行政院國家科學委員會專題研究計畫成果報告

應用於傳輸線暫態分析的尺域方法及
應用於印刷電路板由地反彈雜訊問題之線方法研究

Scale-Domain Method for Transmission Line Problems and Modified Method of Lines for PCB Ground Bounce Problems

計畫類別：■個別型計畫　　□整合型計畫
計畫編號：NSC　89－2213－E－002－139－
執行期間：89 年　8 月　1 日至　90 年　7 月　31　日

計畫主持人：鄭士康
研究人員　：江怡霆，張景程，許可欣，朱紹儀，林宗慶

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執行單位：　國立台灣大學電信工程學研究所

中　華　民　國　90 年　8　月　15　日

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中文摘要

本報告以線方法來分析多層印刷電路板之地彈雜訊與共振效應，此效應對整個系統之雜訊傳播有重要之影響。當問題的尺寸為最高頻的數倍波長時，以線方法分析較其他法如 FDTD 會較有效率。以所獲得之數值與實驗數據，我們即可預測去耦合電容與細縫之平面共振頻率。另一方面，尺域小波法將用來分析傳輸線之暫態反應。本法能處理多樣化之問題，包括色散、非線性特性與任意波形。此能力無疑地為暫態計算提供了另一有效之計算工具。本研究成果並已發表於國際期刊 [1] [2]。

關鍵詞：線方法，基版反彈雜訊，平面共振頻率，尺域，小波，色散，非線性。

Abstract

The Method of Lines (MoL) is used to analyze the ground bounce noise and resonance effects of a multi-layered PCB power plane, which affects the propagation of ground bounce noise in actual motherboard system. When the problem size is extended to several wavelengths of the highest operating frequency, this approach is more efficient than other methods like the FDTD. From the obtained numerical and experimental information, we can predict the effects of decoupling capacitors and slits on planar resonance frequency. On the other hand, we use Haar wavelet scale domain (HWSD) method to compute the transients of transmission lines. This method is versatile to handle dispersive line and nonlinear loads with arbitrary signal shapes. The ability of the HWSD method makes it another efficient tool for transients computation.

Keywords: method of lines, ground bounce noise, planar resonance frequency, scale domain, Haar, wavelet, dispersion, nonlinear effects.

I Introduction

When multi-layer PCB assemblies are used, it can be divided into layers of signal, power, and ground planes. The power plane is sometimes split on the same plane to separate, for example, the analog circuitry and digital logic and voltage reference areas. Due to the lead inductance of package, the output stage of chip will produce current surge during transitions. We call it ground bounce noise or delta-I noise. The noise can degrade the machine cycle time by causing delays and false switching and must be controlled for high performance system. Moreover, it becomes very significant on power and ground planes at the resonance frequencies of the plane. Such a noise may pollute the sensitive regions, increase the noise level in the system and produce some unpredictable radiation interference. Hence the prediction of resonance frequencies of power planes becomes important. Using the cavity model we can obtain a series of resonance frequencies for regions of rectangular shape. But the cavity model is not applicable when the coupling effects of slits and decoupling capacitors exist. Thus, we have to use numerical electromagnetics to predict it. The MoL is found more efficient than FDTD and MPIE(Mixed Potential Integral Equation) methods when the size of the problem is extended to several wavelengths of the highest operating frequency.

In addition, we propose the HWSD method for transients computation of transmission lines. This method is a general approach to handle lossless, lossy, dispersive lines and both linear and nonlinear loads with arbitrary shapes, which other methods can't do all well.

II Method of Lines and Equivalent Circuit Model

For our problem, constant permittivity ε is assumed for each dielectric layer. Conducting strips of vanishing thickness are located at the interfaces between each layer. From the Maxwell equations, we can derive a Helmholtz equation

$$\frac{\partial^2\varphi}{\partial x^2}+\frac{\partial^2\varphi}{\partial y^2}+\frac{\partial^2\varphi}{\partial z^2}+k^2\varphi=0$$

To limit the area of discretization, we enclose the structure by perfect magnetic wall on the edges. Each region in the power plane is separated by slits whose widths are around 20 to 40 mils. Both planes may sprinkle with small holes. To simplify the problem we assume that these small holes have slight effects on resonance frequency. Because the distance between two planes is much smaller than the wavelength of the highest operating frequency, we assume that the field components in the z direction are constant. Then we need only deal with the x and y variation in the Helmholtz equation, which is approximated by

$$\left(-h^{-2}\overline{\overline{P}}_{xy}+k_0^2\varepsilon_r\overline{\overline{I}}\right)\overline{\varphi}=0$$

where $\overline{\varphi}$ is a column vector representing the collection of sampled φ's on the e or h-lines and $\overline{\overline{P}}_{xy}$ is a constant matrix. The above equation is transformed by

$$\overline{\varphi}=\overline{\overline{T}}\overline{\varphi}$$

to get a system of uncoupled ordinary differential equations

$$\left(k_0^2\varepsilon_r\overline{\overline{I}}-h^{-2}\overline{\overline{\lambda}}\right)\overline{\varphi}=0$$

where $\overline{\overline{T}}$ is the transformation matrix composed of eigenvectors according to

$$\overline{\overline{T}}^{-1}\overline{\overline{P}}_{xy}\overline{\overline{T}}=\overline{\overline{\lambda}}^2$$

The resonance frequency then can be obtained

$$\omega_i=\frac{\lambda_i}{h\sqrt{\mu_0\varepsilon_0\varepsilon_r}}=\frac{c\lambda_i}{h\sqrt{\varepsilon_r}}$$

For the feeding, we apply an equivalent current source to simulate the coaxial cable feed as shown in Fig 1. With an impressed current source, the electric field is given by

$$\overline{E}_z=j\omega\mu\left(k_0^2\varepsilon_r\overline{\overline{I}}-h^{-2}\overline{\overline{P}}_{xy}\right)^{-1}\overline{J}_z$$

The numerical effort in computing the field is small because the matrix $\overline{\overline{P}}_{xy}$ is sparse. Since the decoupling capacitors are much smaller than the wavelength and the size of the plane, and an equivalent circuit model can be used to approximate these elements. The conducting and displacement currents flowing through a cell can be expressed as

$$I_{dsp}=j\omega\varepsilon_0\varepsilon_r\int\mathrm{E}\cdot d\bar{s}=j\omega\varepsilon_0\varepsilon_r A_{cell}E_z$$

$$I_{dev}=\frac{1}{Z_L}\int\overline{E}\cdot d\bar{l}\cong\frac{E_z h}{Z_L}$$

respectively. $Z_L=R+j\omega L+1/j\omega C$ is, therefore, the equivalent impedance of decoupling circuit, L is the inter-connect inductance, R is the loss factor, and C is the capacitance of the decoupling capacitor. The total current is the sum of the above two currents and then the equivalent current density can be obtained as

$$J_z=\frac{hE_z}{Z_L'}\cdot\frac{1}{A_{cell}}$$

where $Z_L'=Z_L\parallel Z_{c0}$, $Z_{c0}=1/j\omega C_0$, and $C_0=\varepsilon_0\varepsilon_r A_{cell}/h$ is the inter-plane capacitance. In addition, the coupling between two sides of a slit is modeled as a capacitor connecting both sides. The formulae are

$$\left(-hE_z^{i-1,j}+hE_z^{i+1,j}\right)\frac{1}{Z_L}=I$$

$$J_z^{i-1,j}=\frac{h}{A_{cell}Z_L}\left(E_z^{i+1,j}-E_z^{i-1,j}\right)$$

$$J_z^{i+1,j}=\frac{h}{A_{cell}Z_L}\left(E_z^{i-1,j}-E_z^{i+1,j}\right)$$

Here the width of the slit is small and approximated by one subdivision.

