Wideband Spice-compatible modeling of packages

Marek Schmidt-Szałowski
BL RF Small Signal, NXP Semiconductors, Netherlands
October 29, 2012

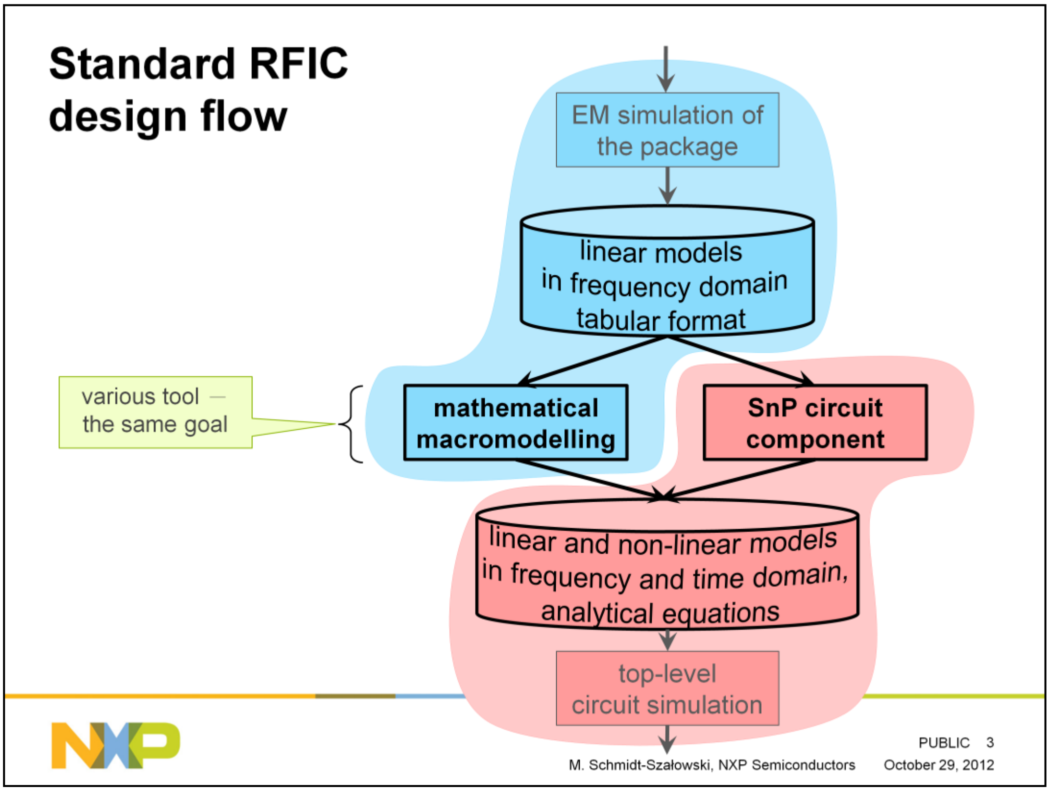
Author: Dr. Marek Schmidt-Szałowski
Affiliation: Business Line RF Small-Signal, NXP Semiconductors Netherlands
Address: room BZ-2.074, Gerstweg 2, 6534AE, Nijmegen, NL
Phone: +31 24 353 6858
Email: marek.schmidt-szalowski@nxp.com

Presented on 29-11-2012 on EuMIC2012 in Amsterdam, paper EuMIC03-3.

Outline

- Definition of the problem
  - Application scope, motivation, and required functionality
- Description of the modeling method
  - Assumptions and implementation
- Two test cases

The main topic: extraction of a Spice-compatible lumped-circuit model from wideband multiport S-parameters of RF package.



The extraction method discussed here serves as a link between an EM simulator and a circuit simulator. The EM simulator delivers a package model in a form of S-parameters defined at discrete frequency points. A direct use of such a model in the circuit analysis is limited to linear simulations in the frequency domain and, if accurate enough, in the non-linear DC simulation. In order to enable time-domain simulations, an analytical model must be available, like the one described here. Extraction of such a model is from the numerical point a challenging problem, mathematically equivalent to Padé approximation.

