



Towards a Description Synthesis of the Entanglement of the Substrate with the Interconnection Network, for Fast Modeling of 3D RF circuits

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ABSTRACT

3D chip stacking is considered known to overcome conventional 2D-IC issues, using Trough Silicon Via (TSVs) to ensure vertical signal transmission between data. If the electrical behaviour of 3D interconnections (redistribution metal lines and through silicon vias) used in 3D IC stack technologies are to be explored in this paper, the substrate itself is of interest, via Green Kernels by solving Poisson's equation analytically. Using this technique, the substrate coupling and loss in IC's can be analysed. We implement our algorithms in MATLAB. This method has been already used; but, it permits to extract impedances for a stacked uniform layers substrate. We have extended for any numbers of embedded contacts, of any shape. On a second hand, we grasp the background noise between any two points, in the bulk, or at the surface, from a transfer impedance extraction technique. With an

analog algorithm, a strength of this work, we calculate unsteady solutions of the heat equation, using a spreading resistance concept. This method has been adapted to stacked layers. With this general tool of impedance field, we investigate on the problems encountered by interconnects, especially the vias, the substrate, and their entanglement. A calculation of thermal mechanical stresses and their effects on substrate crack (max and min stresses), devices (i.e: transistors) and hotspots, are made to track the performance. But, to well understand the interconnection incidence on 3D system performances, it is important to consider the whole electrical context; it seems relevant to consider the possible couplings between vias, not only by the electromagnetic field, but also by any possible energy transfer between interconnects; more generally, one of actual problem is to determine where the energy is really confined in such 3D circuits, before find solutions to limit pollutions coming from electro-magneto -thermal phenomena or background noises.

Index Terms— Substrate, Interconnects, Electronic noise, via, stress, 3D RF ICs, Green kernel, TLM, Energy Transfer, transfer impedance.

INTRODUCTION

For many years, the increase in the performance of microelectronics devices has gone with the miniaturization of these latter. In order to follow Gordon Moore's 1965 forecasts, the microelectronics industry has established a doubling of the number of transistors per unit of surface every 18 months until it sees its limits a few years ago with 28 and 14 nm technologies. As the manufacture of smaller transistors is difficult to make for large-scale marketing, only three manufacturers in the world are still able to continue miniaturization below 10 nm (Intel, Samsung and TSMC). This race for miniaturization is all the more pronounced in the field of technologies smartphones, tablets or connected objects that require an increase in performance without changing dimensions. This results in a gradual decrease in the size of the chips, as well as the intensification of their transistors and poses many challenges in terms of interconnection and thermal management. In contrast to the standard circuit architecture, which consists of having single-functional device on a printed circuit board (PCB), designers developed embedded structures in mobile devices such as the so-called systems-on-chip (SoC). But there is still a long way to better understand the disadvantages inherent in these new technologies, and to try to propose possible solutions to reduce or even eliminate them. Our aim is to build the heart of a design Kit dedicated to RF embedded circuits. First of all, we introduce general tools, that we call spreading or transfer impedance, that permit to calculate general impedances, also associated to heat transfer, and a general electronic noise transfer between any two points . Then, we point out on crucial problems as stress induced by mechanical or temperature hotspots .The paper ends by some modelling stress-induced failures, a predilection of drawbacks in such circuits.

MATHEMATICAL TOOL

The final goal of this work is to define different levels of model abstraction and to develop a hierarchical modelling concept to be able to adequately describe the impact of parasitic effects on the performance of blocks at each relevant abstraction level during the architectural model based design of 3D systems. Effects which are exacerbated are e.g. noise coupling and thermal and stress effects by enabling high device density in all dimensions in 3D integration. Due to the different physical effects, their local or global existence and influence in the 3D system,

different levels of abstraction have to be used to decrease the simulation time or to make the simulation even possible. These abstraction levels will vary from detailed field calculations using e.g. finite element methods to behavioural models used in circuit simulation, e.g. Spice, VHDL-AMS, and Verilog-AMS.

This task will also focus on the development of a flexible high-level interconnect model using MATLAB, taking into account technological data and TSV models, and aiming to enable the estimation of the communication cost in terms of latency, throughput, power, silicon cost (based on spatial organization), and its consequent impact on function performance and system performance. First of all, in this paper, Green functions are applied to homogeneous layers of the circuit substrate model extraction, as opposed to global numerical methods like differences or finite elements methods (FEM). For sure, these method are actually used in this domain, and others [1- 7] . But, our deep will is to try a possible synthesis. Some basic recalls and concept are first introduced. The use of discrete cosine transforms (DCT), applying twice a FFT - Fast Fourier Transform - in this model, accelerate the computation speed, compared to classical numerical techniques notably FEM). All the calculations are made in the reciprocal domain. That means we take also FFTs both for the injected current, I source, and the injected current, - I sink. Then, a global model, which can be applied on any substrate with vias, themselves decomposed in stacked basic vias along the layers they goes through, or embedded contacts, these of any shape, any thickness, in any number , is shown; it can treat the case of contacts, or probes, lying in different layers. All the calculations made, we come back to real space. As a check, we compare with some FEM commercial simulators[8-9].

Substrate Extraction Technique

1. Green Kernels

The high density integration and high system frequency make substrate noise coupling becoming one of the most significant considerations in the design, due to its great impact on the performance of ICs. The aim of the 3D substrate analysis is, first of all, to efficiently extract the Z impedance parameters between any contacts (electrodes, via, probes) which are located on or into the silicon substrate. An efficient impedance extraction tool could help the designer accelerate the design and optimize the final layout. Like in planar technologies, 3D interconnects can be built as an 'RLCG' - divider bridge: $(G//C)/(R \text{ series } L)$ - equivalent electrical model, with a Π or a T network. Often, a simple compact model is constructed by modeling the substrate as a simple node. However, this assumption is only viable when substrate is highly conductive, in low and medium frequency domains, and is not suitable for the multi-layer substrates. That is the reason why, in this paper, we propose a global substrate extraction method relying first on a Transmission Line Matrix (TLM) one, more physical method analog to the Green one.

