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SPICE-Compatible Cavity and Transmission Line Model for Power Bus with Narrow Slots

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***Abstract*— Segmental lumped circuits are derived from coupled transmission line model for a narrow slot on the power bus. Both electric and magnetic coupling are taken into account by distributed inductances and capacitances. Then a SPICE-compatible circuit model for the power bus with the narrow slot is proposed. In this model, the segmental lumped circuits are connected to the equivalent circuit, which is derived by a hybrid cavity model and segmentation method for irregular power/ground planes. The model is validated by comparing with the calculations of finite element method (FEM) for the self or mutual impedances of the two port networks located in the power bus.**

I. Introduction

Power bus noise is becoming one of the major power integrity (PI) concerns due to high digital logic switching rate and layer transition of high-speed signal traces[1]. Decoupling capacitors are usually used to reduce the impedance of the power/ground planes, and thus stabilize the power supply voltage in high-speed, high trace density package or PCB design [2]-[5]. Power/ground plane segmentation or power islands are another popular technique to isolate the power noise [6][7]. On the other hand, various digital logic levels result in different voltage power supplies. Mixed digital and analog circuit design also requires separation of digital power/ground planes and its analog counterparts. This naturally results in some slots on power/ground pairs. Therefore, an analysis of a power bus with slots or a splitting are crucial in high-speed PCB or package designs.

Power bus with slots has been analyzed by a number of rigorous numerical methods. Mixed-potential integral equation method for multilayered structures was adopted in [8]. the hybrid finite element and the method of moments (MoM) is also used for gapped power bus structures [9]. While these numerical solutions are very flexible and accurate, they are relatively time-consuming, and thus not very friendly for engineering applications. A SPICE-compatible model is more preferable to a digital design engineer than full-wave approaches. For this reason, cavity model has been extensively used in power bus analysis, as it can lead to an equivalent circuit consisting of lumped resistance, capacitance and inductance [10][11]. With the aid of the segmentation method, the cavity model can be extended to more general shaped power buses [12]. Recently, several authors proposed lumped circuits to model the coupling effects along two sides of a slot [14],[15]. Comparing with the full-wave approaches, lumped circuit model provides more insights on the physics of coupling and usually takes less computation time.

In this paper, a SPICE-compatible model is proposed by using a resonant cavity model and coupled transmission lines representing the regular power/ground plane and slots, respectively. Moreover, the segmentation method is employed to make the cavity model capable of simulating more general cavity structures. To employ this model, the power bus with a narrow slot is first divided into several large regular portions and a small part along the slot. The regular portion of the power bus can be modelled as resonant cavities, and their impedance properties are easily calculated via summation of the corresponding cavity modes [10]. On the other hand, the slot is viewed as a coupled microstrip line [16]. Through the even and odd mode analysis, the distributed inductances and capacitances are extracted and expressed by analytical formulas related to the structure's geometry and dielectric properties.

It is worthy to note that this research work is motivated by the previous paper of one of the co-authors [16]. In that paper, the main concept of the transmission line model is proposed. The present paper contains the detailization of implementation. Moreover, a new point of view on the drawback of the model is provided, and an improvement path is suggested for the further development.

II. Formulation of Cavity and Distributed Circuit Model

A schematic of a rectangular power/ground plane with a narrow slot is shown in Fig. 1. The dimensions are depicted. Between the power and ground plane there is a lossy dielectric material with relative permittivity ε_r and loss tangent $\tan\delta$. Two ports are located at the points (x_1, y_1) and (x_2, y_2), respectively.

