

Causal Power Plane Models

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Causal and frequency-dependent models and simulations are important for today's high-speed signal integrity simulations. But are causal models also necessary for power-integrity simulations? When we do signal-integrity eye-diagram simulations, we define the source signals, so if we use the correct causal models for the passive channel, we will get the correct waveforms and eye reduction due to distortions on the main path and 'noise' contributions from the coupling paths. The details of the simulated waveforms will make a difference in the accuracy of the eye closure; in such cases the causality of the passive component models is important. At the other end of the spectrum, when we do point-of-load PDN simulations, causality today is much less important, not because we don't care for the accuracy of the result, but primarily because the excitation signal, the signature of the current demanded by the load as a function of time, is usually not known very well. One of the areas in between, which involves PDN models and requires good causal models is when we do SI-PI co-simulation. For instance, when we want to simulate the eye closure due to simultaneous switching noise and PDN noise on memory signals, the memory excitation signals are set by us and therefore it makes sense to use good causal models for the PDN to get accurate results. This article shows you a few important points how you can achieve it.

For any interconnect, which has non-negligible delay for the particular application (in technical terms: which is not electrically short), we need to use a simulation model that describes not only the interconnect impedance, but its propagation delay as well. Extrapolating from the world of lumped circuits, cascaded segments of series inductance and parallel capacitance is a simple and brute-force approach to model both impedance and delay. A simple ladder circuit shown in *Figure 1* can be used to simulate the approximate behavior of package pins, connector pins, traces and cables.

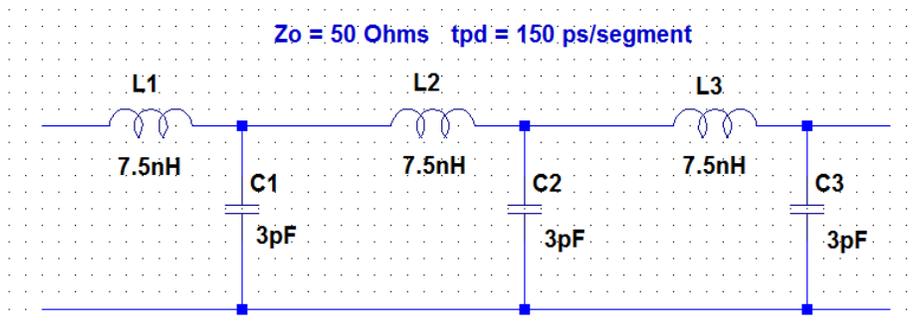


Figure 1: Ideal one-dimensional LC ladder.

This model represents a lossless interconnect with 50-Ohm characteristic impedance and 150 ps delay in each segment, a total of 450 ps delay for the three cascaded segments. The model works well as long as the equivalent rise time of the signal we want to pass through this circuit is much longer than the delay in one segment. If all what we want to accurately simulate is the rise time of the signal passing through this circuit, this ratio can be as low as three to four. If we need good accuracy of the entire waveform, say down to one percent, we need a ratio of ten to twenty. If we take the tighter accuracy estimate of twenty, the signal passing through this example network should have a rise time of $20 * 150 \text{ ps} = 3 \text{ ns}$, or longer. This limits the bandwidth to approximately 100 MHz. If we need wider bandwidth, we can proportionally scale the L and C values. For instance, using $L = 0.75 \text{ nH}$ and $C = 0.3 \text{ pF}$ will produce a 50-ohm interconnect model with a bandwidth of 1 GHz, but if we need the same 450 ps end-to-end delay, we will need to use ten times more, namely thirty cascaded segments. This model is so simple that any circuit simulator can take it, but to achieve high accuracy and high bandwidth at the same time, we may end up with a lot of circuit nodes and considerable run time. Of course for the ideal lossless case we have in *Figure 1*, we could just use a single SPICE Tline element with the required characteristic impedance and total delay, but that way we would lose the horizontal connectivity (see more on it in [1]).

Package or connector pins, traces and cables are one-dimensional transmission lines in the sense that the majority of the signal will only propagate along the conductor. Power planes are two-dimensional transmission lines: any signal (mostly the power noise) can and will propagate in any direction on the plane. To model this two-dimensional nature, we can take the ladder equivalent circuit from *Figure 1* and turn it into a two-dimensional grid circuit. *Figure 2* shows the conceptual circuit diagram.