2. Substrate analysis

Generally, the Z parameters could be defined as:

$$Z_{mn} = \frac{V_m}{I_n} \Big|_{I_{k \neq n} = 0} \quad (1)$$

In practical operation, by injecting an unitary current as an excitation source, I, at an n^{th} point, and a unitary current excitation source, -I ,(equivalent to a sink current I) at a m^{th} point, and

calculate the resulting voltage at a probe point, we can get the impedance Z_{mn} between the m and n contacts, directly.

If we can get the resulting potential distribution in the substrate caused by the unitary current, we could get the Z parameters by (1) easily.

Under quasi-static conditions, the potential over the substrate satisfies Laplace's equation:

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2} \right) \varphi(x, y, z) = 0 \quad (2)$$

- *Brief recalls* [1].

If we assume that $\varphi = X(x, x') \cdot Y(y, y') \cdot Z(z, z')$, we can rewrite equation (2), thus resulting in:

$$\begin{aligned} Y \cdot Z \cdot \frac{d^2 X}{dx^2} + X \cdot Z \cdot \frac{d^2 Y}{dy^2} + X \cdot Y \cdot \frac{d^2 Z}{dz^2} \\ = - \frac{\delta(x - x') \cdot \delta(y - y') \cdot \delta(z - z')}{\epsilon_N} \end{aligned} \quad (3)$$

The solution of such equation can also be considered as the superposition of a transmitted and reflected RF waves at any layer interface.

The Green function then changes to:

$$\begin{aligned} \varphi(x, y; x', y')_{z=z'=0} (G_0)_{z=z'=0} + \\ \left(\sum_{m=0}^{\infty} \sum_{n=0}^{\infty} f_{mn} C_{mn} \cos\left(\frac{m\pi x}{a}\right) \cos\left(\frac{m\pi x'}{a}\right) \cos\left(\frac{n\pi y}{b}\right) \cos\left(\frac{n\pi y'}{b}\right) \right) \end{aligned} \quad (4)$$

where $C_{mn} = 0$ for $(m=n)>0$, $C_{mn}=2$ for $m=0$ or $n=0$, and $C_{mn}= 4$ for all others m and n ($m>0$ and $n>0$).

The function f_{mn} is thus given by:

$$\begin{aligned} f_{mn} \\ = \frac{1}{ab\gamma\epsilon_N} \frac{\beta_N \tanh(\gamma_{mn}d) + \Gamma_N}{\beta_N + \Gamma_N \tanh(\gamma_{mn}d)} \end{aligned} \quad (5)$$

β_N and Γ_N can be computed recursively from:

$$\begin{aligned} \begin{bmatrix} \beta_k \\ \Gamma_k \end{bmatrix} \\ = \frac{1}{2} \begin{bmatrix} 1 + \frac{\epsilon_{k-1}}{\epsilon_k} & (1 - \frac{\epsilon_{k-1}}{\epsilon_k})e^{2\theta_k} \\ (1 - \frac{\epsilon_{k-1}}{\epsilon_k})e^{-2\theta_k} & 1 + \frac{\epsilon_{k-1}}{\epsilon_k} \end{bmatrix} \begin{bmatrix} \beta_{k-1} \\ \Gamma_{k-1} \end{bmatrix} \end{aligned} \quad (6)$$

Where $\theta_k = \gamma_{mn} (d-d_k)$, $\beta_o = 1$, $\Gamma_o = 1$.

As an example, we show on Fig. 1, a stacked layers system modelling the substrate, coming from a profile P+/P of a 0.35 μm BiCMOS technology, and a calculation of a impedance modulus between two embedded contacts.

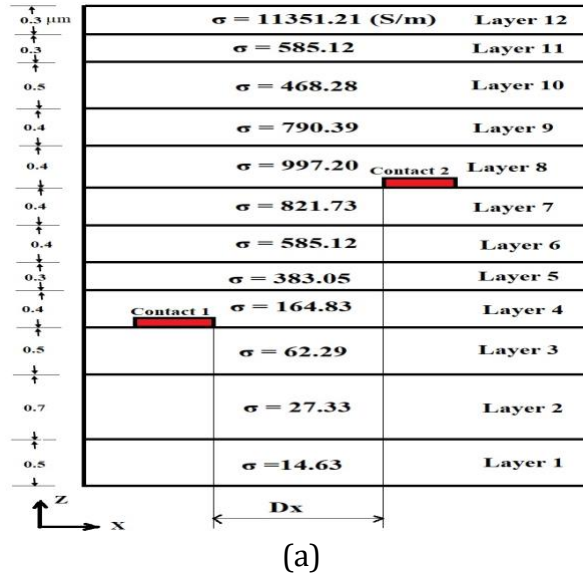
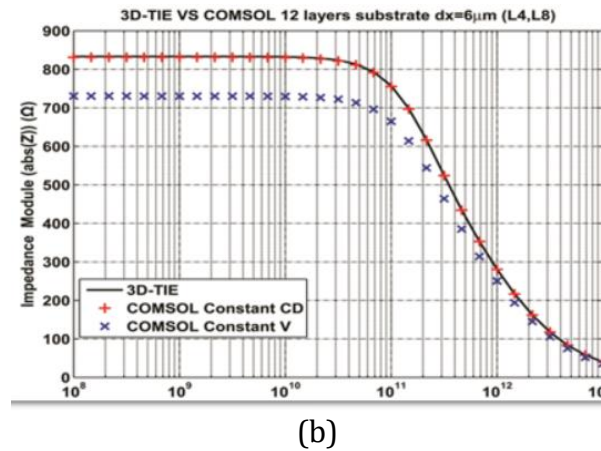


Fig. 1. (a) Stacked layers, with their conductivity and thicknesses; (b) impedance modulus between two contacts (see Fig. 1. (a)): our Method and COMSOL

3. Unstationary Solution of Heat Equation

We consider now a three-dimensional transient heat conduction in a nonhomogeneous finite medium, that which is more general that what we see in the literature [10-13].

