's function.

The derivation of $Q_{pq}(k_{xn},k_{ym})$ is similar to the method in [7] , therefore only the results are given below:

$$Q_{xx}(k_{xn},k_{ym}) = -\left(\frac{k_{xn}^2}{k_\rho^2}Q_{TM} + \frac{k_{yn}^2}{k_\rho^2}Q_{TE}\right) \quad (3a)$$

$$Q_{xy}(k_{xn},k_{ym}) = Q_{yx}(k_{xn},k_{ym}) = -\frac{k_{xn}k_{ym}}{k_\rho^2}(Q_{TM} - Q_{TE}) \quad (3b)$$

$$Q_{yy}(k_{xn},k_{ym}) = -\left(\frac{k_{yn}^2}{k_\rho^2}Q_{TM} + \frac{k_{xn}^2}{k_\rho^2}Q_{TE}\right) \quad (3c)$$

where

$$Q_{TV} = \frac{1}{Y_{LV}^{(1)} + Y_{LV}^{(2)}} \quad (4)$$

$$Y_{LV}^{(1)} = -jY_{TV}^{(1)}\cot(k_{z1}d_1) \quad , \quad Y_{RV}^{(3)} = -jY_{TV}^{(3)}\cot(k_{z3}d_3)$$

$$Y_{RV}^{(2)} = Y_{TV}^{(2)}\frac{\left[Y_s + Y_{RV}^{(3)}\right] + jY_{TV}^{(2)}\tan(k_{z2}d_2)}{Y_{TV}^{(2)} + j\left[Y_s + Y_{RV}^{(3)}\right]\tan(k_{z2}d2)}$$

$$k_{zi} = \sqrt{\varepsilon_{ri}\mu_{ri}k_0^2 - k_\rho^2}$$

$$Y_{TM}^{(i)} = \frac{\varepsilon_{ri}k_0}{k_{zi}\eta_0} \quad , \quad Y_{TE}^{(i)} = \frac{k_{zi}}{\mu_{ri}k_0\eta_0}$$

$$k_\rho^2 = k_{xn}^2 + k_{ym}^2 \quad , \quad k_{xn} = \frac{n\pi}{a} \quad , \quad k_{ym} = \frac{m\pi}{b}$$

V=E or M , i=1, 2, or 3, and $Y_s = 1/Z_s$.

Moment Method Solution

The current, $\vec{J}_s$, is expanded as follows,

$$\vec{J}_s(x,y) = \hat{x}\sum_{i=1}^{N_x} I_{xi}J_{xi}(x,y) + \hat{y}\sum_{j=1}^{N_y} I_{yj}J_{yj}(x,y) \quad (5)$$

Rooftop functions are used to expand the current in the areas of discontinuities [8] and at the connectors. An x-directed rooftop current has the form,

$$J_{xi}(x,y) = t(x - x_i)p(y - y_i) \quad (6)$$

where $t(x)$ is a triangular function of length $2\Delta_{xi}$ and $p(y)$ is a pulse function of length Δ_{yi}. The points (x_i,y_i) are the midpoints of the x-directed currents. The x-directed currents overlap each other in the x-direction but not in the y-direction. For a y-directed current, x and y are interchanged in the above expressions. In uniform sections of transmission lines, currents with an edge condition are used. For a section of transmission line in the x-direction, these currents have the form

$$J_{xi}(x,y) = t(x - x_i)f(y - y_i) \quad (7a)$$

$$J_{yj}(x,y) = p(x - x_j)g(y - y_j) \quad (7b)$$

where $f(x)$ and $g(y)$ are functions that satisfy the edge condition [7].

Applying Galerkin's method [2,3] results in a set of algebraic equations for the such that

$$[\mathbf{Z}][\mathbf{I}] = [\mathbf{V}] \quad (8)$$

where [**V**] is the excitation vector (see reference [2]) and [**I**] is the unknown current vector comprised of the complex coefficients I_{xi} and I_{yj}. After solving for the unknown currents, the scattering parameters are then calculated.