There are already various commercial tools available, including those that are integrated with either the EM simulator or the circuit simulator, as well as stand-alone ones. To the author's knowledge, all the currently available solutions are meant as general-purpose modeling tools applicable to very broad range of linear devices. The extracted models have purely mathematical character. Neither their topology nor extraction strategy reflects the physical relationships governing the devices to be modeled.

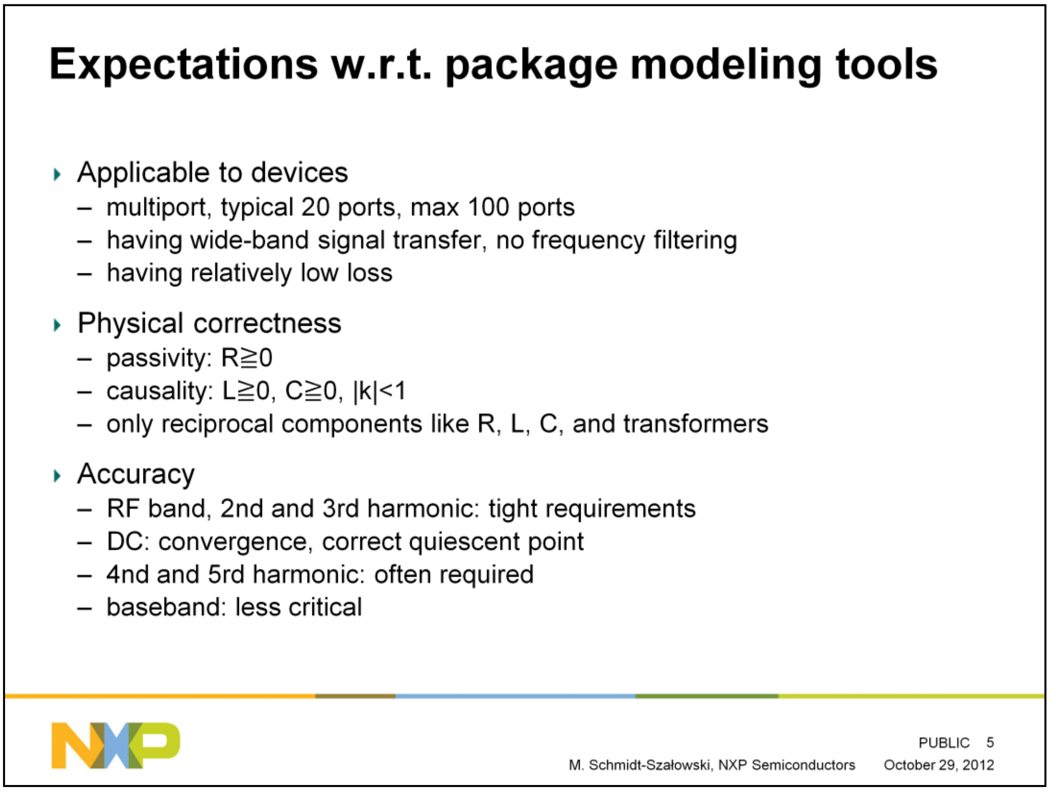
Shortcomings in the current way of working

- Pure mathematical modeling.
  - Generally seen as incomprehensive and disconnected from the physics.
- Passivity enforcement remains a challenge.
  - If not enforced: DC and transient simulations may fail to converge.
  - If enforced: the model accuracy degrades.
- "DC leakage"
  - Tiny fitting errors at 0 Hz result in leakage resistance between the IO's.
- Large sensitivity to glitches in the input data.
  - The input data often have to be "polished" before running the extraction.
- Extrapolation in frequency domain not possible.
  - Models behaves strange at frequencies other than those in the input data.
- Noise contribution often wrong.
  - Controlled sources extensively used in the netlist have no noise model.

The purely mathematical character of the currently available modeling tools leads to their bad publicity within the RF design community. Incomprehensively constructed netlists, passivity-related issues, inability to extrapolate in the frequency domain, and the necessity for labour-intensive pre-processing of the S-parameters are commonly heard complaints. There however two other that often overlooked both by the EDA vendors as wells as by the RF designers themselves.