Here also, contacts or probes numbers is not limited; the method, for such a thermal impedance calculation, is very fast, defining a new impedance, analogue to the classical impedance one presented above. The mathematical formulation of the problem is taken as:

$$\frac{d^2T}{dx^2} + \frac{d^2T}{dy^2} + \frac{d^2T}{dz^2} - \frac{1}{k} \cdot \frac{d\tau}{dt} = 0 \quad (7)$$

If the last term of the left hand of the equation equals 0, the Laplace equation is considered corresponding to the steady state for equations (7).

As it can be seen, we get analogies with equations (2) – (3).

Then, we assume that $T = X(x, x') \cdot Y(y, y') \cdot Z(z, z') \cdot \tau(t, t')$, we can rewrite equation (7), thus resulting in:

$$\begin{aligned} & Y \cdot Z \cdot \tau \cdot \frac{d^2X}{dx^2} + X \cdot Z \cdot \tau \cdot \frac{d^2Y}{dy^2} + X \cdot Y \cdot \tau \cdot \frac{d^2Z}{dz^2} + X \cdot Y \cdot Z \cdot \frac{d\tau}{dt} \\ &= - \frac{\delta(x - x') \cdot \delta(y - y') \cdot \delta(z - z') \delta(t - t')}{k} \end{aligned} \quad (8)$$

for $x \neq x' \ y \neq y' \ z \neq z', t \neq t'$. τ is assimilated to a Green kernel.

The only non-trivial solution, in time, is then:

$$\tau/\tau'' = -\kappa^2 \quad (9)$$

which has the evanescent solution:

$$\tau = Ae^{-\beta t} \quad (10)$$

Finally the general solution of eq.(8) takes the following expression:

$$\begin{aligned} \varphi &= X(x, y, x', y', z, z') = (1 - e^{-\beta t}). \\ \sum_{m=1}^{\infty} \sum_{n=1}^{\infty} [2 - \frac{(A_N e^{-\gamma_{mn}(d+Z_l)} + B_N e^{-\gamma_{mn}(d+Z_l)})(e^{-\gamma_{mn}Z_h} + e^{-\gamma_{mn}Z_h})}{ab\epsilon_N \gamma_{mn}(B_N e^{\gamma d} + A_N e^{-\gamma d})} \cdot \cos \frac{m\pi x}{a} \cos \frac{n\pi y}{b}] \end{aligned} \quad (11)$$

where $Z_l = \min (z, z')$, $Z_h = \max (z, z')$

We give, on Figure 2, a solution example when a 1V pulse is applied on two embedded contacts (see Fig1), and 2.3 μs after the plateau onset of this temporal pulse.

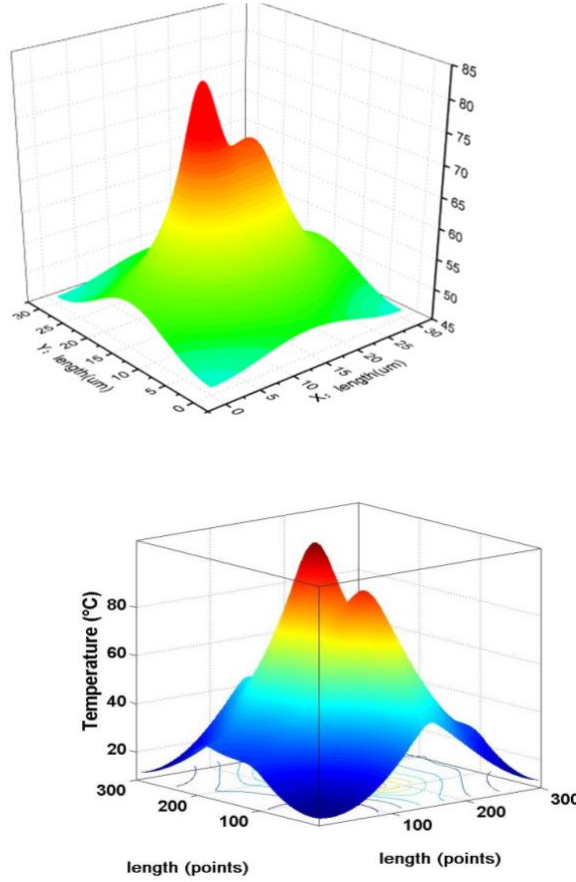


Fig. 2. Temperature map, after a temperature pulse applied in L4 and L8 layers (up) COMSOL[5], (down) MATLAB [6], (see Fig.1)

By referring to a TLM (Transmission Line Matrix) method, we can define a “thermal spreading impedance” as following:

$$R = A_i / B_i \quad (12)$$

where A_i and B_i are the Fourier coefficients (FC) of the general i solution for the temperature in the direction perpendicular to the layers:

$$FC_{si} = A_i \cosh(\gamma_{si}) + A_i \sinh(\gamma_{si}) \quad (13)$$

An impedance associated to Fourier coefficients ratio development of temperature solution can be defined:

$$R_{FCDi,z}(\gamma_n) = \frac{k_{i+1} R_{FCDi+1,z}(\gamma_n) + k_i \tanh(\gamma_{i,z} t_i)}{k_\lambda R_{FCDi+1,z}(\gamma_n) \tanh(\gamma_{i,z} t_i)} \quad (14)$$

where k_i is the thermal conductivity of the i^{th} layer,

$$\gamma = \left[\left(\frac{m\pi}{a} \right)^2 + \left(\frac{n\pi}{b} \right)^2 \right]^{1/2}.$$

We calculate a spreading impedance associated to Fourier coefficients ratio development of temperature solution, which we think from great interest; particularly, it permits to work in the reciprocal domain (see Figure 3). For instance, at each frequency, we can extract temperature profiles (not presented here).