Mode Power

After determining the unknown current coefficients, the power lost to the enclosure is calculated. This power is given by [9]

$$P_{\text{loss}} = -\frac{1}{2}\text{Re}\iint \vec{E}_t(x,y) \bullet \vec{J}_s^*(x,y)dxdy \quad (9)$$

Near a resonance of the enclosure, P_{loss} is approximately equal to

$$P_{\text{mode}} = -\frac{1}{2}\text{Re}\iint \vec{E}_{\text{mode}}(x,y) \bullet \vec{J}_s^*(x,y) \quad (10)$$

where P_{mode} is the power lost to the resonant mode that corresponds to the resonance. In the following analysis we will discuss only the TM mode, but the analysis is easily extended to the TE mode.

For the TM mode, the tangential electric field is given by

$$E_{TM}^x(x,y) = A_{TM}^x\cos(k_{xn}x)\sin(k_{ym}y) \quad (11a)$$

$$E_{TM}^y(x,y) = A_{TM}^y\sin(k_{xn}x)\cos(k_{ym}y) \quad (11b)$$

where n and m correspond to the indices of the resonant mode. Substituting (5) and (11) into (10) yields

$$P_{\text{mode}} = P_{TM} = P_{TM}^x + P_{TM}^y \quad (12a)$$

$$P_{TM}^x = \sum_{p=1}^{N_x} P_{TM}^{xp} \quad (12b)$$

$$P_{TM}^y = \sum_{p=1}^{N_y} P_{TM}^{yp} \quad (12c)$$

$$P_{TM}^{xp} = -\frac{1}{2}\text{Re}\left\{\iint E_{TM}^x(x,y)I_{xp}^*J_{xp}(x,y)dxdy\right\} \quad (12d)$$

$$P_{TM}^{yp} = -\frac{1}{2}\text{Re}\left\{\iint E_{TM}^y(x,y)I_{yp}^*J_{yp}(x,y)dxdy\right\} \quad (12e)$$

Inspection of (11) and (12) indicates that P_{TM} can reduced by locating areas of high current in areas of low electric field and visa-versa.

Consider a circuit where the current is predominately x-directed and centered at y=y_c, it can be shown that P_{loss} as a function of y_c, is given by

$$P_{\text{loss}} \approx \left(P_{TM}^x\right)_{\max}\sin^2(k_{ym}y_c) \quad (13)$$

and thus P_{loss} can be reduced by changing y_c. For example, positioning the circuit in the center of the box will result in a maximum power loss for the TM_{110} ($k_{y1} = \pi/b$) resonance and a minimum power loss for the TM_{120} ($k_{y2} = 2\pi/b$) resonance.

Results

The full-wave analysis described in the previous section was used to analyze a large gap in a transmission line and a transmission line with a single shunt open circuit stub attached midway between the input and output connectors. The scattering parameters were determined as well as the power lost to the resonant modes of the enclosure.

Large Gap

Consider a cavity of the following dimensions: a=30 mm, b=48 mm and c=12.7 mm (Fig. 1). The substrate thickness is d_1=1.27 mm and the relative permittivity is ε_{r1}=10.5(1-j0.0023). In the band 4-9 GHz, the TM_{110} (5.6 GHz) and the TM_{120} (7.5 GHz) are resonant in such an enclosure. Note that this cavity scales (excluding loss) to a reasonably sized enclosure for an MMIC operating in the band 40-90 GHz.