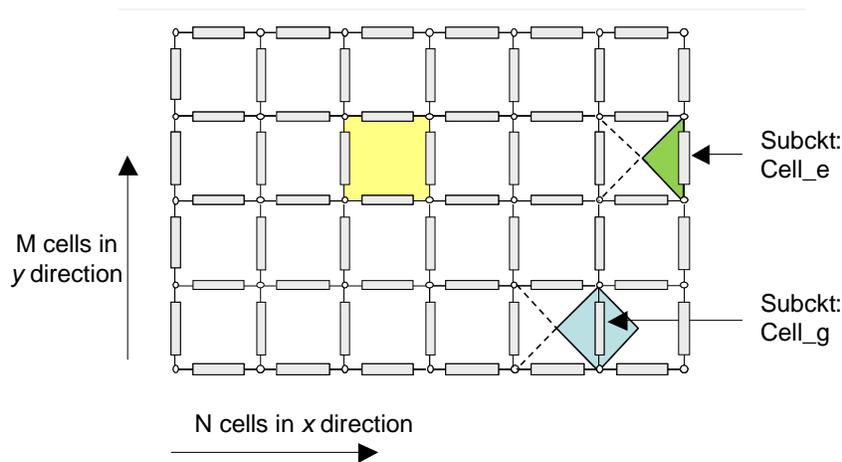


Figure 2: A two-dimensional SPICE grid network to simulate a pair of rectangular planes.

Each subcircuit in the grid can be one LC segment from *Figure 1* with appropriate L and C values. The calculation of the subcircuit element values is explained in [2]. If we have plane outlines or slots/cutouts in the planes that can not be easily approximated with

rectangular shapes, the same grid concept can be extended to utilize variable-size grid cells [3]. The simulation models based on grid equivalent circuits will provide results at each of the circuit nodes; in between circuit nodes we need to rely on interpolation. Alternately, for simple plane shapes, analytical formulas can also be used to simulate the self and transfer impedance of planes at any arbitrary point [4].

Of course, today we have a number of commercial 2.5D tools available that can simulate two-dimensional structures [5], [6], [7]. We could also use 3D tools, though simulating electrically very large structures that way is not really feasible.

Turning our attention back to causality: the model in *Figure 1* is fully causal, but it is not only band limited, it represents only lossless interconnects. This means no series resistance representing the conductor losses, no parallel conductance to capture the dielectric losses, and as a result, the capacitance and inductance values are frequency independent constants. If we want causal, lossy models, each segment needs to be represented with an RLGC network, where all four parameters are frequency dependent.

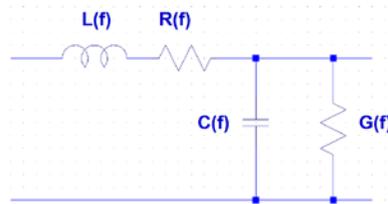


Figure 3: One segment of a generic, causal, lossy interconnect model.

For the frequency-dependent $C(f)$ capacitance and $G(f)$ conductance we can use one of the causal dielectric models, for instance the wide-band multi-pole Djordjevic-Sarkar model [8]. The material constants can be taken from the laminate data sheet or from measurements [9]. For instance, based on the data sheet [10], a glass-reinforced FR406 4-mil core laminate at 1 GHz will have a D_k and D_f value of 3.95 and 0.0161, respectively. If we chose an unreinforced HK04J25 1-mil polyimide laminate [11], the D_k and D_f data sheet values at 1 MHz are 3.5 and 0.005, respectively. If one segment of the plane model represents one square inch area, the FR406 laminate model at 1 GHz will have 226 pF capacitance and 0.022815 S conductance; the HK04J25 laminate model at 1 MHz will have 800 pF capacitance and 0.0257 S conductance. For a causal model we need these parameters as a function of frequency, which can be readily calculated based on the chosen wide-band multi-pole model [12]. *Figure 4* and *Figure 5* show the causal dielectric parameters as a function of frequency for the FR406 and HK04J25 laminates. On both figures the red dots on the left chart indicate the values provided by the data sheets.

To obtain the series elements of the equivalent circuit, we can start with the DC sheet resistance and asymptotic high-frequency spreading inductance of the planes. The DC sheet resistance depends on the thickness and conductivity of the conductor.