To calculate the self and mutual impedance at the ports 1 and 2, the power bus is divided into three rectangular cavities A, B and C, and one coupled microstrip line, as shown in Fig.1 by dashed lines. In the segmentation method, many internal ports are set along the connection boundary of two cavities. Each port is connected directly with its counterpart in another cavity to assure the continuity of tangential fields along the boundary. While cavity-to-cavity connection has

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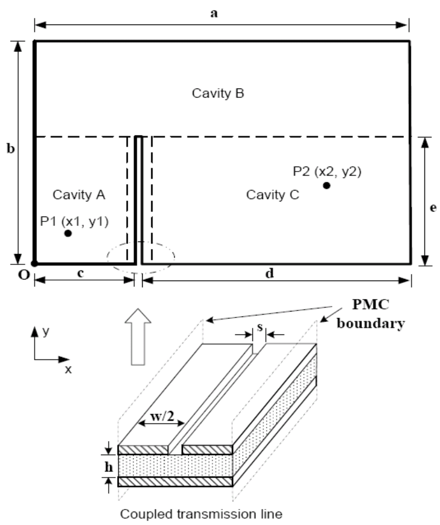


Fig. 1. The example structure analyzed

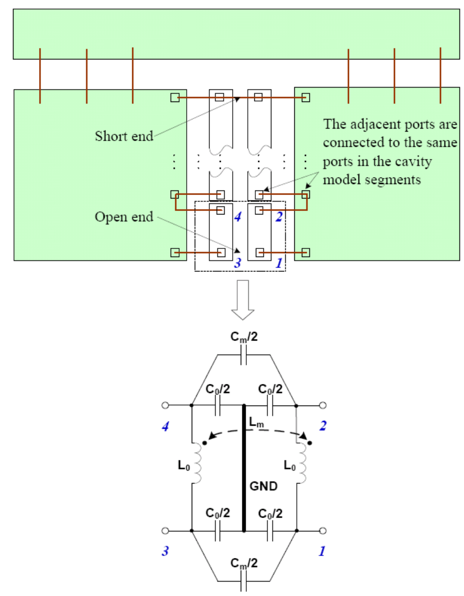


Fig. 2. Equivalent four port circuit for one segment of the slot.

been solidly set up by the segmentation method, careful consideration needs to be undertaken for cavity-transmission connections.

In Fig. 9 of [16], the cavity ports are put alone the edge of the coupled transmission line while the transmission line ports are along the center of the transmission lines. As cavity ports and transmission line ports are not located at same physical positions, it's not correct to connect them directly to enforce boundary conditions on these ports with different locations.

To provide a more reliable foundation for the hybrid model, a new point of view is desirable. Fig. 2 presents an illustration of the model proposed in this paper. The dashed line in Fig. 1 is exactly same to the dashed line in Fig.3. This means we put the ports of cavity boundary exactly same to the transmission line center. The voltages and currents of these ports are obtained by either the cavity model or the coupled transmission lines. This naturally satisfies the boundary condition along the dashed lines.

To derive segmental lumped circuits, the slot is first divided into small segments and then each segment is represented by a four-port lumped circuit, as is shown in Fig. 2. All the equivalent circuits of segments are connected by sharing the same ports along the slot. Therefore, the dashed line is the boundary between cavity model and transmission model in the present approach. For example, ports 3 and 4 are related together by either the cavity model or equivalent circuit derived from the coupled transmission lines. This satisfies the continuity of tangential electric and magnetic field along the slash line (boundary of cavity and lumped circuit models). An equivalent circuit is steadily available for a cavity model. Thus, a SPICE-compatible circuit model is derived by connecting all cavities and the distributed LC circuit of the slot together.

Since the equivalent circuits of cavity models have been studied for many years, [10]-[13], this paper is focused on the derivation of the lumped circuit model of the slot.

The geometric scheme and its equivalent circuit are illustrated in Fig. 2. The mutual electric and magnetic coupling are represented by the capacitance and inductance in the circuit, respectively.

These parameters are derived from the per unit length (p.u.l) parameters of even and odd modes of coupled microstrip lines, as is shown in Fig. 3.