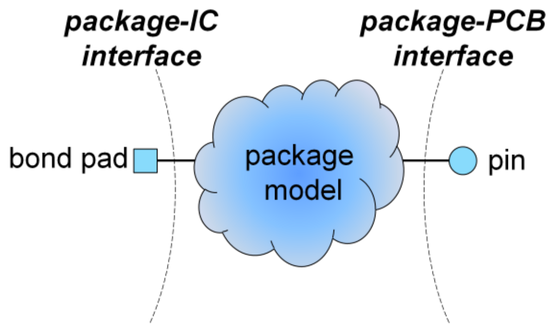
1. The first of these drawbacks, called here "DC leakage", is an inevitable result of residual fitting errors, which although sometimes well below the -60dB level can substantially deteriorate the accuracy of the DC simulations. Imagine that some of these residual errors are equivalent to a 1MΩ resistance between pins differing in voltages by 1V. That will cause an error current of 1μA whish is sufficient to significantly influence the base current of a HBT transistor. This phenomena is magnified by the fact that the IO's of an analog IC may sometimes have extremely high DC resistance, which is orders of magnitude greater than 50 Ω to which the modeled S-parameters are normalized. Obviously, the DC leakage as described here will not be the only effect of the fitting inaccuracies of the model. Some other fitting residuals can translate into excessive series resistance of the IO's or a "mutual resistance" (like mutual inductance but observed in the real part of Z). Fortunately, typical analog IC's are not very sensitive to these inaccuracies, except for very low-ohmic power devices.

2. The other drawback becomes evident when the noise contribution of the package is being calculated during frequency-domain noise analysis. The netlist representations of mathematical models involves often a lot of controlled sources. Each of these circuit components itself is non-reciprocal and only combined all together they constitute a reciprocal package model. This sort of netlist models work fine as long as only the impedance (or a related parameter) needs to be simulated. It fails, however, during the noise analysis for which the simulator must calculate the contribution of each individual components. The non-reciprocal components have no (physical) noise model and most of circuit simulations treats them as noise-less devices. In theory, this problem can be fixed by moving the noise calculation from the component level to the subcircuit level. It would mean that the simulator calculates first the Y-matrix of the package model and then, assuming purely thermal origin of the noise, derives the noise correlation matrix CY from it. Unfortunately, not all the leading simulators support this calculation method of the noise contribution.



Considering the challenges of mathematical models, there is a need for a more physical approach to the package modeling. Before starting to develop a new method it is good to summarize the expected functionality and performance. After all, RF packages are relatively narrow class of passive devices and share a lot of common features. This helps in making proper choices at the very beginning.

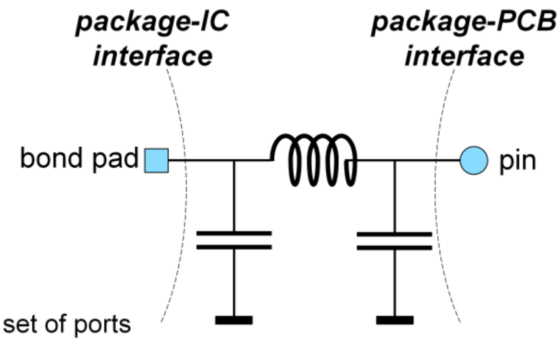
Design of the new modeling method



- Assumptions
  - As simple as possible
  - Tailored to package modeling
  - Maintain physical interpretations as much as possible
  - S-parameter table from an EM simulator as the target data

Instead of taking one of general methods for modeling of linear time-invariant systems, the author chose to start with a simple physical network and add those features that are necessary for modeling of packages.

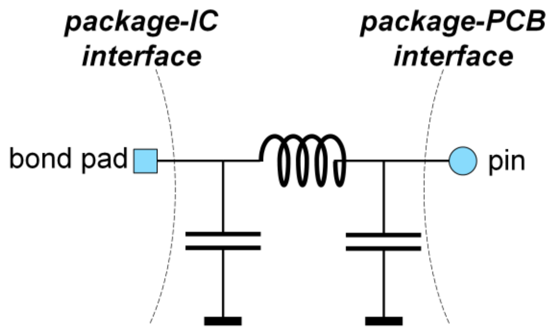
Starting point: Π-section



- Features
  - Comprehensive netlist
  - Clear distinction between two set of ports
    - at the package-IC interface
    - at the package-PCB interface
  - No positive gain
  - Easy to extract

One of the most obvious features of an RF package if its ability to interconnect an IC with the PCB. One can identify two clear interfaces: package to IC and package to PCB. The ports located at these two interfaces will be treated differently throughout the whole modeling procedure.