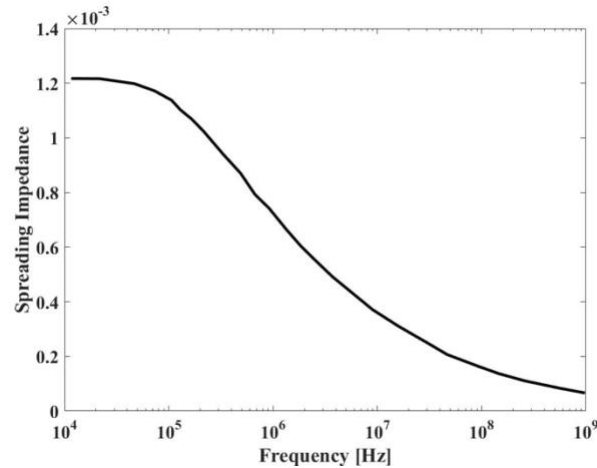


Fig. 3. Fourier coefficients spreading impedance: temperature pulse input (contacts in L4 and L8; (see Fig 1)

Noise Calculation Using the Impedance Field Method

It is an understatement to say that background noise is not well taken into account in 3D circuits, even in 2D: very few results for several decades, the pioneering papers of Van der Ziel. What more is, the dimensions decreasing drastically, this is definitely a key point. For example, sources of noise, once independent, are no longer uncorrelated, because of their proximity. This field is very poorly regarded nowadays, while it is increasingly unavoidable.

We calculate now electronic noise at any point of the 3D structure (Fig 1), induced by some GR noise source (see Fig 4). [14,15]

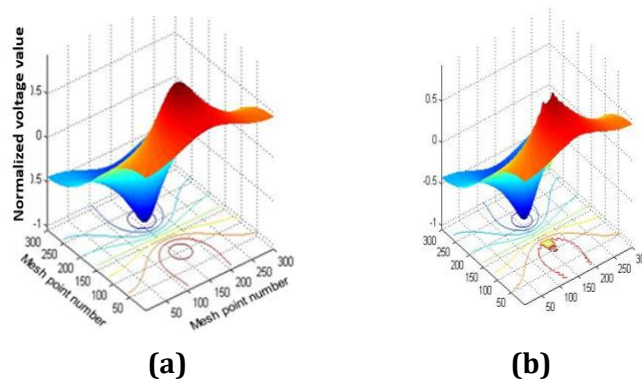


Fig. 4. Potential difference between 2 embedded contacts

**(a) without noise; (b) GR , on the upper contact
(L8)**

1. Principle of the method [14-15]

The impedance field method was introduced by W. Shockley et al. [14]. This method differs from the Langevin's method in the way it links the macroscopic quantity to elementary microscopic fluctuations. The spectral density of voltage fluctuations can be brought down to two basic calculations: that of the spectral density of voltage fluctuation and that of elementary fluctuations.

2. Impedance Fields

The method, below, seems well suited to our stacked layer structures, the noise spreading through cells, correlated or not, as to the point of view of background noise.

Let us consider a polarized specimen between an electrode N and the ground. If we inject a little current ΔI , flowing out of N, at a coordinate point $\vec{r}$ in the volume, the potential of N varies by $\Delta V(N, \vec{r})$ – see Fig 5 - .

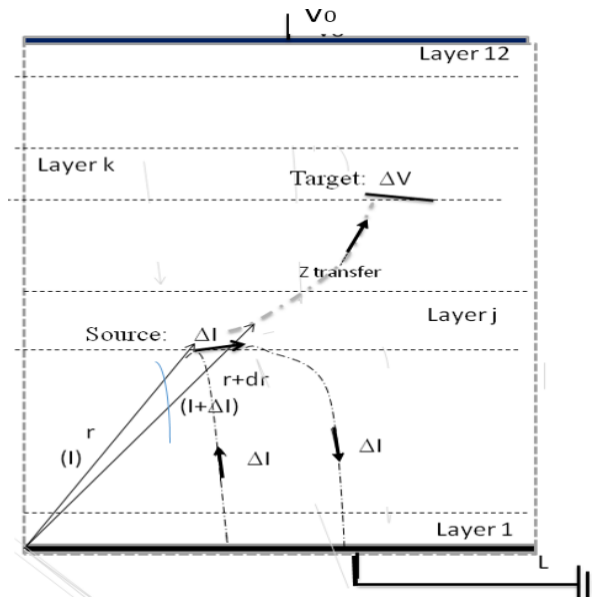


Fig.5: Principle of the transfer impedance.

Each noise source at the coordinate point $\vec{r}$ in the substrate (and also in the devices) produces a noise at another point of the volume and the surface (and electrodes).

The noise induced depends on its sources and its way of propagation via the impedance field.

In order to study the noise between two contacts on the surface or in the substrate, we inject a little current (noise) that is complex, ΔI , at frequency f , at a coordinate point $\vec{r}$ ($\pm \vec{dr}$) in the substrate, superimposed on a continuous current (cf. operational point) between the contacts. This little current produces a potential $V(f)$ in another contact. Thus, we can write:

$$\Delta V(r+dr) = Z(r+dr) \delta I \quad (15)$$

Equation (15) means that a current fluctuation in the dr^3 volume produces a voltage difference between two other contacts. Integrating on the domain considered, the total voltage produced by these currents is:

$$\Delta V_f = \int \vec{V} Z(r, f) \cdot \delta \vec{j}(r) \cdot dr^3 \quad (16)$$

In order to explicitly calculate the noise density $S_v(f)$, we use, for example, Cartesian coordinates, from x, y and z directions; we get in an α direction:

$$\Delta V_f = \sum_{\alpha} \left[\int \frac{\partial Z(r, f)}{\partial \alpha} \vec{V} Z \cdot \delta j_{\alpha}(r) \cdot dr^3 \right] \quad (17)$$

Thus, the voltage noise spectral density becomes:

$$S_v(f) = 2 \int_{-\infty}^{\infty} \Gamma_{\delta V}(N, \vec{r}, t) \cdot e^{2\pi i f t} dt \quad (18)$$

Where $\Gamma_{\delta V}(N, \vec{r}, \tau)$ is the correlation function;

$$\Gamma_{\delta V} = \langle \delta V_f^*(t) \delta V_f(t + \tau) \rangle \quad (19)$$

The density becomes

$$S_v(f) = \sum_{\alpha} \sum_{\beta} \iint \frac{\partial Z(r, f)}{\partial \alpha} \cdot \frac{\partial Z^*(r', f)}{\partial \beta} \cdot 2 \int_{-\infty}^{\infty} \overline{\delta j_{\alpha}(r, t) \cdot \delta j_{\beta}^*(r', t + \tau)} e^{2\pi i f \tau} dt \cdot dr^3 dr'^3 \quad (20)$$