The even mode capacitances are obtained as ([17], Chapter 8.5)

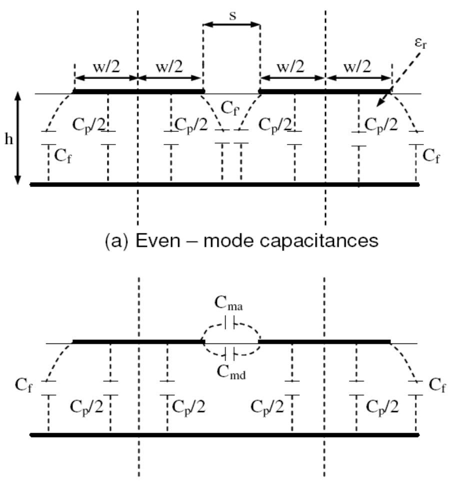
$$\begin{aligned} C_p &= \varepsilon_0 \varepsilon_r \frac{w}{h}; && (1) \\ C_f &= \frac{\sqrt{\varepsilon_{eff}}}{2cZ_0} - \frac{C_p}{2}; && (2) \\ C_f' &= \frac{C_f}{1+(Ah/s)\tanh(8s/h)} \sqrt{\frac{\varepsilon_r}{\varepsilon_{eff}}}, && (3) \end{aligned}$$

where $A = \exp\{-0.1 \exp(2.33 - 2.53w/h)$ and the effective permittivity is ([17], Chapter 4.5)

$$\begin{aligned} \varepsilon_{eff} &= \frac{\varepsilon_r+1}{2}\left[1+\frac{29.98}{Z_0}\left(\frac{2}{\varepsilon_r+1}\right)\left(\frac{\varepsilon_r-1}{\varepsilon_r+1}\right)\right. \\ &\qquad \left.\left(\ln\frac{\pi}{2}+\frac{1}{\varepsilon_r}\ln\frac{4}{\pi}\right)\right]^{-2} \quad (4) \end{aligned}$$

The characteristic impedance of microstrip line is expressed as ([17], Chapter 4.5)

$$Z_0 = \frac{119.9}{\sqrt{2(\varepsilon_r+1)}} \ln\left[4\frac{h}{w} + \sqrt{16\left(\frac{h}{w}\right)^2 + 2}\right] \quad (5)$$



(a) Even – mode capacitances

(b) Odd-mode capacitances.

Fig. 3. Illustration of p.u.l. parameters for even and odd modes of a coupled microstrip line

for $w/h < 3.3$, and

$$
\begin{aligned}
Z_0 &= \frac{119.9\pi}{2\sqrt{\varepsilon_r}} \left\{ \frac{w}{2h} + \frac{\ln 4}{\pi} + \frac{\ln(e\pi^2/16)}{2\pi} \left(\frac{\varepsilon_r - 1}{\varepsilon_r^2} \right) \right. \\
&\quad \left. + \frac{\varepsilon_r + 1}{2\pi\varepsilon_r} \left[\ln \frac{\pi e}{2} + \ln \left(\frac{w}{2h} + 0.94 \right) \right] \right\} \qquad (6)
\end{aligned}
$$

for $w/h > 3.3$. The coupling capacitances through upper air of the odd mode can be evaluated as

$$
C_{ma} = \begin{cases} \frac{\varepsilon_0}{\pi} \ln \left[2(1+\sqrt{k'})/(1-\sqrt{k'}) \right] & 0 \le k^2 \le 0.5; \\ \varepsilon_0 \pi / \ln \left[2(1+\sqrt{k})/(1-\sqrt{k}) \right] & 0.5 \le k^2 \le 1.0, \end{cases} \qquad (7)
$$

where

$$
k = \frac{s/h}{s/h + 2w/h}
$$

and $k' = \sqrt{1-k^2}$, and the coupling capacitance through dielectric substrate is

$$
C_{md} = \frac{\varepsilon_0 \varepsilon_r}{\pi} \ln \left\{ \coth \left(\frac{\pi s}{4h} \right) \right\} + 0.65 C_f \left(\frac{0.02}{s/h} \sqrt{\varepsilon_r} + 1 - \varepsilon_r^{-2} \right) \qquad (8)
$$

Since only the slot region between two dotted lines in Fig.3 is necessary to be considered, the total capacitances for each mode can then be written as

$$
\begin{aligned}
C_e &= C_p/2 + C_f'; \qquad (9) \\
C_o &= C_p/2 + C_{ma} + C_{md}. \qquad (10)
\end{aligned}
$$