Starting point: Π-section

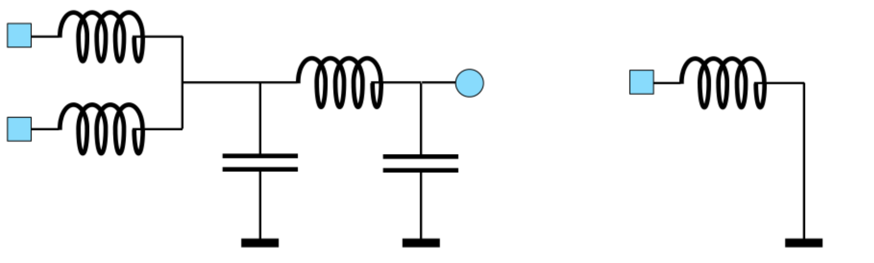


- But how to deal with
  - Differential ports
  - Irregular IO's
    - single pin — multiple bond pads
    - ground bond wires, no explicit ground pin
  - Conductor and dielectric losses
  - Coupling between IO's
  - Distributed character of the IO's
  - Numerical challenges resulting from nearly shorts and nearly opens

A Π-section model is one of the most popular low-frequency models of interconnect. It is, however, clearly too simple to model a whole RF package. In the following slides, several enhancements will be added.

Enhancements of the Π-section model

- Differential ports
  - not needed in 99% of encountered cases
  - all ports referred to the ground, global or local
- Irregular IO's
  - Solved by adding series *L* sections



single pin — multiple bond pads grounding bond wire

In the current form, the model is suitable only to packages with single-ended ports. It means that all the ports at the IC-package interface must share the same reference terminal. The same holds for the ports at the PCB-package interface. Till now, it has never been a disadvantage of any kind. Moreover, as soon as the demand for packages with differential ports or more irregular topologies shows up, the underlying model equations can be easily extended without significant computational efforts.

Typically each net representing an IO of the package has two ports: one at the IC-package interface, the other at the PCB-package interface. However, the package model has to deal with IO's having more than one bondwire attached to the same pin. Moreover, for the sake of improved RF grounding, the package may include downbonds, ie. short bondwires connecting a bondpads with the local ground of the package. Both situation can be easily accounted for by adding more inductive components in the equivalent circuit. Note that the modeling of downbonds is limited to the first-order effects and the self-resonance of the downbond will not be modeled.

Enhancements of the Π-section model

- Conductor and dielectric losses
  - Solved by adding the skin loss model and dielectric relaxation model

$$Z = j\omega L_{ext} + R_{DC} M_{skin}(\omega) \qquad Y = j\omega C_{\infty} + j\omega C_{\Delta} M_{diel}(\omega)$$

with four unknowns to be extracted
- DC resistance R_{DC}
- external inductance L_{ext}
- high frequency capacitance C_{∞}
- excess capacitance C_{Δ}

any combination of positive R_{DC} and L_{ext} is physical

any combination of positive C_{∞} and C_{Δ} is physical

and two pre-defined complex functions
- M_{skin} — derived from surface impedance model
- M_{diel} — equivalent to Djordjevic-Sarkar model

  - In the netlist this analytical equation is replaced with a lumped models
    - conductor losses: continued-fraction expansion
    - dielectric losses: multi-pole Debye model

PUBLIC 10

M. Schmidt-Szałowski, NXP Semiconductors October 29, 2012

RF package are usually built of high quality conductors and dielectrics but at high frequencies even they cannot be treated as lossless. For this reason, the ideal L and C components use so far need to be replaced now by lossy L and lossy C with frequency dependent real and imaginary part of Z and Y.