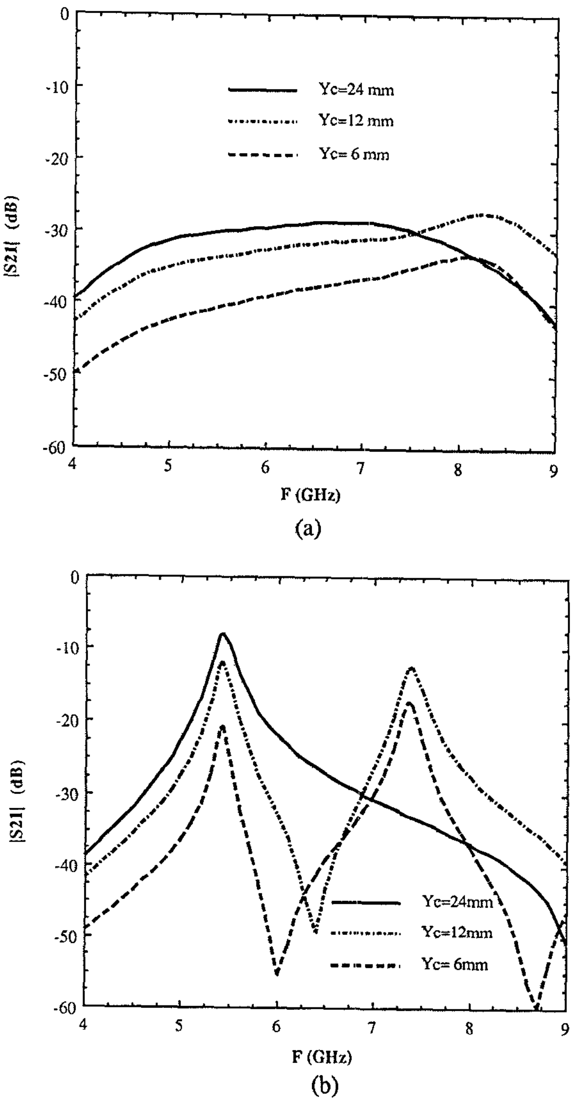

Fig. 2 Computed $|S_{21}|$ for a large gap in a transmission line enclosed in (a) cavity A and (b) cavity B with dimesions a=30 mm, b=48 mm, c=12.7 mm, w=1.2 mm and g=15 mm.

To reduce the coupling of power to resonant modes, two methods are employed to lower the Q. The first method is to fix a microwave absorbing layer to the cover of the enclosure. The thickness of this layer is d_3=1.27 mm, the relative permittivity is ε_{r3}=21.15(1-j0.04) and the relative permeability is μ_{r3}=2.3(1-j0.88). This cavity will be defined to be cavity A. The second method used to decrease coupling of power to resonant modes was recommended by Williams [1]. A low loss dielectric substrate coated with a resistive film was placed on the cover. The thickness of this substrate is d_3=1.27 mm with a relative permittivity of ε_{r3}=10.5(1-j0.0023). The optimal surface impedance, Z_s, was determined to be 35+j35 Ω/sq [1]. This cavity will be defined to be cavity B.

A transmission line, centered at y=y_c, with a large gap, g=15mm, was analyzed. Ideally, a very low transmission should occur. Fig. 2a shows the predicted transmission response of cavity A for three different values of y_c. A 10 dB improvement in $|S_{21}|$ is gained for the TM_{110} resonance by placing the circuit at y_c=6 mm (b/8) instead of at y_c=24 mm (b/2).

Fig. 2b shows the predicted transmission response of cavity B for three different values of y_c. For this circuit, the resonance effect from the TM_{110} and TM_{120} mode was not sufficiently reduced using the method suggested by Williams [1].

Eq (13) predicts that the maximum power will couple to the TM_{110} mode ($k_{y1} = \pi/b$) by placing the circuit at y_c=24 mm (b/2) and that no power will couple to the TM_{120} mode ($k_{y2} = 2\pi/b$) for the circuit located at y_c=24 mm (b/2). The full-wave analysis supports this prediction.

Stub

Consider a cavity of the following dimensions: a=15 mm, b=24 mm and c=12.7 mm (Fig. 1). The substrate thickness is d_1=1.27mm and the relative permittivity is ε_{r1}=10.5(1-j0.0023). For an enclosure of this size, there is only one resonant mode, the TM_{110} (10.8 GHz), in the band 9-12 GHz. To reduce the coupling of power to the TM_{110} mode, a microwave absorbing layer was attached to the cover of the enclosure. The thickness of this layer is d_3=1.27 mm, the relative permittivity is ε_{r3}=21(1-j0.02) and the relative permeability is μ_{r3}=1.1(1-j1.4).

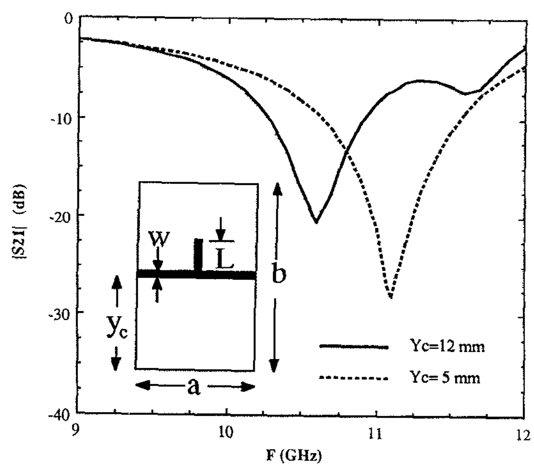

Fig. 3 Computed $|S_{21}|$ for a transmission line with a single shunt open circuit stub having dimensions; a=15 mm, b=24 mm, c=12.7 mm, w=1.4 mm and L=1.9 mm.