Ⅲ Haar Wavelet Scale Domain Method

The general telegraphist's equations in time domain are

$$-\frac{\partial}{\partial x}v(x,t)=R\cdot i(x,t)+L\frac{\partial}{\partial t}i(x,t)$$

$$-\frac{\partial}{\partial x}i(x,t)=G\cdot v(x,t)+C\frac{\partial}{\partial t}v(x,t)$$

Approximate the signals by

$$v(x,t)\approx\sum_{j=1}^{n}v_j(x)h_{j,k}(t)$$

$$i(x,t)\approx\sum_{j=1}^{n}i_j(x)h_{j,k}(t)$$

where $h_{j,k}(t)$ is the scaled Haar wavelet [3] with duration Δt defined

$$h_{j,0}(t)=\begin{cases}1, & (j-1)\Delta t\le t<j\Delta t\\ 0, & others\end{cases}$$

$$h_{j,k}(t)=\begin{cases}2^{\frac{p}{2}}, & j-1+\frac{2q}{2^{p+1}})\Delta t\le t<(j-1+\frac{2q+1}{2^{p+1}})\Delta t\\ -2^{\frac{p}{2}}, & (j-1+\frac{2q+1}{2^{p+1}})\Delta t\le t\ (j-1+\frac{2q+2}{2^{p+1}})\Delta t\\ 0, & others\end{cases}$$

By inserting the approximation into the transmission line equations and integrating with respect to t, the original system of continuous time-dependent equations transforms to a system of discrete time-independent matrix equations

$$-\frac{d}{dx}[v]=([R]+\frac{1}{\Delta t}[L])[i]$$

$$-\frac{d}{dx}[i]=([G]+\frac{1}{\Delta t}[C])[v]$$

Once these matrices are determined, equations can be manipulated in a way similar to those in the time-harmonic case. Problems with linear loads thus can be solved directly, and nonlinear loads can be dealt with by a common iterative scheme.

IV Numerical Examples

Consider an FR4 double side rectangular plane whose size is 10cm*10cm and dielectric constant is 4.4. We set two test ports on the plane. One is at (5cm, 5cm), and another is at (5cm, 2cm). Using the cavity model the resonance frequencies are shown in Table 1 up to 3 GHz. Figure 1 displays simulation result, which agrees with the experimental data. The peaks in Fig. 1 are consistent with the resonance frequencies in Table 1, but we can't find the 01, 10, or 11 modes in our simulation, because port 1 is on the central line of the plane, and is at a null-field point of these mode patterns. For the same structure, we place eight 1nF capacitors near port 1. The equivalent inductance of the via hole is about 1.25nH for FR4. After adding these capacitors, we find a minimum around 100MHz in Fig 2. Because the operating frequency is very low, we can explain this phenomenon by its equivalent circuit. Assume that all capacitors and inductances are identical and the board can be regarded as a capacitor of capacitance C_0. The input impedance of the equivalent circuit is

$$Z_r = \frac{1-\omega^2 L_{\mathit{eff}} C_{\mathit{eff}}}{j\omega\left[\left(C_{\mathit{eff}}+C_0\right)-\omega^2 L_{\mathit{eff}} C_{\mathit{eff}} C_0\right]}$$

where $L_{\mathit{eff}} = L/8$,and $C_{\mathit{eff}} = 8C$. The minimum around 100 MHz is just the resonance frequency of a series LC circuit composed of the inter-connect inductor and the decoupling capacitor. Around this band port 2 can't receive much power. On the other hand, the first maximum is due to the resonance of a parallel LC circuit composed of the inter-connect inductor and the inter-plane capacitor. The effect of the number of capacitors is shown in Fig 3. From Fig. 3 we can observe that an increase in the number of capacitors has little influence since it is related to a resonance due to lumped circuits. Also the first peak moves toward higher frequencies as the number of capacitors increases. Another consideration is the capacitance of individual decoupling capacitors. If we use different capacitance, from Fig. 4 we find that the first peak around 700MHz is almost the same in the first two cases because the capacitance of the decoupling capacitors are much greater than that of the inter-plane capacitors of the PCB. Serious influence on the first peak will occur when the capacitance of capacitors is near the inter-plane capacitance.

Fig 5 is the verification of the HWSD method with the 1D FDTD. The figure also shows the convergence of the HWSD method by two different parameters. Fig 6 is the verification with the phasor technique and the IFT (inverse Fourier transform), which a unit rectangular pulse propagating along a infinite Debye dispersive transmission line. The results are consistent. In the last example, we try to compute the transients of a Debye dispersive line with nonlinear loads. The convergence of Fig 7 shows the ability of the HWSD method to deal with the combined problem.

V Conclusion

The resonance effect of a multi-layered PCB power plane has been predicted by the MoL. Compared with other method, the MoL is more efficient when the problem size is extended to several wavelengths of the highest operating frequency. For the HWSD method, we have validated its usefulness in dispersive transmission lines with nonlinear loads. It is indeed a effective and simple technique for a variety of transient problems of transmission lines.

VI Reference

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[2] I.T. Chiang, S.K. Jeng, " Staircase Approximation for Transients of Multisection Dispersive Transmission Lines with Nonlinear Loads, " IEEE AP-S 2001.

[3] I.T. Chiang, S.K. Jeng, " Haar Wavelet Scale Domain Method for Solving the Transient Response of Dispersive Transmission Lines with Nonlinear Loads," submitted to IEICE Transaction on Communications.

m n	$f_{mn}(GHz)$
00	0
01 or 10	0.715
11	1.011
20 or 02	1.430
12 or 21	1.599
22	2.022
31 or 13	2.261
32 or 23	2.578

Table 1: Propagation mode frequencies in the parallel plate resonator.

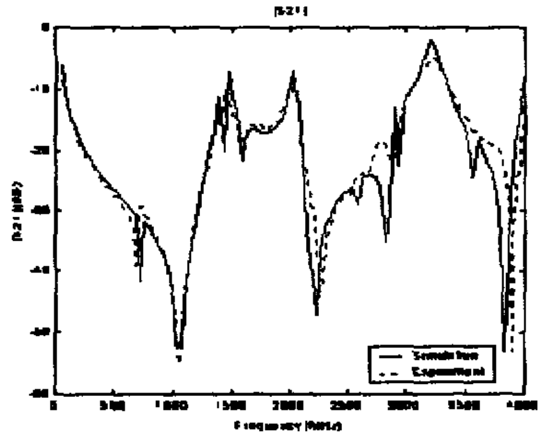

Fig. 1: The magnitude of S21 for a 10cm*10cm rectangular plane.

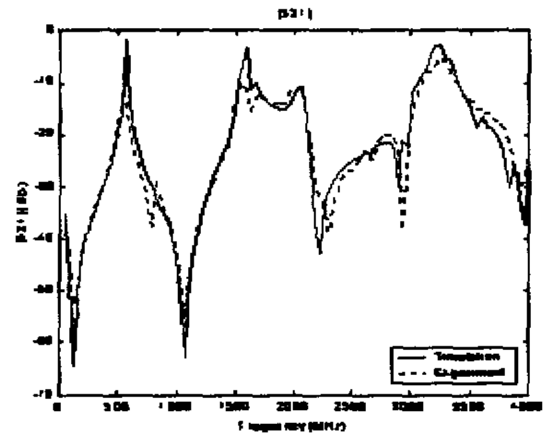

Fig. 2: The magnitude of S21 for a 10cm*10cm rectangular plane with eight decoupling capacitors.

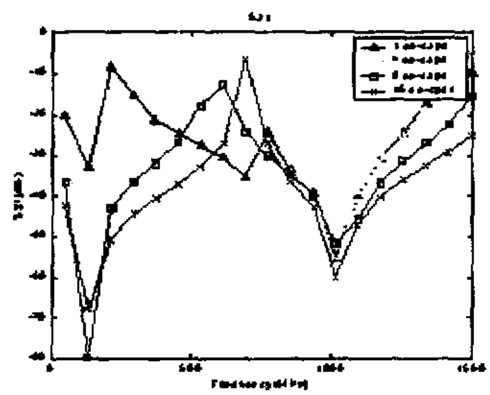

Fig. 3: The effect of the number of capacitors.