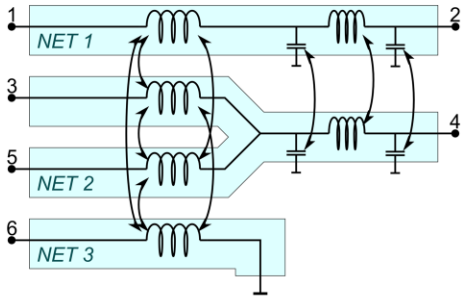
The loss model used for the conductors is based on the equation derived from surface impedance model and includes two model parameters: the DC resistance (R_{DC}) and the external inductance (L_{ext}). The dielectric losses are described using Djordjevic-Sarkar model controlled by two parameters: the capacitance at f=∞ (C_{∞}), and the capacitance change between f=0 and f=∞ (C_{Δ}). There are also two complex coefficients M_{skin} and M_{diel} with a predefined frequency dependence.

Both models are well described in the references. It is worth emphasizing that in the formulation used in this work the lossy impedance Z and admittance Y are linear function of the corresponding model parameters. Another observation is that the real and imaginary parts of Z and Y remain positive as long as the model parameters are positive. These two features simplify the extraction routine.

The analytical expressions M_{skin} and M_{diel} include infinite number of poles and zeros. Upon composing the netlist the expressions for Z and Y are approximated with proper lumped models.

Enhancements of the Π-section model

- Coupling between IO's
  - Solved by replacing single *L* and *C* components with arrays of them.
  - Sparse matrix methods reduce the complexity of the problem.



- Distributed character of the IO's
  - An *LCLC* ladder accurate at least up to **¼ λ**.
  - Sufficient for 99% of encountered cases.

PUBLIC 11

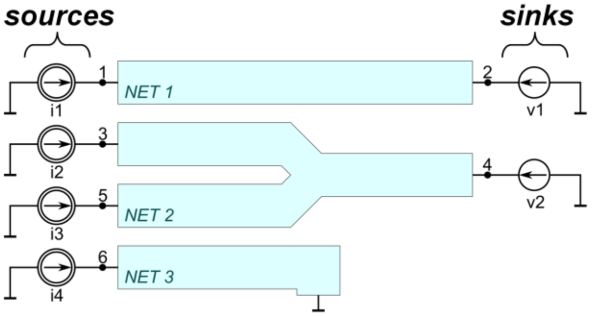
M. Schmidt-Szałowski, NXP Semiconductors October 29, 2012

The model considered up to now described the properties of individual IO's. In order to describe the interactions *between* the IO's the mutual Z and Y components must come into play. As a result, all the impedances and inductances that were scalar up to now should be extended to matrices. As shown in the figure, the complete circuit model will consist of two matrices of coupled series impedances and two matrices of coupled shunt admittances. Note that the two impedance matrices may have different size since one of them includes additional inductors that model the irregularities of the topology. Owing to a careful modeling of losses and mutual coupling, the real and imaginary parts of the impedance and admittance matrices are always symmetric positive definite.

The L-C-L-C topology has a fundamental limitation in terms of maximum electrical length that can be accurately modeled. The end of the frequency range where the model is accurate lies in practice between the ¼λ and ½λ frequency. The only way to extend it is to add more L-C sections, which would make the extraction much more complicated. Fortunately, the RF packages modeled up to now rarely required more bandwidth than the model could offer.

Enhancements of the Π-section model

- Dealing with numerical singularities
  - Solved by using hybrid parameters *H*
    - Ports facing the IC treated as sources.
    - Ports facing the PCB treated as sinks.



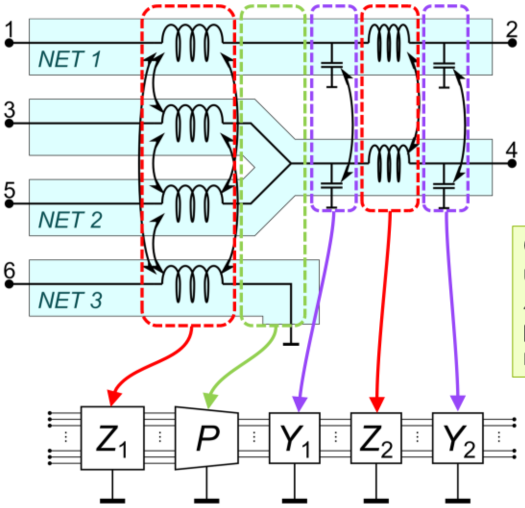
  - No singularities at the DC (in contrast to *Y*- and *Z*-parameters)
  - Good link to impedance/admittance models (in contrast to *S*-parameters)