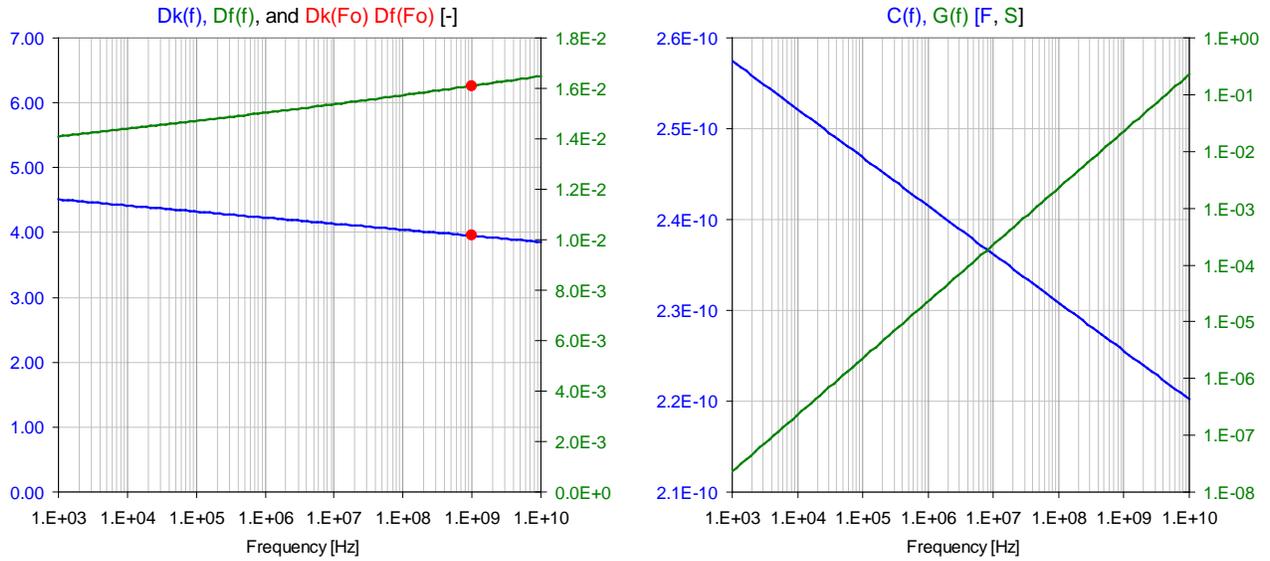


Figure 4: FR406 $Dk(f)$ and $Df(f)$ as a function of frequency (on the left) and $C(f)$ and $G(f)$ as a function of frequency for a one square inch of laminate (on the right).

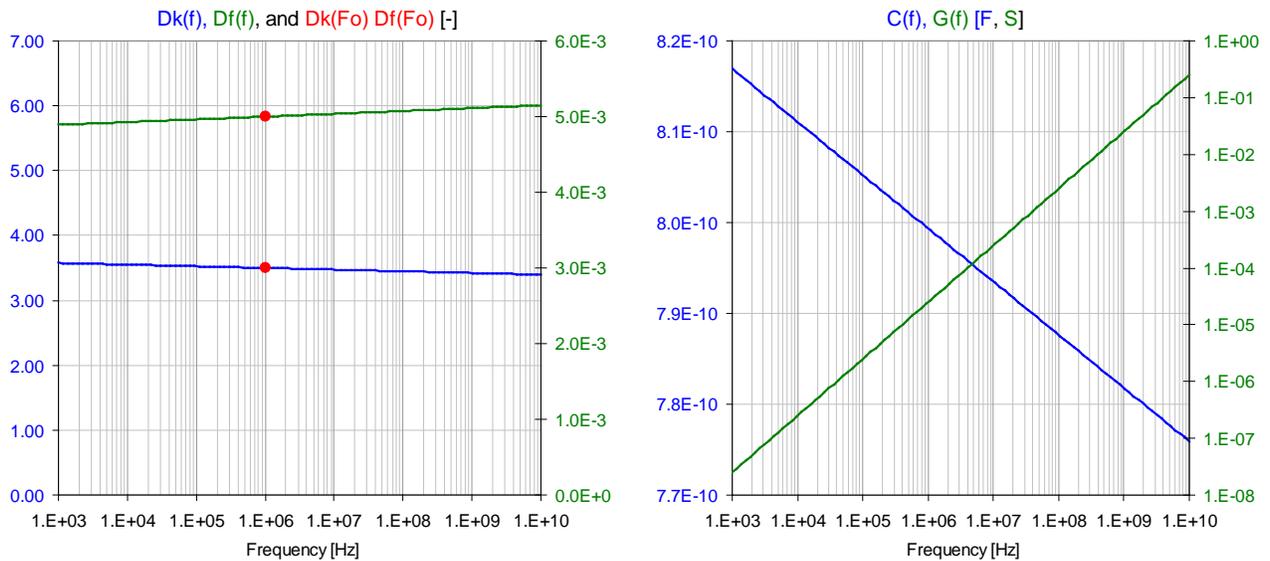


Figure 5: HK04J25 $Dk(f)$ and $Df(f)$ as a function of frequency (on the left) and $C(f)$ and $G(f)$ as a function of frequency for a one square inch of laminate (on the right).