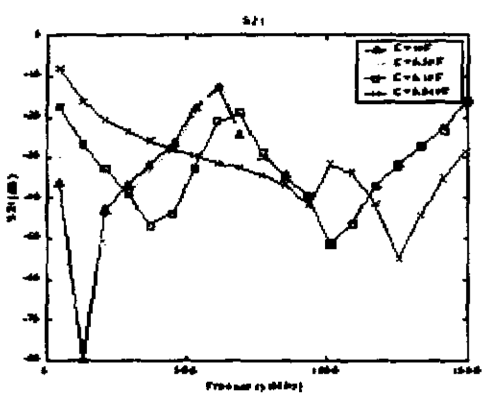

Fig. 4: The effect of different capacitances.

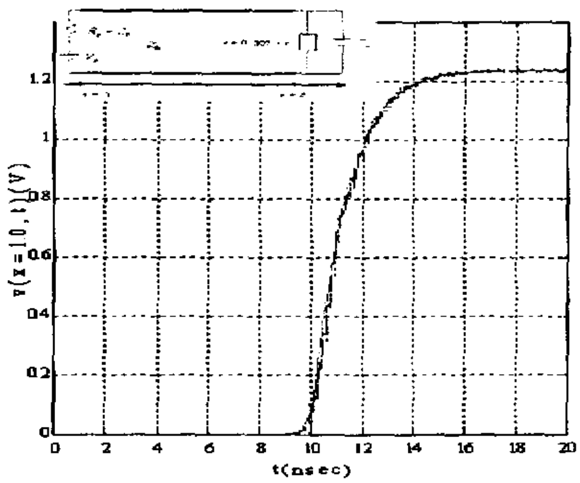

Fig. 5: Comparison of FDTD and HWSD

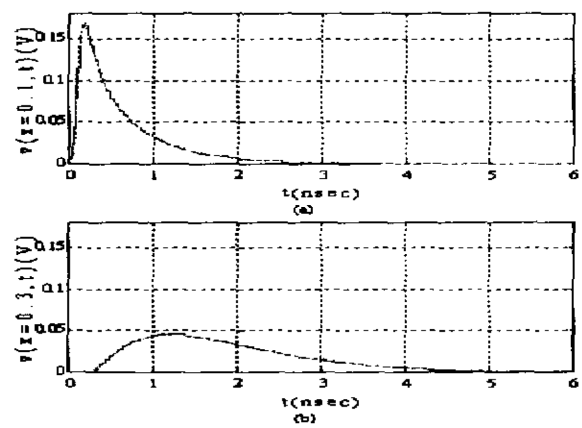

Fig. 6: Comparison of IFT and HWSD

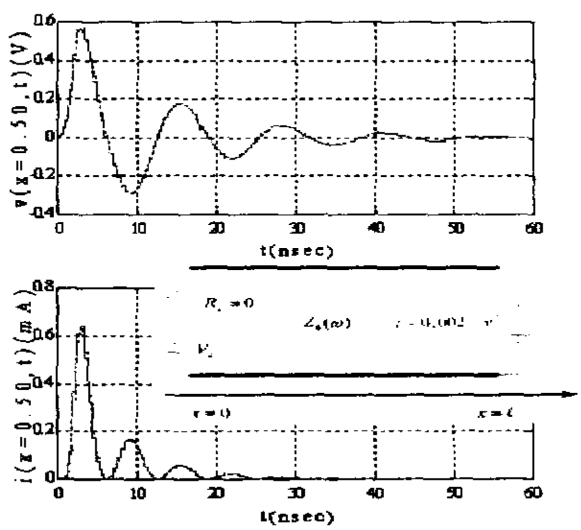

Fig. 7: Convergence of the HWSD

2001 IEEE-APS/URSI 開會報告 博士生江怡憲

本屆（2001 年）的 IEEE-APS/URSI 在 Boston 的 Sheraton 飯店舉行，時間為七月八日至七月十二日。其中第一天為報到日兼歡迎餐會，可以看到大部分來參加此次會議的人，包括教授、業界人士與在學的研究生。正式的會議報告於七月九日早上至七月十二日的下午舉行，依論文性質分別在不同的會議室報告，而每個會議主題都有一至兩個主持人負責主持會議。此外，在會議進行期間亦有一些包括儀器設備生產商和書商等相關廠商招攬生意。

基本上，本次會議內容算是相當豐富、涵蓋主題也相當廣泛，相信參加的每個人都可以找得到自己有興趣的主題聆聽。不過因為本人在國內英文的聽力較缺乏練習，故聽別人報告時頗為吃力；再加上會議過程房間大都保持黑暗以利清楚呈現投影片內容，故過程難免容易疲累。不過好在，除了學術上，也可以有充分的時間抽空到 Boston 走走，除了市容外，兩個有名的大學：MIT（麻省理工學院）以及哈佛大學也理所當然地參訪了一下。雖然台大的面積沒有人家大、草地也不像國外大學那樣又大又綠可以躺在上面睡覺、看書以及野餐，但麻雀雖小卻也是五臟俱全。比較了一下，台大仍然是相當好的學校，台大電機也是相當好的科系，並不會上不了國際舞台。此外，我曾數了數各個學校的投稿篇數，本次會議投稿最多的學校為伊利諾大學（UIUC）、俄亥俄（Ohio）大學與密西根（Michigan）大學，而電波組投稿的篇數雖比不上這三所，但共投八篇在所有的學校中也算是不錯的了。

這是我第一次出國參加學術研討會，給我不少我出國前所沒有的新觀感。基本上，我認為只要有機會，博士班研究生都應該儘量出國參加會議以吸取第一手資訊，這不但是一種自我訓練，還可以兼培養國際觀，才不至於被一些眼前短視近利的現象所迷惑。我也希望，政府可以特別補助這些第一次出國的博士班研究生。很多事情唯有自己走一趟，用自己的眼睛親眼看、用自己的耳朵親耳聽才會有感觸。此外，加強英文聽力與演說能力亦頗為重要，這是台灣高等教育裡面所缺乏的，我甚至覺得可以將之列為大學甚至研究所每學期的必修課。

此次會議資料，最基本的包括會議進行程序（每個人的報告題目與時間）以及所有投稿的論文集。此外，就是一些 Boston 的相關觀光資料。較特別的是，我報告後聽眾中有兩人對我的題目頗有興趣，在交換了一下心得與討論後，他們分別給了我一張他們的名片，希望能繼續保持聯絡。看了一下他們的名片，分別是德州農工大學（Texas A&M University）的教授（此然還兼任副系主任）與馬特拉(MATRA)公司的一位研究員。台大電信所電波組這次共投了八篇會議論文，但只有我有收到別人主動給我名片，或許也是對我的一種肯定吧！

後投之論文

Staircase Approximation for Transients of Multisection Dispersive Transmission Lines with Nonlinear Loads

I-Ting Chiang and Shyh-Kang Jeng
Graduate Institute of Communication Engineering and
Department of Electrical Engineering
National Taiwan University
Taipei , Taiwan , ROC.
Email:skjeng@ew.ee.ntu.edu.tw

Abstract

In this paper, we propose a staircase approximation to solve for the transients of dispersive transmission lines with nonlinear loads. Numerical examples, verified with the FDTD method and the traditional frequency-domain approach, are included. Such an approach is with easy formulation and straightforward numerical solutions, especially when dealing with multisection lines.