The extraction routine relies on a fitting of the frequency response of the models to the target data. These target data come from the EM simulator and are originally expressed as the S-parameters. At this stage one need to choose the type of response. While Y- and S-parameters are often used to this end, they both have significant disadvantages. The Y-parameters become (almost) singular as some frequency points, including 0 Hz. The S-parameters are bound but distant from the physical interpretation of the model that is expressed in terms of voltages and currents. Therefore is convenient to describe the frequency response of the package to be modeled in terms of its hybrid parameters. The fact that the model has a clear cascaded topology linking two interfaces, makes the hybrid description particularly suitable.

The ports located at the IC-package interface will be treated as source ports with current forcing and voltage sensing. The ports located at the PCB-package interface will be treated as sink ports with voltage forcing and current sensing.

Description of the enhanced Π-section model

- Network
- Block diagram

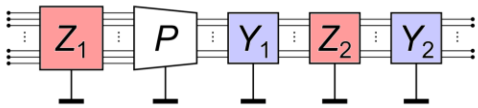


Owing to the choice of unknowns the Z_1, Y_1, Z_2, and Y_2 are always passive, causal, and reciprocal.

At this moment it is useful to introduce a compact block diagram representations of the model topology. The model includes four complex unknown matrices Z_1, Y_1, Z_2, and Y_2 as well as a fixed real matrix P. Matrix P represents the connectivity of the package and can be deduced from the S-parameters at 0 Hz, where Z_1, Y_1, Z_2, and Y_2 virtually vanish.

Description of the enhanced Π-section model

- Multiport hybrid parameters $\mathbf{H}\begin{bmatrix}\mathbf{i}_1\\ \mathbf{v}_2\end{bmatrix}=\begin{bmatrix}\mathbf{v}_1\\ \mathbf{i}_2\end{bmatrix}$
- Model equation



$$\mathbf{H}=\begin{bmatrix}\mathbf{H}_{11} & \mathbf{H}_{12}\\ \mathbf{H}_{21} & \mathbf{H}_{22}\end{bmatrix}=\begin{bmatrix}\mathbf{Z}_1+\mathbf{P}\mathbf{M}\mathbf{Z}_2\mathbf{P}^T & \mathbf{P}\mathbf{M}\\ -\mathbf{M}^T\mathbf{P}^T & \mathbf{Y}_1\mathbf{M}+\mathbf{Y}_2\end{bmatrix}$$