Assuming one-ounce (30 μm) pure copper for each plane, the sheet resistance is approximately 0.6 mOhms for each square in each plane layer. The high-frequency spreading inductance is approximately 33 pH for each mil dielectric thickness, resulting in a 132 pH for the 4-mil FR406 example and 33 pH for the 1-mil HK04J25 laminate example. Note that the resistance is given so far at zero frequency, whereas the starting point of the inductance is calculated at infinite frequency. The calculations are

based on the assumption that at DC the current penetrates vertically the plane thickness uniformly (large skin depth), but at infinite frequency the current flows only on the facing surfaces of the planes (very small skin depth). At interim frequencies the current penetration depends on the ratio of the skin depth and conductor thickness. At each frequency we can calculate an effective plane separation, which is the dielectric thickness and the sum of the equivalent layer thickness values in the upper plane and lower plane. Similar calculations can be applied to obtain the AC resistance on the planes. If needed, corrections for surface roughness can be added, too.

The resulting causal grid model can be simulated in circuit simulators that can take frequency dependent components. Unfortunately the original free Berkeley SPICE does not have this option, but as long as we need only AC simulations, there is a straightforward, though tedious workaround. We can calculate the actual RLGC values separately at each frequency point and then run SPICE AC simulations at those single frequencies. From the output files we can put together the full frequency dependent causal response.

To illustrate the potentials of causal and frequency dependent plane models, *Figure 6* shows how the lossy causal model captures the gradual suppression of modal resonances as the laminate gets thinner. We can see that with 10-mil (0.25 mm) plane separation the impedance swings are substantial at the resonance frequencies, and as we go to 1-mil or thinner, we get a significant reduction. As signal vias may go through these power-ground cavities, capturing the plane resonance correctly in frequency and magnitude becomes important, especially when we combine the plane models with signal interconnects for SI-PI co-simulation.

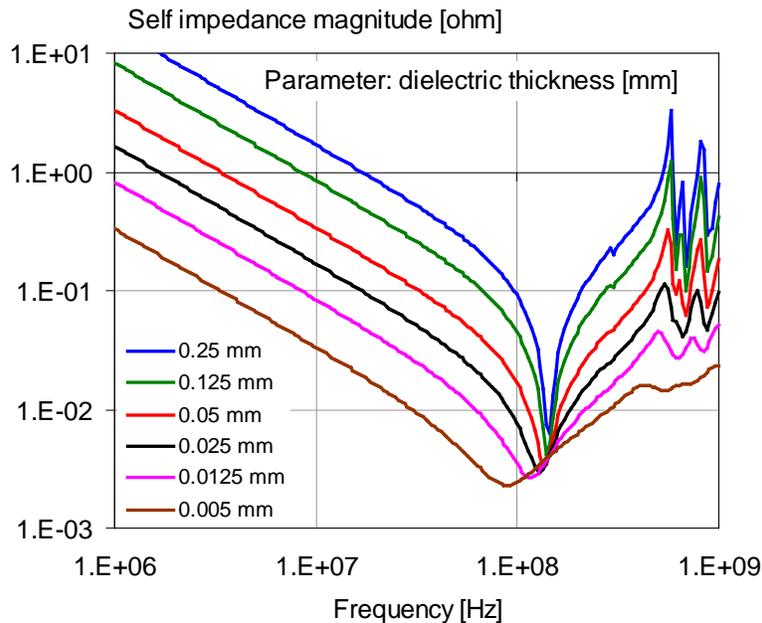


Figure 6: Self impedance magnitude in the middle of a 10" x 10" FR4 dielectric sheet between one ounce copper planes. The parameter is the dielectric thickness.

In a future column we will show the correlation between simulated and measured data for different test boards with different laminates. A peek preview of one such board is shown in *Figure 7*, courtesy of DuPont. The 6" x 6" open-edge board has 1-mil HK04J25 laminate with one ounce copper.

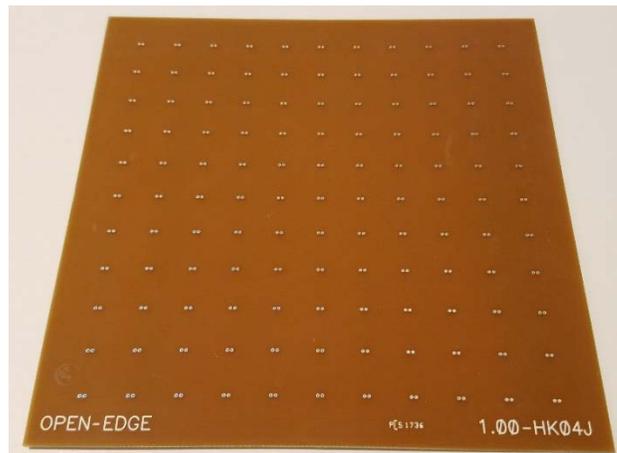


Figure 7: Photo of a 6" x 6" laminate test board, courtesy of DuPont.

Next time when you do memory eye simulations with PDN effects included, make sure your models are causal and your tools can properly handle them.

References

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