I Introduction

As the clock rate of digital circuits goes higher and higher, dispersion of transmission lines takes a more crucial role than before. With dispersion, the pulse shape is distorted, which leads to variations of duration, rise time and fall time. If the pulse shape is distorted, nearby signals are difficult to distinguish if care is not taken. In short, only when signals can be predicted and controlled beforehand do the digital circuits function correctly, especially when the dispersion effect is a major concern.

For microwave circuits, frequency response can be obtained either directly in frequency domain or indirectly in time domain via FFT. Although the harmonic balance method [1] is widely adopted in solving nonlinear loads in frequency domain, transmission lines in that method are usually dealt with in time domain, which inevitably faces the problem with dispersion and nonlinear loads if high frequency effects are taken into account.

To treat the transient response of dispersive transmission lines, the frequency-domain method is adopted traditionally. It suffers, however, not only the difficulty with the presence of nonlinear loads but also the slow numerical integration when taking the inverse Fourier transform. The FDTD, another candidate, is capable of settling nonlinear loads, but has difficulties when frequency-dependent factors must be considered. In addition, the FDTD becomes involved in dealing with multisection transmission lines, which further limits its applications.

The scale domain method [2] can solve for the transients of dispersive lines with nonlinear loads through a straightforward formulation. The next section will give a sketch of the staircase approximation, a simplification of the scale domain method. Section III provides some numerical examples, and the last section draws some conclusions.

II Formulation

The dispersive telegraphist's equations in time domain are

$$-\frac{\partial}{\partial x}v = Ri + \frac{\partial}{\partial t}(L * i) \qquad (1)$$

$$-\frac{\partial}{\partial x}i = Gv + \frac{\partial}{\partial t}(C * v) \qquad (2)$$

where the star "*" represents the convolution operation to account for dispersion. Approximate the signals by

$$v(x,t) \approx \sum_{j=1}^{n} v_j(x) h_j(t) \qquad (3)$$

$$i(x,t) \approx \sum_{j=1}^{n} i_j(x) h_j(t) \qquad (4)$$

where $h_j(t)$ is the unit rectangular pulse with duration Δt. This is called the staircase approximation since the voltage and current signals at a given x is now approximated by a staircase- like function. By inserting (3) and (4) into (1) and (2) and integrating with respect to t, the original system of continuous time-dependent equations transforms to a system of discrete time-independent matrix equations

$$-\frac{d}{dx}[v] = ([R] + \frac{1}{\Delta t}[L])[i] \qquad (5)$$

$$-\frac{d}{dx}[i] = ([G] + \frac{1}{\Delta t}[C])[v] \qquad (6)$$

where $[v]$ and $[i]$ are column matrices and $[R]$, $[L]$, $[G]$, $[C]$ are square matrices with

$$[R] = R[I] \qquad (7)$$

$$[G] = G[I] \qquad (8)$$

$$[L] = L[M]^{-1} \qquad (9)$$

$$[C] = C[M]^{-1}(\varepsilon_i[I] + \frac{1}{\Delta t}[Dis]) \qquad (10)$$

Here [I] stands for identity matrix and

$$[M] = \begin{bmatrix} 0.5 & 0 & \cdots & \cdots & 0 \\ 1 & 0.5 & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \vdots & & \ddots & 0.5 & 0 \\ 1 & \cdots & \cdots & 1 & 0.5 \end{bmatrix}$$

$$[Dis] = d_{ij} = \begin{cases} \frac{a}{b^2}[e^{b(i-j+1)\Delta t} - 2e^{b(i-j)\Delta t} + e^{b(i-i-1)\Delta t}] & , \quad i > j \\ \frac{a}{b^2}[e^{b(i-j+1)\Delta t} - e^{b)i-j)\Delta t} - b\Delta t] & , \quad i = j \\ 0 & , \quad i < j \end{cases}$$

Assume Debye dispersion [3]

$$C(\omega) = C\varepsilon_r(\omega)$$

$$\varepsilon_r(\omega) = \varepsilon_\infty + \chi(\omega)$$

$$\chi(\omega)=\frac{\varepsilon_s-\varepsilon_\infty}{1+j\omega t_0}$$

So

$$\chi(t)=\frac{\varepsilon_s-\varepsilon_\infty}{t_0}e^{-t/t_0}U(t)$$

by inverse Fourier transform, and

$$a=\frac{\varepsilon_s-\varepsilon_\infty}{t_0}$$

$$b=-\frac{1}{t_0}$$

correspondingly. Here ε_s means the relative permittivity at dc, ε_∞ is the relative permittivity at $\omega=\infty$, and t_0 stands for the Debye relaxation constant.

Once these matrices are determined, equations can be manipulated in a way similar to those in the time-harmonic case. Problems with linear loads thus can be solved directly, and nonlinear loads can be dealt with by a common iterative scheme [2].

Ⅲ Numerical Results

To prove the usefulness of the staircase approximation, consider a two-section lossless ($R=G=0$) transmission lines shown in Fig. 1, and shunt with a capacitor and a nonlinear load at the interconnection and the right terminal. At the same time, generator with an internal resistor is placed at the left end.

As the first example, assume both sections of the transmission lines are with the same parameters, $\ell=0.5(m)$, $L=0.5(\mu H/m)$, and $C=0.2(nF/m)$. In addition, a matched generator excites unit rectangular pulse with duration $w=1(n\sec)$. The nonlinear loads are described by $i=0.01\times v^2$ for $v>0$ and $i=0$ for $v\le 0$, and shunt with capacitors $C_s=50(pF)$. The resultant voltage signal at $x=1.0(m)$ calculated by the staircase approximation (solid line) and the FDTD (dashed line) are shown in Fig. 2. Both match well. The zero voltage before $10(n\sec)$ is due to the delay of propagation. The rise and fall of the voltage signal between $10(n\sec)$ and $20(n\sec)$ are related to the charge and discharge of capacitors caused by the finite duration pulse. The rise after $20(n\sec)$ is excited by the reflection from the interconnection.

Next, let's remove the nonlinear loads and introduce the Debye dispersion to both transmission lines with parameters $\varepsilon_s=9$, $\varepsilon_\infty=4$ and $w_0=1/t_0=5\pi\times10^8$. Apply the same pulse excitation and replace the internal resistor by $R_g=150(\Omega)$. The voltage response calculated at $x=0.5(m)$ and $x=1.0(m)$ by the staircase approximation (solid line) and the frequency-domain transform method (dotted line) are illustrated in Fig. 3. The zero voltage are again due to the propagation delay. The smoother shapes and smaller magnitude, compared with the previous example, reveal the effect of dispersion. The agreement of both curves validates the capability of our method in dealing with dispersive transmission lines.

Last, but not least, assume the same parameters used in the second example and apply the nonlinear loads utilized in the first example. The results are exhibited in Fig. 4, in which the solid line is with the parameters $\Delta t = 0.1$ ($n$sec) and 512 bases while the dashed line adopts $\Delta t = 0.2$($n$sec) and 256 bases. The results show that the convergence of our method is pretty good.

IV Conclusion

We have proposed the staircase approximation and shown its usefulness in dealing with transients of multisection dispersive transmission lines with nonlinear loads. Numerical results verified with the FDTD and the conventional frequency-domain method have been exhibited. This method can be easily formulated and applied to problems with frequency-dependent loads, which is important for more realistic applications.