We present on the Figure 6 some Lorentzian response to a GR source applied on the upper contact (with a local conductivity $\sigma = \sigma_0/(1+j\omega\tau)$, where τ is taken, here, as $10^{-9}s$) of Fig1.

$$S_v(f) = \sum_{\alpha} \sum_{\beta} \iint \frac{\partial Z'(r, f)}{\partial \alpha} \cdot \frac{\partial Z'(r', f)}{\partial \beta} \cdot S_{j_{\alpha\beta}}(r, r'', f) dr^3 dr'^3 \quad (21)$$

If K is a noise source:

$$S_{j_{\alpha\beta}}(r, r'', f) = K_{\alpha\beta}(r, f) \delta(r' - r)$$

equation **Error! Reference source not found.** simplifies into:

$$S_v(f) = \int \vec{V} Z(r, f) \cdot K(r, f) \cdot \vec{V} Z^*(r', f) \quad (22)$$

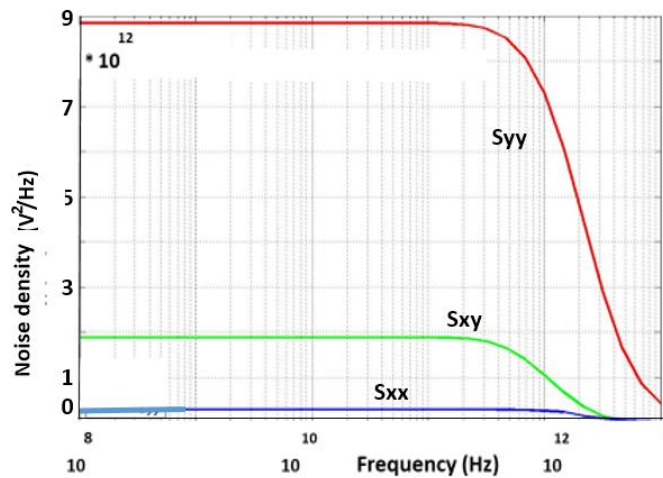
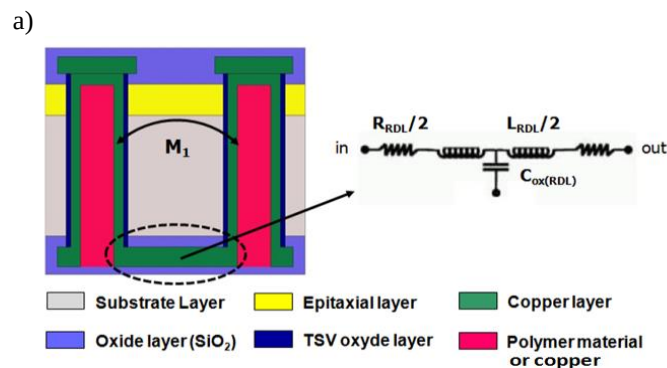


Fig. 6 Transfer impedance associated to GR noise

If now, we look at the reduction of parasitic effects, with a higher interconnection density, faster inter-chip communication speed and less energy consumption are the main advantages of TSV interconnection [16-20]; crosstalk becomes a key problem in reducing the TSV performance due to the fact that the silicon substrate with conductive properties, has parasitic capacitance effects on the insulating oxide (SiO_2 , isolation material) surrounded by the TSV and the isolation layer formed by the TSV metallization, especially in TSV arrays.

3. TSV Coupling Analytical Approach

Consider a basic configuration (Fig 7): two via linked by an RDL (ReDistribution Layer) [17-20].



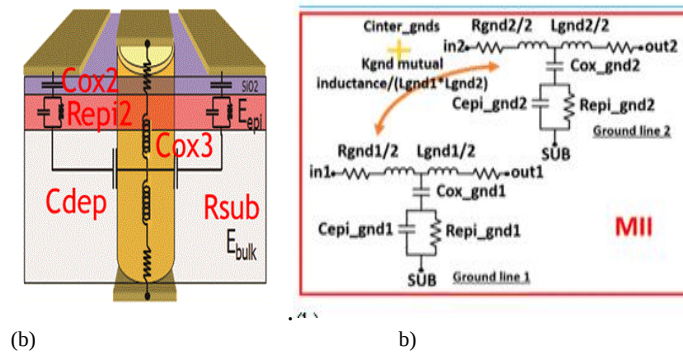


Fig. 7. (a) 2xTSVs chain side view: representation of the inductive coupling between the TSVs (b) RLCG compact model of the TSVs/RDL line

The total magnetic field, induced by the two via1 1 and 2, is $\vec{B}_1 + \vec{B}_2$. Electric charges are distributed non-uniformly on the surface of these conductors, although considering static or quasi-static currents. The presence of these surface charges is necessary for setting in motion the load carriers and for the transfer of electromagnetic energy *via* the Poynting vector.

To analyze some via effects, namely here capacitive, of the coating oxide of the via, it seems sufficient to solve analytically Poisson's equation, here in cylindrical coordinates with a full depletion approximation; this full depletion approximation assumes that there are no mobile charge carriers (holes or electrons) in the depletion region.

To calculate the Poynting vector anywhere in a plane orthogonal to the TSVs axis, it is efficient to use bipolar coordinates (α, β) (Fig.8.a) related to Cartesian coordinates by the relations:

$$x = a \cdot \frac{\sinh \alpha}{\cosh \alpha - \cos \beta} \quad ; y = b \cdot \frac{\sin \beta}{\cosh \alpha - \cos \beta}, \quad (23)$$

with: $\beta \in [0, 2\pi[$, $\alpha \in]0, 2\pi[$

Then, we can straightly deduce the new form of Poisson's equation:

$$\Delta \Phi = \left[\frac{\cosh \alpha - \cos \beta}{b^2} \right] \cdot \left[\frac{\partial^2}{\partial \alpha^2} f + \frac{\partial^2}{\partial \beta^2} f \right] \quad (24)$$

In this system of orthogonal curvilinear coordinates, we can still separate α and β .

