Analyzing Distortion Contributions in a Complex Device Model

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Summary. This paper studies how distortion contribution analysis is affected by the internal structure of the device models. It is shown that to keep the distortion contributions tractable and physically meaningful we need to lump the contributions so that they resemble the contributions of a classical transistor pi-model. The technical challenges related to this are also discussed.

1 Introduction

Until now, distortion contributions have been analyzed mostly fully analytically, using symbolic Volterra analysis and simplified schematics[1,2]. The authors have proposed a fully numerical distortion contribution analysis method called Volterra-on-Harmonic-Balance (VoHB) [3,4], that both builds the polynomial models needed for Volterra analysis, and propagates the distortion contributions to any chosen node. VoHB proceeds in the following steps: HB is run to obtain voltage and current spectra in all nodes and branches. Using these, a polynomial model is built for each non-linear VCCS and VCQS. Then, a linearized network is built (using the linear terms of the fitted polynomials), and - using the direct current method - the response of the injected distortion currents is calculated in a given node.

VoHB operates at VCCS level and does not need modifications into the device models. However, the output contains the response of all non-linear VCCS and VCQS found inside the device model, so that in addition to the dominant sources it includes reverse biased pn junctions, or parasitic devices of BJTs, for example. Many of these insignificant terms can be masked away simply based on their low magnitude, but another problem has emerged.

Many new device models are distributed as an executable Verilog-A code, which is automatically converted into a model structure. The syntax of verilog-A makes it possible to write models whose structure differs notably from traditional device models and is not necessarily optimal for Spice-like simulators, or distortion contribution analysis in particular. For example, Fig. 1 shows a rather typical example where the transistor's Ids source is split into two by generating an intermediate node in the middle of the source. From terminal current point of view this modification is absolutely ok, but it scrambles the distortion contribution analysis. VoHB calculates the contributions of all the sources, and adding an equivalent (and equal) distortion current sources parallel to Ids and Idsx as in Fig 2a generates two large distortion contributions that cancel each other. The net current sum for example in the collector terminal is still correct, but the physical meaning and intuition is lost: the names of the sources do not mean anything to the engineer, any more, and the mutually cancelling contributions have no physical meaning.

For the above reasons, there is a clear interest to reduce the entire internal structure of a complex device model (Fig. 2a) into something resembling a classic transistor pi-model (Fig. 1a) to keep the distortion contribution analysis results tractable. The idea is to lump all the input and output-related non-linear currents together, as shown in Fig.2b. Now the designer can again clearly recognize the effects of input and output related conductive or capacitive non-linearities.

Building the equivalent distortion current model is straight-forward. HB simulation is run, and terminal distortion current spectra Id, Is and Ig are recorded. Then equivalent current sources are fitted so that the terminal distortion currents are correctly modeled by these imaginary distortion sources. These sources are fitted using the spectra of terminal currents and intrinsic node voltages.

Even in a lumped model the terminal distortion currents consist of distortion generated in several sources. For example, the current in the drain terminal comes from the gm-element Ids(vgs,vds), and drain charge Qds(vgs,vds), both of which are controlled by intrinsic vgs and vds voltages. Moreover, the current from Qds is proportional to the tone frequency Ω, while Ids has a transit delay τ that rotates the phases of the tones by exp(-jωτ). Hence, i_d would be described as

$$\begin{aligned} i_d &= i_{gm} + i_{qds} \\ &= diag(e^{-j\Omega\tau}) \cdot gmpoly(vgs, vds), \\ &\quad + diag(j\Omega) \cdot qdspoly(vgs, vds) \end{aligned} \qquad (1)$$

where diag() is a diagonal matrix, Ω is the frequency of a given tone, and gmpoly() and qdspoly() are model function matrices. In the model function matrices each row corresponds to one frequency in the spectrum, and each column corresponds to one $vgs^{i}*vds^{j}$ product term in the polynomial model.

Similar equation can be written for the gate current ig:

$$\begin{aligned} i_g &= i_{gpi} + i_{qgs} \\ &= gpipoly(vgs, vds) \\ &\quad + diag(j\Omega) \cdot qgspoly(vgs, vds) \end{aligned} , \qquad (2)$$

where gpipoly() corresponds to the possible conductive part of the input current (needed in BJTs) and qgspoly() models the current caused by the input charge.

Solving the polynomial coefficients from (1) has some technical challenges. Currently, VoHB has igm and iqds available separately, and can fit the two polynomials independently. This keeps the number of unknowns in some bounds. In a lumped model we must fit (1) simultaneously, which increases the number of unknown coefficients. As we can fit only as many coefficients as there are equations this approach may not be possible using a 1-tone spectrum, but we must use 2- or 3-tone excitations. The jΩ emphasis also means that the effect of capacitive lower harmonics is attenuated, and more easily buried underneath the effect of conductive non-linearities. Second, the frequency response diag(jΩ) of capacitive current is known a priori, but the transit delay tau of the ids source is not necessarily known, if we want to keep the analysis independent of the device models. Hence, tau needs to be found by iterating (1). Third, the above only gives the equivalent polynomials that can be used to calculate the nonlinear distortion currents i_{NL}. Addition to this, we need to build the linearized circuit model to propagate the currents to a given node. For this reason we also need to find linear models for all VCCS elements in the original model.

As an example, we took the MET model [5] of Freescale's LDMOS power transistor MRF21030. Its output drain current consists of currents from three sources: the gm source Ids(Vgs,Vds), drain charge Qds(Vds), and gate-drain charge Qgd(Vgd). Ids strongly dominates the total drain current, and Qgd is insignificant. In a simple example, we tried to model total drain current Id as a sum of Ids + IQds. The combined model matrix is very badly ill-conditioned (cond ~3e13), and Qds fits poorly. Much better results was obtained by iterating a couple of times in a loop, where Ids was first estimated, and substrated from Id before fitting IQds. This reduces the order of the fitted system and improves especially the fitting of the capacitive currents that are buried underneath the dominating Ids current. The condition number of the Ids fitting matrix drops to ~ 1e7, which is mostly set by the heavy correlation betweenVgs and Vds signals.

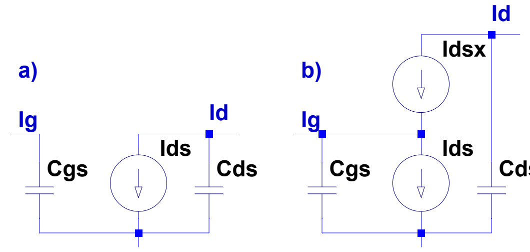

Fig. 1 a) Transistor pi-model, b) example structure that may result when automatically generated from a Verilog-A source.

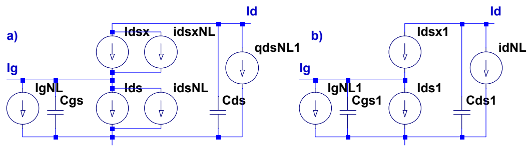

Fig. 2 a) Distortion current sources next to each VCCS, b) lumped distortion current sources.

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