V Reference

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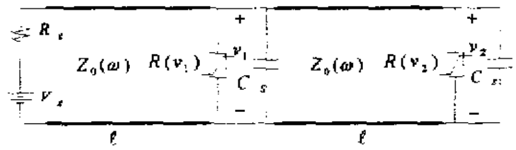

Figure 1

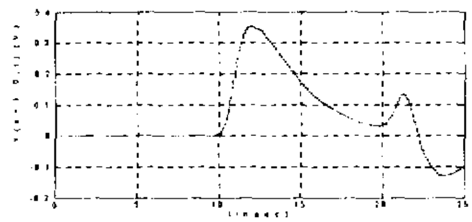

Figure 2

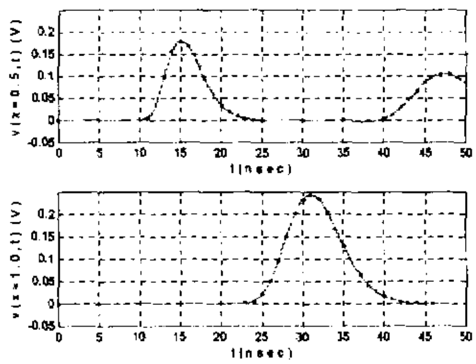

Figure 3

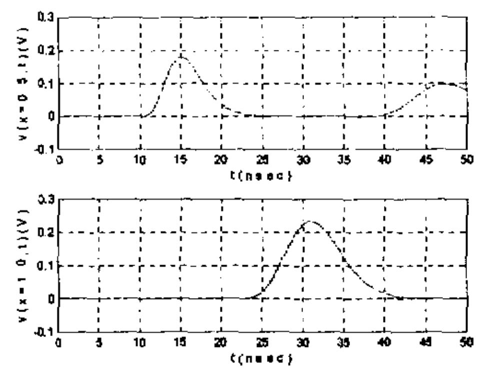

Figure 4

參加第十四屆蘇黎世國際電磁相容研討會與相關產品展示會（14th International Zurich Symposium and Technical Exhibition on Electromagnetic Compatibility, EMC2001 Zurich）報告

鄭士康
台大電機系/電信研究所
中華民國九十年三月十七日

一、引言

隨著電子產品快速發展，產品內外之電磁波干擾日益嚴重，如何減少彼此電磁干擾之電磁相容問題日趨重要。敝人數年前指導學生開始此方面研究，略有心得。此次獲國科會補助，進行「應用於傳輸線暫態分析的尺域方法及應用於印刷電路板由地反彈雜訊問題之線方法研究」研究（補助編號：NSC89-2213-E-002-139），已具初步成果，乃投稿 2001 年蘇黎世國際電磁相容研討會與相關產品展示會（International Zurich Symposium and Technical Exhibition on Electromagnetic Compatibility, EMC 2001 Zurich），經審查通知錄取，於大會發表論文。此一研討會自 1975 年開始，由瑞士聯邦理工學院（Swiss Federal Institute of Technology Zurich, ETHZ）會同 IEEE 電磁相容學會（IEEE Electromagnetic Compatibility Society）等多個國際組織，每兩年於瑞士蘇黎世舉行一次，此次爲第十四屆。每屆會議除論文發表外，另有特定主題之講習班（Tutorial Lectures）、專題討論會（Workshops）、產業論壇（Industrial Forums）、電磁相容國際標準公開討論會（Open Meeting）及電磁相容有關產品展示；除 IEEE 電磁相容學會之年會外，此會議應爲世界最具規模之電磁相容會議。

本屆會議於二月二十至二十二日舉行。由於在當地停留六天以上之機票較爲便宜，決定二月十七
日星期日出發，二十四日星期六晚間返抵台北。經歷十數小時過境等候及飛行，於當地時間十八日上午抵達蘇黎世，住進 ETHZ 附近一家小型旅館，名爲 Leoneck，房間設備裝潢就其價位而言，均屬上乘。在旅館稍事休息，下午至蘇黎世市區最繁華之車站大街（Bahn Strass）參觀，因商店週日皆不營業，僅能觀賞其櫥窗，但遊人依然不少。隨後至附近蘇黎世湖及湖濱公園遊覽，由於天氣陰沉，湖光山色未如預期。傍晚來到其舊市區，教堂鐘塔、大小街道、老式建築與雕像，均整潔典雅，別有特色。十九日上午前往 ETHZ 電子研究所，尋訪現在當地進修之本系闕志達教授不遇，當日下午至會場報到。二十至二十二日參加會議。二十一日晚終於連絡上闕教授，並蒙其安排，於二十三日到附近知名滑雪場參觀，見識滑雪運動後，於晚間搭機回國。

二、會議經過

報到

二月十九日下午前往 ETH Main Building 會場報到，領取資料袋及論文集。資料袋中裝有識別名牌、最後議程手冊、會場附近生活有關資訊（如午、晚餐參考地點與價格範圍，一般餐廳及旅館小費慣例等等）、蘇黎世地圖、會議期間大眾交通工具搭乘票券、作者午餐（Authors' Lunch）餐券、會議間休息（Coffee

Break）之飲料券、註冊費收據、參展廠商及產品目錄等，極其周到。以上詳細列出資料袋內項目，以供日後國人舉辦國際會議參考。

會場各種指示牌標示清楚，位置適當，並有電視機多部播放最新訊息。註冊報到程序流程亦繪成流程圖，以單槍投影機投射至牆壁上，一目瞭然。除註冊報到處外，另外設有詢問處（含外幣匯兌服務）、視聽設備協助等，工作人員態度親切，令人印象深刻。

第一日

開幕式於於 ETH Main Building 之大會堂（Auditorium Maximum）舉行，會場座位幾乎全部坐滿，參加人數約在三百餘人左右，相當盛大。開幕式由大會創辦人兼主席，ETH 教授 Osterwalder 主持，多位瑞士學界與業界名人發表演講。IEEE 電磁相容學會主席 Mr. J. Butlert 除致賀詞外，並簡介 IEEE 電磁相容學會近況。會中之 Keynote Speaker 由美國加州大學柏克萊校區之 Professor W. J. Welch 擔任，介紹外星生物搜尋計劃之大要。顯然，來自外星生物之訊號必定相當微弱，因此電磁干擾之隔絕，即電磁相容問題，在此種研究中扮演非常重要的角色。

開幕式後即於三個會場平行展開論文宣讀。此三個會場均佔地寬敞，可容納二百餘人，設備極佳。會議進行時並有電腦連線，由專人操控，以單槍投影機即時顯示各個會場議程（Session）及專題討論會（Workshop）進行狀況，方便聽眾掌握時間，不致遺漏想聽講之論文宣讀。論文發表進行時，每位講員約有二十分鐘，由主席設定定時鐘，剩餘八分鐘以上時定時鐘顯現綠燈，二分鐘至八分鐘時顯示黃燈，二分鐘內出現紅燈，逾時則紅燈閃爍，對講者控制時間極有幫助。講者可自由選用一般投影機或以筆記型電腦連接單槍投影機。會場銀幕及白板相當大，可同時容納講者之投影片內容與議程進行資訊。

上午議程爲 EMC Protection、 Adverse Effects of High Power EM 、Medical and Biological Issues，平行進行。敝人參加 Session A: EMC Protection，共有五篇論文發表，包括以 MPIE 積分方程分析金屬容器之屏蔽效果、頻率選擇薄片之理論分析、多金屬傳輸線之不均勻處特性標定、圓形線圈低頻磁場之多層屏蔽效果、導電性織品之應用等。其中導電性織品可製成衣物、帳蓬等，隔離電磁雜訊，頗爲有趣。

下午之平行議程有 EMC in Networks、Sensors and Probes、EMC in Power Systems，敝人參加 EMC in Networks，有六篇論文。論文內容印象較深者包含汽車、飛機內電線之電磁場耦合、無線電手機天線與電線之電磁場耦合、電車電力輸送線對通訊電纜之影響，因爲涉及之電線爲數眾多，所以多採取機率、統計分

析，或統計與物理模式混合之分析。

就在參加下午議程之前，敝人巧遇昔日在伊利諾大學香檳校區（University of Illinois at Urbana-Champaign）進修時認識之友人 Omar Ramahi，晤談甚歡。Omar 曾於多家知名電腦公司任職，研究電磁相容問題多年，績效卓著，現在馬里蘭大學（University of Maryland）任教。敝人與之交換電磁相容研究及教學心得，彼此均有所獲。