Considering null the electrical field E , at the infinite, the non-uniform linear charge over each TSV are opposite (these charges are necessary to ensure the creation of an axial current into the TSVs). Although the calculation of E ($= -\text{grad}V$) is a little bit cumbersome, we get *in fine*:

$$\vec{E} = (V(\alpha_1) - V(-\alpha_1)) \left(\frac{\cosh \alpha - \cos \beta}{2b\alpha_1} \right) \vec{e}_\alpha \quad (25)$$

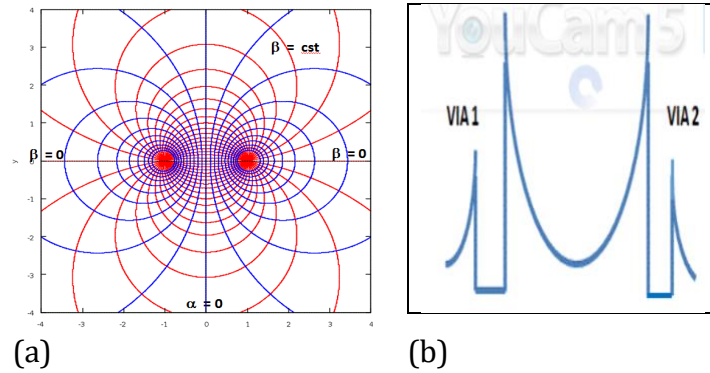


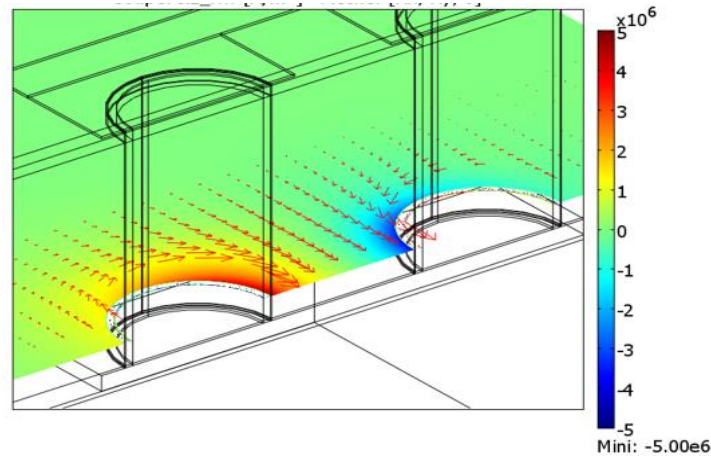
Fig.8 Left: Bipolar coordinates. Right: Poynting vector modulus, around 2 TSVs.

The modulus of the Poynting vector, of a dangling chain shape, takes particularly high values between the TSVs, but quickly goes to zero as the distance of sleeve surrounding the conductors, as shown in Figure 8 (b).

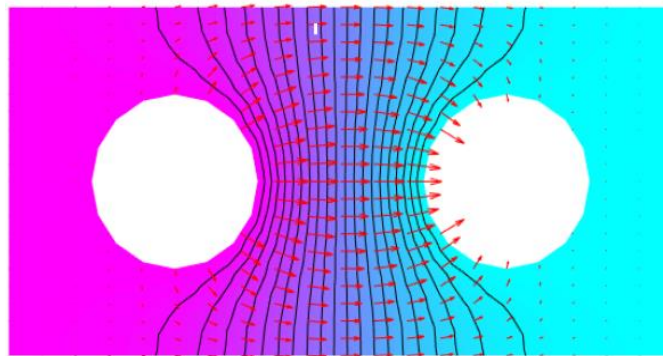
Magnetic fields created by the TSVs

Start by looking at the direction of the magnetic field in the approximation of quasi-stationary regimes. In this context, the electric field created by charges surface can be linked to the electrostatic field. If the wavelength is very large compared with the length of the circuit, the component axial E is negligible. This way, the electromagnetic wave carried by the waveguide that is the two-TSV line is a transverse electric TE of Oz axis. But for this type of wave, consider the component B in the cross-sectional plane orthogonal to E . Neglecting edge effects, this plan is a plan of anti-symmetry for the currents distribution. Quasi-steady and steady state being the limiting case of very low frequencies, the $\vec{E}$ and $\vec{B}$ fields are orthogonal everywhere outside the TSVs and regardless of the TSVs pitch. In other words, the lines of magnetic field are the equipotential lines in a straight section plane. The magnetic field lines are the α -iso lines: $\vec{B} = B(\alpha, \beta) \cdot \vec{e}_\beta$. What more is $\text{div} \vec{B} = 0$ (conservative flux). A direct consequence is the non-uniformity of current density in the cylindrical muffs of the two TSVs, In fact, the circle limiting a section of the conductor number 1 is a magnetic field line $B_1 + B_2$ of the whole line. The currents distributed in the two-TSVs line produce the same magnetic field (outside conductors) that two wires off-center from the axes of the cylindrical TSVs, centered at $x = + / - b$; $y = 0$; in fact, the positions of these equivalent wires coincide with the electrical images of the two cylinders brought to the potentials $+/- V$.

We can see, on Fig 9, the densities of current (COMSOL), not homogenous, and equipotentials or electric fields (MATLAB).



a) COMSOL Cut : J_z A/m²; arrow ($H_x, H_y, 0$)



b) MATLAB.

Fig 9: TSVs with metal sleeves: a) Current density and magnetic field distribution, b) Electric field and equipotentials

Importance of the RDL (Redistribution line)

For a closed two-TSVs line on a resistance R of a few Khoms, electrical energy is largely predominant compared to the magnetic energy. Lowering R increases the magnetic energy, via L , while the electrical energy remains constant. For small R , the magnetic energy outweighs, the transition taking place for a neighboring R value of the impedance of the vacuum. As static charges coexist with moving charges, the interaction between two parallel TSVs is both magnetic and electrical nature.