Then the inductances and capacitances of one segment's circuit in Fig. 2 can be calculated as

TABLE I

GEOMETRIC PARAMETERS AS SHOWN IN FIG. 1 USED IN THE NUMERICAL EXAMPLE (UNIT:CM)

a	11.1
b	7.0
c	3.0
d	8.0
e	4.0
s	0.1
h	0.2
w	0.6
(x_1, y_1)	(1.0,1.0)
(x_2, y_2)	(8.6, 2.5)

$$
\begin{aligned}
L_o &= \frac{\mu_0 \varepsilon_0}{2} \left(\frac{1}{C_e^a} + \frac{1}{C_0^a} \right) \qquad (11) \\
L_m &= \frac{\mu_0 \varepsilon_0}{2} \left(\frac{1}{C_e^a} - \frac{1}{C_e^a} \right) \qquad (12) \\
C_o &= \frac{1}{2} \left[C_o^a + C_e^a \right] \qquad (13) \\
C_m &= \frac{1}{2} \left[C_o^d - C_e^d \right] \qquad (14)
\end{aligned}
$$

Superscripts 'a' or 'd' represent 'air' ($\varepsilon_r = 1$) or 'dielectrics' (actual ε_r) for the ε_r in (9) and (10).

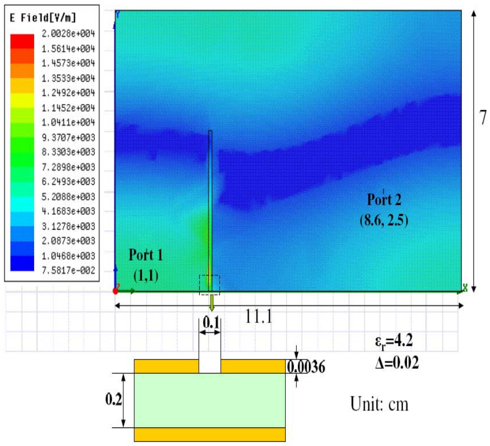
III. NUMERICAL EXAMPLE AND DISCUSSIONS

Table I gives the dimensions of the power bus with a narrow slot and the locations of two ports. Between the power and ground plane is the dielectrics with permittivity $\varepsilon_r = 4.2$ and tangent loss $\tan\delta = 0.02$.

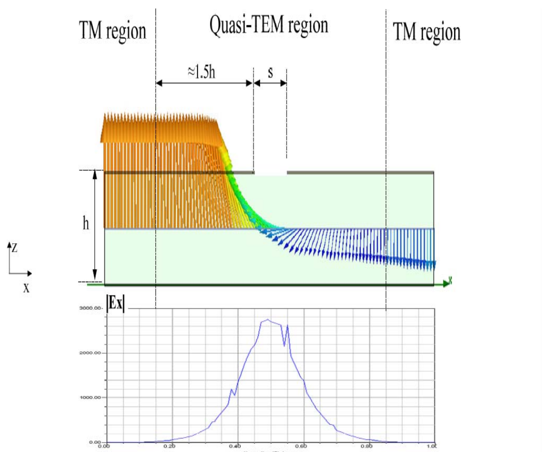
To illustrate the coupling of two sides along the slot, the field distribution obtained by FEM at 2.5GHz is provided. Fig. 4 (b) shows that the electric field across the narrow slot is quite strong. So The electric field at the slot cannot be neglected. This means there is a strong capacitance coupling along the slot. The cavity model alone cannot capture this phenomenon. To some extent, this justifies the new hybrid model proposed in this paper.

Fig. 5 compares the simulation results of the self and mutual impedances between two ports, obtained using three different approaches: finite element method (FEM) by Ansoft's HFSS, 'cavity model' and cavity transmission line model (cavity+TXL). Although the FEM is not a SPICE-compatible approach, it is a rigorous full-wave solution which serves as a benchmark herein.