第二日

第二日上午之議程，分爲 Transients、Transmission Lines、Modeling Large Chips and Packages 三部份，敝人論文爲 Modeling Large Chips and Packages 議程發表八篇論文中之第三篇。每一議程開始前，出席之論文講員均須至議場辦公廳（Symposium Officers Lounge）集合，與議程主席及其他講員會面，並聽取論文發表過程注意事項。在此一聚會中，得知敝人之議程主席 Dr. A. Ruehli 爲本系同仁及本次論文共同作者江簡富教授過去於 IBM T. J. Watson 研究中心服務時之舊識，言談甚歡。同一議程之論文集中討論大型積體電路於多層電路板之電磁干擾問題。敝人發現，我們探討的地彈雜訊（Ground Bounce Noise）爲甚多研究單位關心之課題，敝人學生賴明佑完成之線方法(Method of Lines)分析，簡單易懂，應用範圍廣泛，爲其他相關論文所不及。

同議程其他論文則多報告以較傳統之 Method of Moments、PEEC、FD-TD 方法模擬分析電路板上金屬傳輸線之干擾問題的結果。

下午議程爲 Lightning、Measurement Techniques、Computer Codes and Validation。敝人對 Computer Codes and Validation 較感興趣。該議程發表之論文亦有八篇，內容涵括電磁干擾計算方法及軟體技術之應用。其中敝人最注意者爲英國 De Montfort University 學者 H. G. Sasse 與 A. P. Duffy 所發表之 Topics in the Design of a Parallel TLM Solver。該論文指出目前大部分之個人電腦閒置之時間與記憶體甚多，而電磁相容問題計算常需甚長時間與龐大記憶，若能連接多部個人電腦，使資源得以互通有無，將可增加計算效率。雖然該論文僅探討 TLM 方法之平行化，但同樣想法應可推廣至其他領域之計算方法。過去之平行計算方式，多爲同一電腦內多個處理器之平行處理，而現在個人電腦及網路普及，應有不同之平行計算方式。此種研究或許於計算機科學中已在進行，或已有成果，當予留意。

第三日上午議程爲 High Frequency Methods and Analysis、Test Chambers and Cells、PCBs in the GHz Range。敝人參加 PCB in the GHz 部分，共有七篇論文，

重點仍爲印刷電路板之電磁干擾問題。其中有關地彈雜訊之研究論文有兩篇，其分析方法仍爲傳統之特徵函數展開、FD-TD 等。Omar 在此 Session 亦發表一篇有關電路板上延遲線（Delay Lines）之全波分析之論文，提及較長之蜿蜒線（Meander Lines）將產生週期性諧波，影響其延遲訊號之特性，相當有趣；由此出發，Omar 提出方形螺線等結構，可避免此種週期性諧波，相當有用。

下午有四個議程：EMC Innovation、EMC in Communication Systems、Reverberation Chambers、Chip-Level EMC。敝人參加 Chip-Level EMC 部分，亦有七篇論文，重點爲晶片及電路板組合之電磁干擾問題，同樣也有一篇論文討論地彈雜訊。一位義大利學者提出以系統辨識（System Identification）方法建立積體電路之黑箱模型，以供電磁相容分析，論點新穎。一位英國學者利用 FD-TD 及 Method of Moment 分析積體電路散熱器（Heat Sink）的電磁輻射，亦有特殊之處。接下來有位日本學者分析 CMOS 積體電路的電磁干擾，其中所舉之電路實例爲傳輸線接電晶體電路，量測其暫態訊號變化。此種問題恰爲敝人學生江怡霆在敝人同一國科會補助計劃中之另一研究課題，顯然具有很高之實用價值。江同學之初步結果亦已由 2001 年 IEEE 天線與傳播學會年會(Antenna and Propagation Symposium）接受，將於本年七月赴美國波士頓發表。

下午五點半，所有議程結束後，籌備單位舉行歡送派對（Farewell Party），提供簡易點心及紅酒果汁，與會人士互相攀談，爲本屆會議畫下句點。

三、整體印象及心得

此屆會議參與之學者專家多半來自歐洲、美國，日本、韓國各有十餘人，而敝人爲唯一來自台灣之參與者。印象中台灣之科技發展與美國聯繫較多，對歐洲之來往較少。然而，歐洲在許多方面之科技，例如本次研討會之主題：電磁干擾防治，並不比美國遜色多少，因此有必要加強對歐洲的認識。

參加此次會議印象最深刻之處爲會場設施之完善、籌備之細密、與展示廠商之眾多及敬業。廠商之敬業精神可由以下事實看出：一般會議最後一日之展示廠商多半早已停止展出；但據筆者觀察，本次會議直至最後一日下午，所有四十餘家軟硬體產品廠商代表仍據守攤位，與參觀人士洽談。設施完善及籌備細密或許與主辦單位已舉行同一會議十餘次有關，但廠商敬業則可能與參展廠商多爲德國籍有關。德國人民認真、注重條理、有始有終之敬業精神，以往僅有耳聞，此次實地見識，確實可供台灣借鏡。

四、結語

電磁相容問題對電子產品之影響愈來愈大，不僅產品需符合各國之電磁相容標準，以免妨礙其他產品之運作，而且產品內部元件之間，也需減少其因產品輕薄短小而劇增之電磁干擾，方能達到其設計之功能。鑒於台灣經濟與電子業息息相關，國人實應體認電磁相容研究之重要。歐美各國對此方面之研究起步較早，技術設備較爲先進。但在學理方面，因涉及因素甚多，先前僅能對問題進行簡單分析；近年則因計算工具有快速進展，進步較多。目前國內電磁相容技術僅有二、三政府單位及若干大廠商研發，而相關學理之探討研究，在各學術單位中亦屬少數。然而，國內若干大學之計算電磁理論專家不乏其人，若能將其專才運用於此一領域，其成效當可直追世界水平，敝人此次與同仁江簡富教授及學生共同發表之論文或可視爲其一例證。

另外，由此次會議可以看出，成功之國際會議可提高主辦單位之學術聲望，令外國高級知識份子留下良好印象，並可爲主辦國帶來不錯之商業利益。然而，國際會議之籌辦經緯萬端，有賴經驗傳承，方可達到如本次會議之水準。台大慶齡工業中心日前成立專責機構，協助校內單位籌辦國際會議，應是一個良好的起步。

發表之論文

Multi-Layered PCB Power Plane Resonance Analysis Using Method of Lines

Ming-Iu Lai, Jean-Fu Kiang and Shyh-Kang Jeng
National Taiwan University
Graduate Institute of Communication Engineering and
Department of Electrical Engineering
(E-mail: skjeng@ew.ee.ntu.edu.tw)

Abstract: The Method of Lines (MoL) is used to analyze the resonance effect of a multi-layered PCB power plane, which affects the propagation of ground bounce noise in actual motherboard system. When the problem size is extended to several wavelengths of the highest operating frequency, this approach is more efficient than other methods like the Finite Difference-Time Domain (FD-TD) method and the Mixed Potential Integral Equation (MPIE) method. From the obtained numeric and experimental information, we can predict the effects of decoupling capacitors and slits on planar resonance frequency.

Index Terms-Method of Lines, ground bounce noise, planar resonance frequency.