Can there be a transfer of energy between two neighboring circuits? If the electric power passes near the conductors, should we consider an energy transfer between two separate but neighboring circuits? This question must be considered carefully since the magnetic field in the vicinity of the resistance R is the sum of the magnetic fields created by the circuit via1 and circuit via2; $\vec{B} = \vec{B}_1 + \vec{B}_2$.

Both these contributions are found through the Poynting vector: $\vec{P} = \vec{P}_1 + \vec{P}_2$. Energy transfers are therefore to be taken into account for unconnected circuits. But, if the electromagnetic energy does not pass through the connections, it should be possible to counter these energy

transfers, for instance by interposing metal screens, although it is not so simple to pattern them, especially for high frequencies.

As a first trial, consider some crosstalk attenuation with ground plane structures. To the efficiency of ground plane structures in 3D integrated circuits, we inspect the connection system depicted on Figure 10. The wire acts for a TSV, the square conductor as a magnetic shield. The stimulus is a smooth Gaussian pulse, amplitude of 1, standard deviation: 10^{-11} s.

The “active area” - the target- acts as a lower device’s analog transistor, the metal line (“wire” – the aggressor) represents an upper device’s digital interconnect lines that induces coupling; the ground plate isolates both these two connecting metals. Presented are on Figure 10.b some first simulations. In fact, results point out the size dependent shadowing of the ground plate. Optimum has to be obtained; with a medium size grounded plane; a surge appears in the aggressor’s voltage. The system could act as a LC-like resonator; there is some energy oscillations between attacker and target: the attacker becomes, alternatively the target (the target becomes the attacker). Some dimension, inducing frequency-dependent resonance, could intervene (cf. cavities or surge in low and middle frequency of RLC circuits). We have seen also the inefficiency of a floating shield.

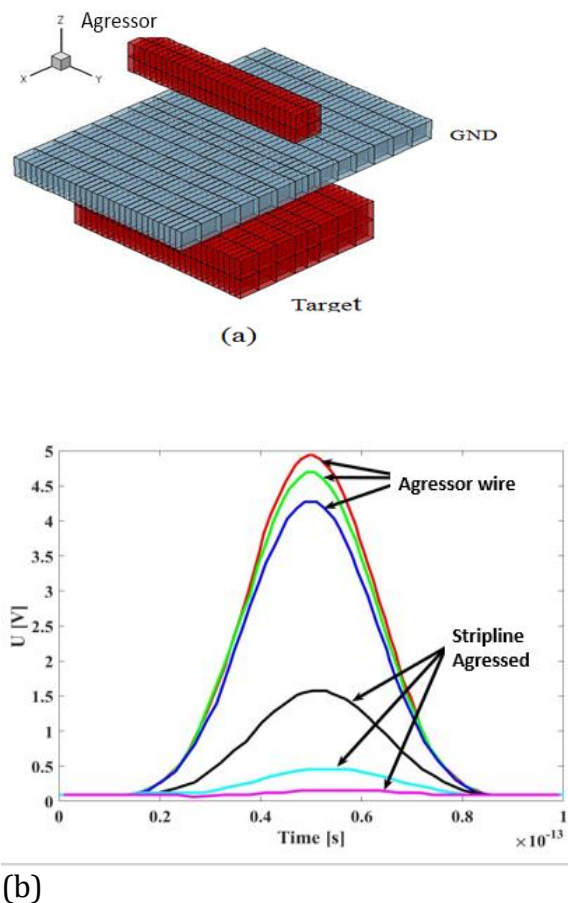


Fig. 10. (a) Multilayered structure, (c) Typical response to a gauss pulse

To come back to via couplings, to decrease the crosstalk, a large TSV diameter should be avoided, if there are no problem concerning with the cost and fabrication. TSV radius is an important parameter in this cylindrical via for the designer or the manufacturer when examining the density of TSVs. The copper via, the insulating layer and the silicon substrate constitute a metal-insulator-silicon capacity (MIS.). Thickening the oxide insulator is favored unless they exceed the density requirement or fabrication craft in some specific circuit parts. Enlarging the distance between aggressor and victim TSVs and decreasing the distance between signal and ground TSV can always reduce the crosstalk noise. In terms of TSV connection bus, appropriately arranging ground between signals should always bear in mind. From preliminary results, crosstalk in a TSV interconnecting network could be improved, leading to enhanced performance of 3D circuits and systems+.

So far, it is often discussed structures in which the distances between the TSVs are uniform. In other words, the TSV on the ground is at the same distance from the TSV signal as the distance from the victim TSV. Reducing the signal-to-ground distance and increasing the signal-to-victim distance, all both reduce the mutual inductance. In addition, they will have the effect of reducing

capacity. As we know, weak capacitive coupling and weak inductive coupling lead to low crosstalk.

Design guidelines should be proposed for designing TSVs in the signal and power distribution networks (PDNs) as well as for TSVs in variable capacitor applications.

Perhaps, another way is to create magnetic fields with chosen orientation, for instance in performing an eccentric cylindrical hole into a full metal via. With the superposition theorem, in a perpendicular section of the final TSV, it is easy to show that the magnetic field in the created cavity of the initial via is perpendicular to the O_1O_2 segment, O_1O_2 being center points in this section, respectively on the two cylinder axes of the full and empty cylinders; mediator plan of O_1O_2 is a symmetric plan of this extruded via.

Perhaps, it should be useful to build honeycomb-like layers, dedicated to these vias, minimizing their heights. Therefore, the choice of the material of the liner of the alveoli and their variation in size should be crucial.

We discuss also some aspects of cooling, via heat pipes in a recent paper [21]. In effect, in such 3D complex circuits, it is crucial to identify hot spots that are susceptible to localized voids formation, debonding, so as cracking due to physical phenomena such as stress, temperature excursions, and electromigration.

Finally, we introduce below an opening on stress analysis on vias, thus complementing our platform, built under MATLAB/GUI, an ante-processor for 3D RF commercial circuits.