Legend 'Cavity model' means that the coupling effects of the slot are totally neglected. For narrow slot ($s/h < 5$), the results obtained by cavity model have obvious differences from those of FEM. Even the first resonant frequency cannot be correctly predicted. On the other hand, the 'cavity+TXL' model takes into account both inductance and capacitance coupling of the slot. Therefore, as seen from Fig. 5, the results for both self and mutual impedances obtained by 'cavity+TXL' model agree very well with those obtained by the FEM up to at least 1.5 GHz. Moreover, 'cavity+TXL' takes about 9 minutes



(a)



(b)

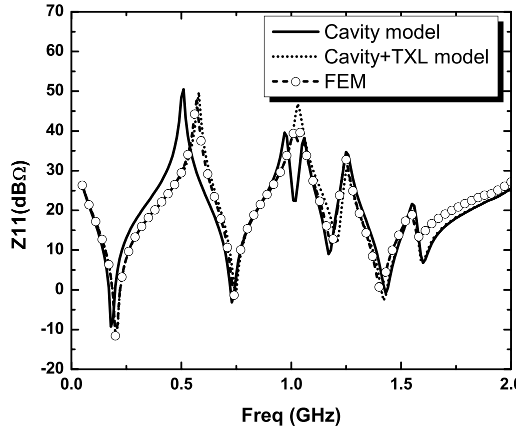
Fig. 4. (a) Electric field distribution in a power bus with a slot. (b) The field distribution near the slot region.

CPU time while the FEM computations take 19 minutes in the same computer. This example demonstrates the effectiveness of the combined cavity and transmission line model for power bus with narrow slots.

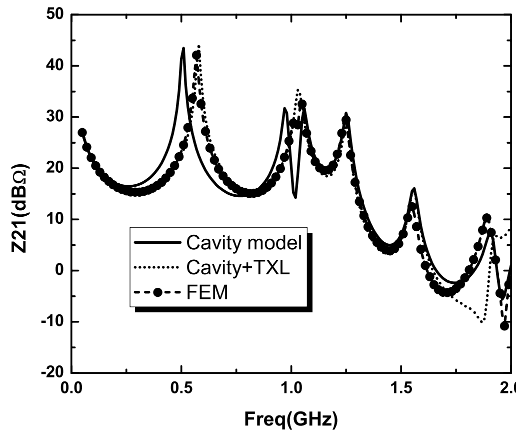
From our new point of view, the equivalent circuit associated with the slot only considers couplings among nearby ports. While this assumption is valid for low frequencies, it neglects the coupling of those ports of separated segments along the slot at higher frequencies. That might be the reason for the eventually enlarged discrepancy between 'Cavity+TXL' method and the FEM results for frequencies above 1.5 GHz. More accurate model is expected to include the coupling of all ports along the slot.

IV. Conclusion

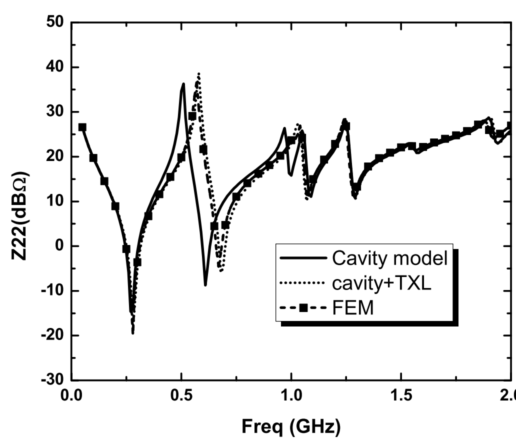
A combined cavity and transmission line model is proposed to analyze the self and mutual impedance of a power bus with narrow slots. The final model is an equivalent circuit which is



(a)



(b)



(c)

Fig. 5. Comparison of self- and mutual impedance among different models.

SPICE-compatible. The coupling effects of the narrow slot are taken into account by even and odd mode of coupled transmission line. A numerical example shows that this combined model predicts accurately with the results obtained by finite element method up to 1.5 GHz, while only the cavity model fails to capture the lowest resonant frequency. The new model is helpful in high-speed PCB or packaging design. It is expected that more accurate model is required to include the coupling of all ports along the slot.

ACKNOWLEDGMENT

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