1. Introduction

When multi-layer PCB assemblies are used, it can be divided into layers of signal, power, and ground planes [1]. The power plane is sometimes split on the same plane to separate, for example, the analog circuitry and digital logic and voltage reference areas [1]. Due to the lead inductance of package, the output stage of chip will produce current surge during transitions. We call it ground bounce noise or delta-I noise. The noise can degrade the machine cycle time by causing delays and false switching and must be controlled for high performance system [2]. Moreover, it becomes very significant on power and ground planes at the resonance frequencies of the plane. Such a noise may pollute the sensitive regions, increase the noise level in the system and produce some unpredictable radiation interference [3]. Hence the prediction of resonance frequencies of power planes becomes important. Using the formula of cavity model we can easily obtain a series of resonance frequencies for regions of rectangular shape. But the cavity model is not applicable when the coupling effects of slits and decoupling capacitors exist.

According to numerical and experimental results [4]-[9], the lowest resonant frequency zone of the power plane is usually up to several hundreds MHz, thus we have to use numerical electromagnetics to predict it. The most popular methods for solving such a problem are the Finite Difference – Time Domain (FD-TD) method [7] and the Mixed Potential Integral Equation (MPIE) method [8][9]. However, these approaches will take long time in computing when the size of the problem is extended to several wavelengths of the highest operating frequency.

In this paper we'll introduce the Method of Lines (MoL) to deal with this problem. The MoL is found more efficient than FD-TD and MPIE methods. In Section 2 we formulate the problem by the Method of Lines and equivalent circuit model. In Section 3 we exhibit some simulation and measurement results and give some discussions on the results. In Section 4 we give a brief conclusion.

2. Method of Lines and Equivalent Circuit Model

2.1 Method of Lines

For our problem, constant permittivity ε is assumed for each dielectric layer. Conducting strips of vanishing thickness are located at the interfaces between each layer. From the Maxwell equations, we can derive a Helmholtz equation

$$\frac{\partial^2\varphi}{\partial x^2}+\frac{\partial^2\varphi}{\partial y^2}+\frac{\partial^2\varphi}{\partial z^2}+k^2\varphi=0 \tag{1}$$

with ω the angular frequency, μ the permeability of free space, φ the z-component of the electric or the magnetic field, $k^2=\omega^2\mu\varepsilon$. The coordinate system is shown in Fig. 1.

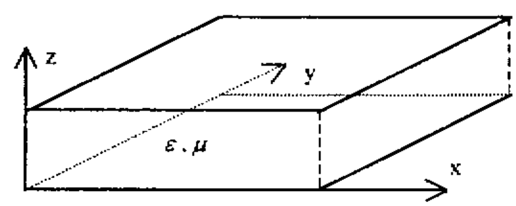

Fig. 1: The coordinate system of the problem.

To limit the area of discretization, we enclose the structure by perfect magnetic wall on the edges [12]. Each region in the power plane is separated by slits whose widths are around 20 to 40 mils. Both planes may sprinkle with small holes. To simplify the problem we assume that these small holes have slight effects on resonance frequency. Because the distance between two planes is much smaller than the wavelength of the highest operating frequency, we assume that the field components in the z direction are constant. Then we need only deal with the x and y variation in the Helmholtz equation. After discretization, the Helmholtz equation is approximated by

$$\left(-h^{-2}\overline{\overline{P}}_{xy}+k_0^2\varepsilon_r\overline{\overline{I}}\right)\overline{\varphi}=0 \tag{2}$$

where $\overline{\varphi}$ is a column vector representing the collection of sampled φ's on the e or h-lines shown in Fig. 2 and $\overline{\overline{P}}_{xy}$ is a constant matrix.

The above equation is transformed by

$$\bar{\varphi} = \overline{\overline{T}}\varphi \tag{3}$$

to get a system of uncoupled ordinary differential equations

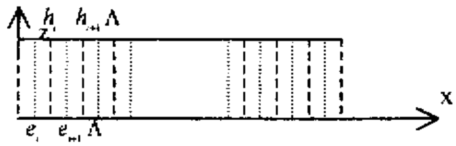

Fig. 2: The e and h-lines for $y = y_0$.

$$\left(k_0^2\varepsilon_r\overline{\overline{I}} - h^{-2}\overline{\overline{\lambda}}\right)\overline{\overline{\varphi}} = 0 \tag{4}$$

where $\overline{\overline{T}}$ is the transformation matrix composed of eigenvectors according to

$$\overline{\overline{T}}^{-1}\overline{\overline{P}}_{xx}\overline{\overline{T}} = \overline{\overline{\lambda}} \tag{5}$$

The resonance frequency then can be obtained

$$\omega_i = \frac{\lambda_i}{h\sqrt{\mu_0\varepsilon_0\varepsilon_r}} = \frac{c\lambda_i}{h\sqrt{\varepsilon_r}} \tag{6}$$

For the feeding, we apply an equivalent current source to simulate the coaxial cable feed as shown in Fig 3. With an impressed current source, the electric field is given by

$$\overline{E}_z = j\omega\mu\left(k_0^2\varepsilon_r\overline{\overline{I}} - h^{-2}\overline{\overline{P}}_{xx}\right)^{-1}\overline{J}_z \tag{7}$$

The numerical effort in computing the field is small because the matrix $\overline{\overline{P}}_{xx}$ is sparse.

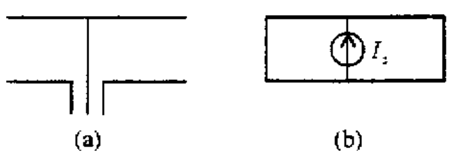

Fig. 3: (a) The structure of coaxial feed. (b) The equivalent method of coaxial feed.

2.2 Equivalent Circuit Model

Since the decoupling capacitors are much smaller than the wavelength and the size of the plane, and an equivalent circuit model can be used to approximate these elements. The conducting and displacement currents flowing through a cell can be expressed as

$$I_{disp} = j\omega\varepsilon_0\varepsilon_r\int \mathrm{E}\cdot d\bar{s} = j\omega\varepsilon_0\varepsilon_r A_{cell}E_z \tag{8}$$

$$I_{dev} = \frac{1}{Z_L}\int \bar{E}\cdot d\bar{l} \cong \frac{E_z h}{Z_L} \tag{9}$$

respectively. Hence $Z_L = R + j\omega L + 1/j\omega C$ is the equivalent impedance of decoupling circuit, L is the inter-connect inductance [12], R is the loss factor, and C is the capacitance of the decoupling capacitor. The total current is the sum of the above two currents and then the equivalent current density can be obtained as

$$J_z = \frac{hE_z}{Z_L'}\cdot\frac{1}{A_{cell}} \tag{10}$$

where $Z_L' = Z_L \,\|\, Z_{c0}$, $Z_{c0} = \frac{1}{j\omega C_0}$, and $C_0 = \frac{\varepsilon_0\varepsilon_r A_{cell}}{h}$ is the inter-plane capacitance.

In addition, the coupling between two sides of a slit is modeled as a capacitor connecting both sides. The formulae are

$$\left(-hE_z^{i-1,j} - hE_z^{i,j}\right)\frac{1}{Z_L} = I \tag{11}$$

$$J_z^{i-1,j} = \frac{h}{A_{cell}Z_L}\left(E_z^{i+1,j} - E_z^{i-1,j}\right) \tag{12}$$

$$J_z^{i+1,j} = \frac{h}{A_{cell}Z_L}\left(E_z^{i-1,j} - E_z^{i+1,j}\right) \tag{13}$$

Here the width of the slit is small and approximated by one subdivision.

3. Results and Discussions

3.1 Rectangular Plane and Decoupling Capacitors

Consider an FR4 double side rectangular plane whose size is 10cm*10cm and dielectric constant is 4.4. We set two test ports on the plane. One is at (5cm, 5cm), and another is at (5cm, 2cm). Through the cavity model the resonance frequencies are given by

$$f_{mn} = \frac{c}{2\sqrt{\varepsilon_r}}\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2} \tag{14}$$

with a and b the length of the sides of the power plane, c the lightspeed in free space and m and n the mode numbers. Table 1 shows the resonant frequencies and the corresponding indexes up to 3 GHz. Figure 4 displays the simulation result, which agrees with the experimental data. The peaks in Fig. 4 are consistent with the resonance frequencies in Table 1, but we can't find the 01, 10, or 11 modes in our simulation, because port 1 is on the central line of the plane, and is at a null-field point of these mode patterns.