denominator of the transfer function

$$\mathbf{M}=\left(\mathbf{1}+\mathbf{Z}_2\mathbf{Y}_1\right)^{-1}$$

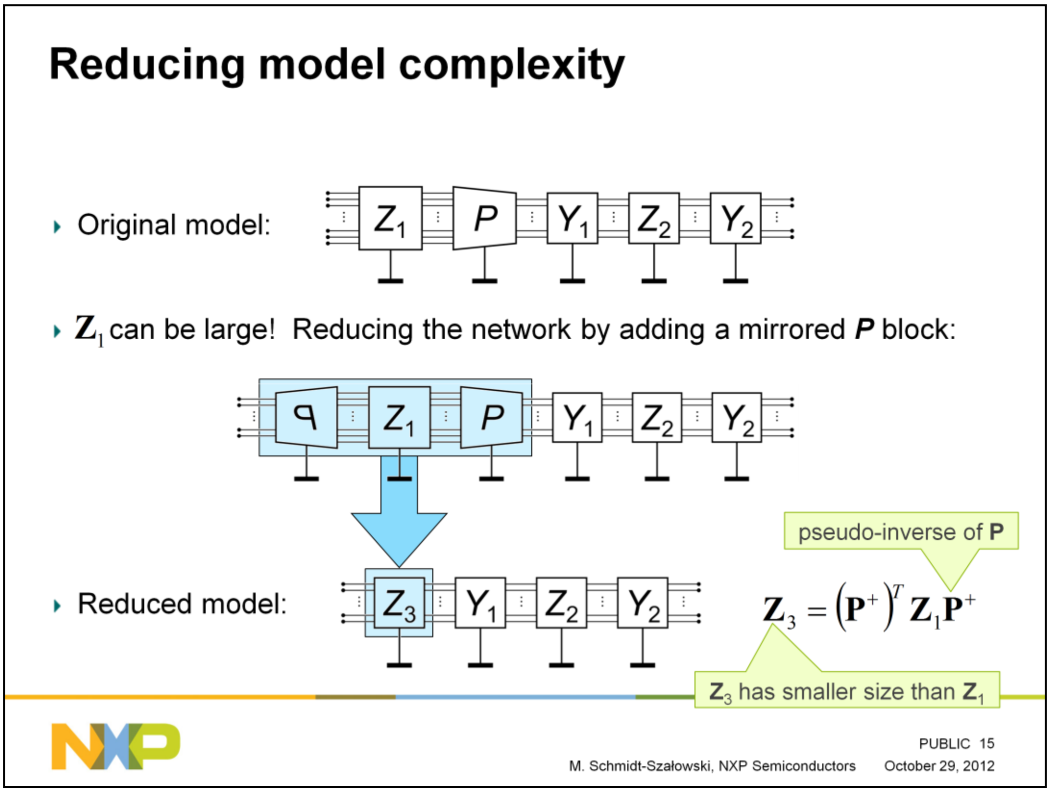
- Four complex matrices $\mathbf{Z}_1,\mathbf{Z}_2,\mathbf{Y}_1,\mathbf{Y}_2$

symmetry, passivity and causality guaranteed by construction

- One real matrix $\mathbf{P}$

rectangular matrix filled with 1's and 0's, extracted from the S-parameters at 0 Hz

The poles of the model are determined by matrix M. Owing to the normalization described in the paper, there will be no singularity at the frequencies in the vicinity of the poles.

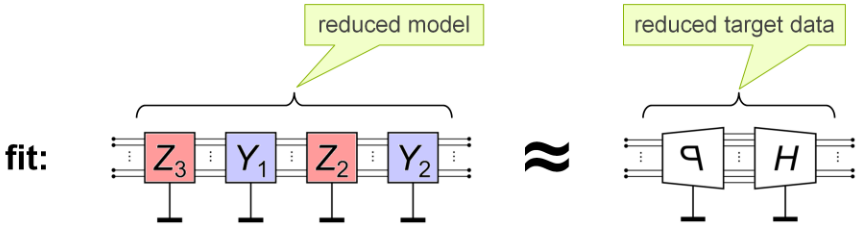


Note that the size of Z_1 is equal to the number of source ports while the size of Y_1, Z_2, and Y_2 is equal to the number of sink ports. When the number of source ports is significantly greater than the number of sink ports (typically when multiple downbonds have been applied) the size of Z_1 can become impractically large. The trick shown here helps to reduce the number of unknowns and the number of equations without any loss of accuracy. It consists in connecting a mirrored P-block to the Z_1-block (both in the model as in the target data) and replacing Z_1 together with both P-blocks attached to it with a new unknown matrix Z_3. This matrix is as large as Y_1, Z_2, and Y_2. Note that the poles of the system remain unchanged.

As a result, the solution process can be split into two smaller problems. First, Z_3, Y_1, Z_2, and Y_2 need to be found. Then Z_1 must be retrieved.