A transmission line with a single shunt open circuit stub attached was analyzed. The stub is centered at x=7.5 mm (a/2) and has a length of 1.9 mm. The transmission line is centered at $y=y_c$. Fig. 3 shows the predicted transmission response of the stub for two different values of y_c. Note that the stub resonance and the box resonance interact significantly when the circuit is located at y_c=12 mm (b/2), but they do not interact when the circuit is located at y_c=5 mm.

The dominate currents on the transmission line and on the stub are, respectively, x and y-directed. Thus they radiate power into the resonant mode via different field components. As a result, moving the circuit will change the power lost from the stub in a different way than it changes the power lost from the transmission line. Inspection of equations (11a) and (12d) indicates that moving the transmission line from y_c=12 mm to y_c=5 mm will reduce P_{TM}^x. An exact calculation of P_{TM}^x shown in Table 1 verifies this. On the other hand, the same shift increases P_{TM}^y. Inspection of equations (11b) and (12e) show that this occurs due to the increases in the E_y field component of the mode near the side wall. The net power lost to the mode, P_{TM}^x plus P_{TM}^y, decreases as the circuit is moved toward the side wall. The third column in Table 1 is the total power lost including all fields and is defined to be $1-|S_{11}|^2-|S_{21}|^2$. Note that when the circuit strongly couples to the TM_{110} mode, the total loss is almost equal to P_{TM}^x plus P_{TM}^y. For weak coupling the total loss is not dominated by the resonant mode loss.

Fig. 4 is a plot of the mode power loss contributed by each subsection of current along the transmission line. This corresponds to P_{TM}^{xi} in equation (12b). For y_c=12 mm a large power loss occurs on the transmission line at x=3 mm. This occurs because a large current standing wave peak is present at that point and because the x-directed mode field is reasonably large there. Moving this standing wave peak to y_c=5 mm moves it into a low field region and thus a decrease in mode power loss occurs.

To verify the analysis of the stub (Fig. 3), both stub circuits were fabricated using duroid RT6010 (DK10.5). Measurements agree well with the theory.

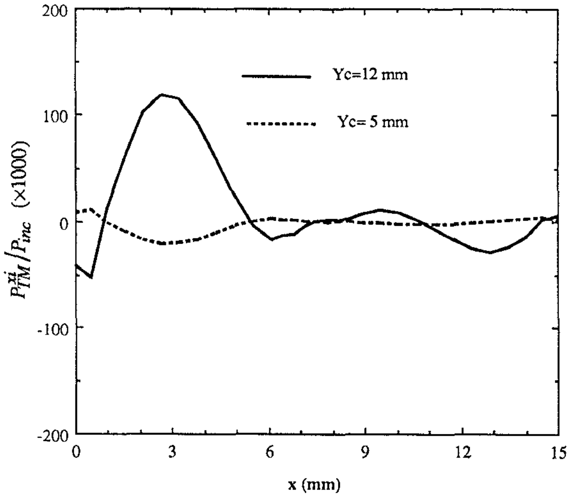

Fig. 4 Fraction of the incident power, P_{inc}, converted to P_{TM}^{xi} versus x at 11.5 GHz

y_c(mm)	P_{TM}^x/P_{inc}	P_{TM}^y/P_{inc}	Loss
5	-0.064	0.1135	0.136
12	0.371	0.0432	0.473

Table 1 The ratio of the TM_{110} mode powers to the incident power for the stub.

Conclusion

The analysis shows that package resonances can have a very significant effect on circuit operation even at frequencies which are not very close to resonance. These effects, can be reduced by including lossy material in the enclosure.

A further reduction in the coupling of power to resonant modes can be obtained by repositioning the circuit. Locating areas of high current in areas of low electric field in the enclosure reduces the power lost to these resonant modes. This principle would find most application in moderately sized enclosures where the resonant frequencies are not closely spaced and the mode field structure is relatively simple.

Acknowledgement

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