For the same structure, we place eight 1nF capacitors near port 1. The equivalent inductance of the via hole is about 1.25nH for FR4.

Table 1: Propagation mode frequencies in the parallel plate resonator.

m n	$f_{mn}(GHz)$
00	0
01 or 10	0.715
11	1.011
20 or 02	1.430
12 or 21	1.599
22	2.022
31 or 13	2.261
32 or 23	2.578

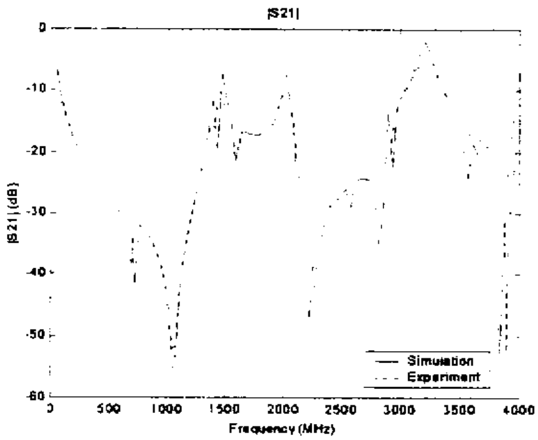

Fig. 4: The magnitude of S21 for a 10cm*10cm rectangular plane.

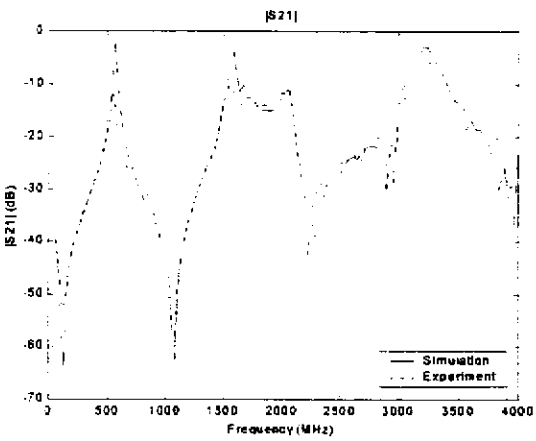

Fig. 5: The magnitude of S21 for a 10cm*10cm rectangular plane with eight decoupling capacitors.

After adding these capacitors, we find a minimum around 100MHz in Fig 5. Because the operating frequency is very low, we can explain this phenomenon by its equivalent circuit. Assume that all capacitors and inductances are identical and the board can be regarded as a capacitor of capacitance C_0. The input impedance of the equivalent circuit is

$$Z_T = \frac{1-\omega^2 L_{eff} C_{eff}}{j\omega\left[\left(C_{eff}+C_0\right)-\omega^2 L_{eff} C_{eff} C_0\right]} \tag{15}$$

where $L_{eff} = L/8$,and $C_{eff} = 8C$. The minimum around 100 MHz is just the resonance frequency of a series LC circuit composed of the inter-connect inductor and the decoupling capacitor. Around this band port 2 can't receive much power. On the other hand, the first maximum is due to the resonance of a parallel LC circuit composed of the inter-connect inductor and the inter-plane capacitor. The effect of the number of capacitors is shown in Fig 6. From Fig. 6 we can observe that an increase in the number of capacitors has little influence since it is related to a resonance due to lumped circuits. Also the first peak moves toward higher frequencies as the number of capacitors increases. Another consideration is the capacitance of individual decoupling capacitors. If we use different capacitance, from Fig. 7 we find that the first peak around 700MHz is almost the same in the first two cases because the capacitance of the decoupling capacitors are much greater than that of the inter-plane capacitors of the PCB. Serious influence on the first peak will occur when the capacitance of capacitors is near the inter-plane capacitance.

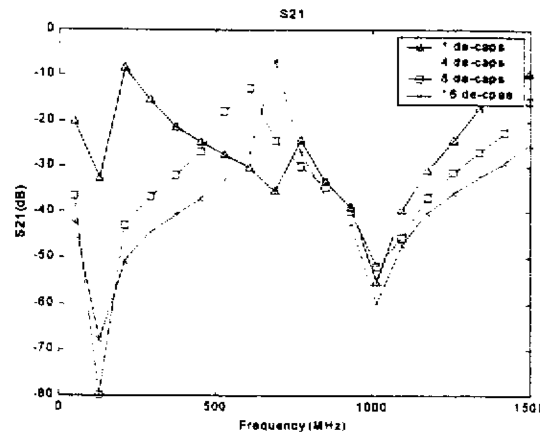

Fig. 6: The effect of the number of capacitors.

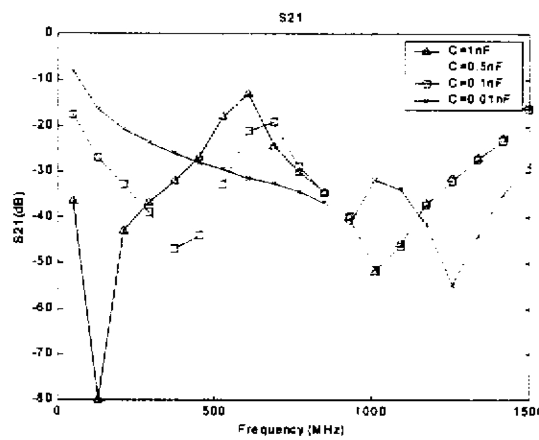

Fig. 7: The effect of different capacitances.

3.2 Effect of slits

Consider a rectangular plane whose size is 10cm*5cm. In the first case we etch two 2mm slits on the upper plane, and in the second case two 2mm slits on the upper and the bottom plane. The simulation and experimental results are exhibited in Fig 8. They are consistent only around the first peak. The simulation results are not as good at high frequency. The reason may be that our algorithm is based on the assumption of constant field in the z direction, while when the e-lines are near the slit, the field component in the z direction should have a large variation. Besides, we only use a capacitor to simulate the coupling effect, which may be good only for low frequency, and more complex coupling effect across the slot could happen in the high frequency region.

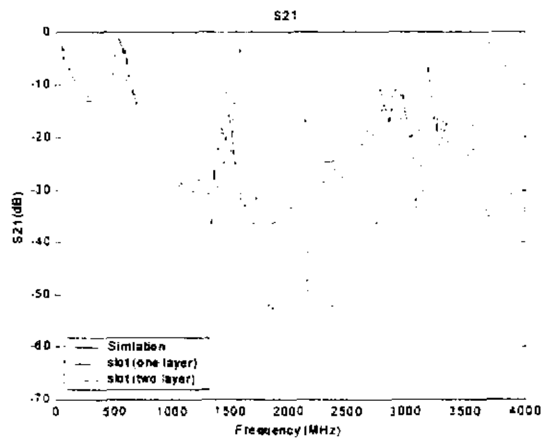

Fig. 8: The effect of slits.

4. Conclusions

Based on the Method of Lines a new numerical tool has been proposed. One advantage of this approach is that it's more efficient than convential numerical methods. For example, when we deal with a rectangular plane whose size is 10cm*10cm, our approach takes about 1 minute for a single frequency on a PC with an AMD K-6-300 CPU. In the same platform using IE3D sofeware to solve the same problem takes about 40 minutes for one frequency. Though this method has some troubles in high frequency, it has good proformance up to 1 GHz. Finally, our numerical results also suggest two rules for the layout of power and ground planes:

1).Use the slits on power plane as less as possible.

2).Put the sensitive chips or intensive noise sources near the central part of a rectanglar region to reduce the number of excited resonance modes.

5. References

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