Stress on via [22-27]

Numerical Experiments are conducted with a newly developed multichamber process that enables substrate preprocessing, catalyst metal deposition, and nanocarbon thermal chemical vapor deposition (CVD) growth in a vacuum batch. The base vacuum pressure was 3.0×10^{-5} Pa. A TaN layer (15 nm thick) was deposited on SiOC dielectric/Si substrate having via holes with a diameter of 160 nm in SiOC dielectric layer. The substrate was subjected to Ar plasma etching in a preprocessing chamber to etch the TaN surface very slightly. Next, layers of Ti (1.0 nm thick) and Co (4.0 nm thick) were deposited in a vacuum vapor deposition chamber. The substrate was then moved to a thermal CVD chamber, where it was heated to a growth temperature of 450 °C in a vacuum. Subsequently, a mixture of acetylene and argon gas was introduced at a gas pressure of 1 kPa. The growth time was 30 min. The CVD growth temperature was as low as 450 °C.

Properties of via

The thermal conductivity of these vias was measured to be $1.0 \text{ Wm}^{-1} \text{ K}^{-1}$ when horizontal grapheme layers were formed and $10 \text{ Wm}^{-1} \text{ K}^{-1}$ when they were removed. These results indicate that, as expected, the conductivity parallel to the graphene plane is several orders of magnitude higher than that perpendicular to the graphene plane.

Simulation Setup

As a target of simulation setup used in comparison, a TSV used has a diameter, height, landing pad size, and liner thickness of 5 μm , 30 μm , 6 μm , and 125 nm, respectively. The SiO_2 is used

as a baseline liner material, Material properties used are as follows: CTE (ppm/K) for Cu = 17, Si = 2.3, and SiO₂ = 0.5; Young's modulus (GPa) for Cu = 110, Si = 130, and SiO₂ = 71.

Simulation Results

To simulate the stresses, as an initial setup used in TCAD tool [28], a single TSV of 5 μ m width and 20 μ m length with a thin layer of 0.15 μ m SiO₂ as a liner is used. . The TSV is filled with copper using isotropic method; then it is annealed for 10 seconds at 1200° C.

Figure (11) depicts some simulation results. As shown, the maximum stresses is concentrated at the top and the bottom of the via between the interface of the TSV and the substrate. The X-direction exerts more stress than the Y and Z directions.

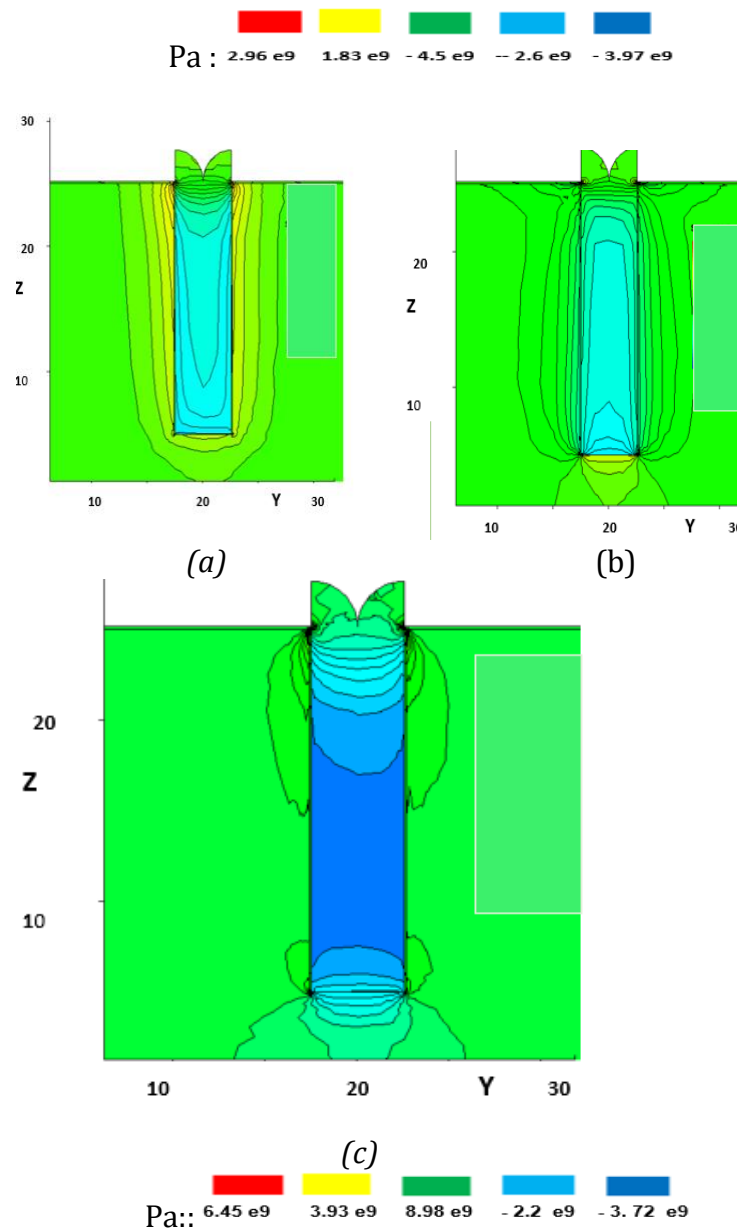


Figure 11: Simulation Results (Pa)on vias (a) Stresses in X direction (b) Stresses in Y direction (c) Stresses in Z direction.

CONCLUSION

We start with considerations for calculating some general impedance between two embedded points in 3D substrates. This spreading or transfer impedance definition is threefold: it concerns the first order electrical regime, the background electronic noise, and temperature spreading. Using the strong analogies between them, we embedded a general and unified algorithm in a MATLAB/GUI environment. Moreover, we give also some first tracks to consider the interconnects, especially the vias, and point out that EM energy localization, for instance between vias, and the vias with the substrate; it can help to go further that the classical studies of electromagnetic couplings, to amend EMI drawbacks. This should be, *in fine*, finalized through a global SPICE-like modelling. Following work is the consideration of non-ideal shapes, such as tapered vias or non-concentric and partially axial coverage of depletion regions.

For the near future, we should consider the study of tri-dimensional noise coupling, and the possible correlations between noise sources, especially at the nano-scale, where these latter, although autonomous, can be coupled.

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