Solution process

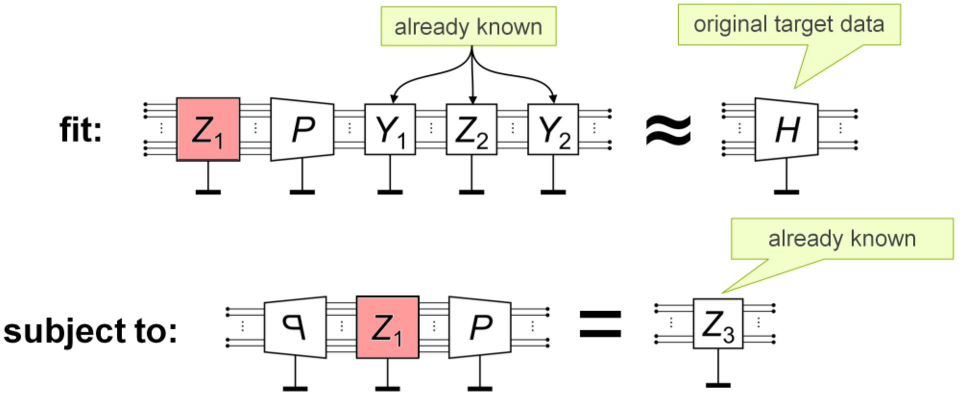
- Step 1:
  - Iteratively find the $\mathbf{Z}_3, \mathbf{Z}_2, \mathbf{Y}_1, \mathbf{Y}_2$ matrices of the model
  - Use the low-frequency approximation as the starting point
  - Quasi-Newton curve fitting in the domain of the *H*-parameters
    - all unknowns optimized at the same time
    - efficient even when the number of unknowns > 1000
    - typical computation time 10 sec to 10 min.



This is a classical least-squares fitting problem with several bounds on the unknowns. Among the other constraints, all the R, L, and C unknowns must stay positive. The model is non-linear in terms of the model parameters so the routine must perform several iterations before the solution is found. It is therefore essential to begin with a proper initial approximation. To this end, first a simplified model is extracted at low frequencies using a non-iterative method.

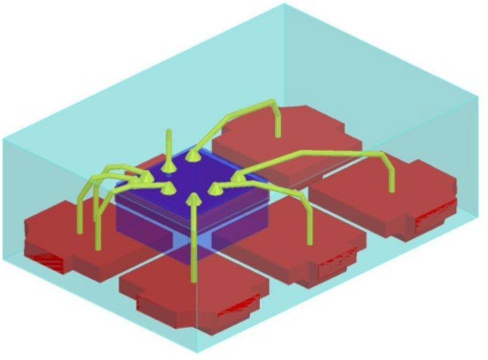
Solution process

- Step 2:
  - Retrieve $\mathbf{Z}_1$ as the solution of a constrained fitting problem



This is again a least-square problem with an equality constraint imposed on the unknowns. Owing to this constraint, the model fit obtained in the previous step will not be ruined.

Example 1: SOT886 package



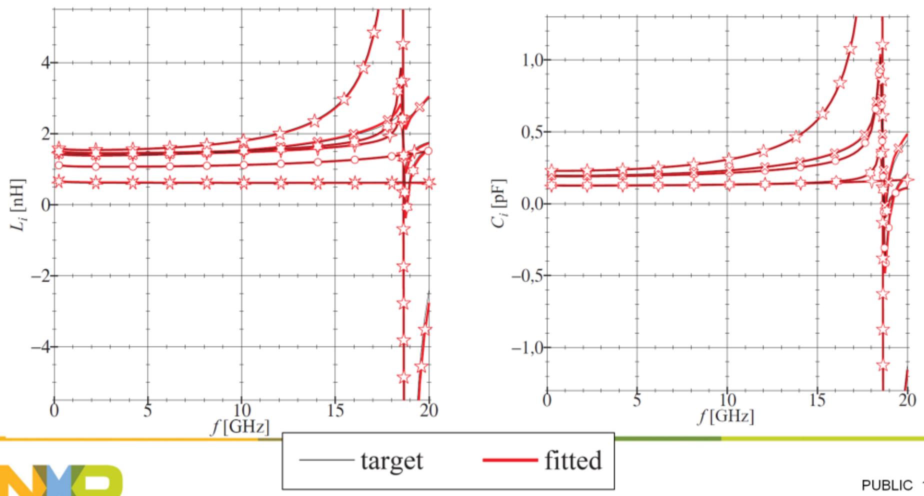
- Characteristics
  - 11 ports (= 6 sources + 5 sinks)
  - 80 unknowns
  - frequency range DC to 20 GHz
  - extraction time < 1 min

An example of a small RF package for a wideband LNA.

Example 1: SOT886 package

Inductance of the IO's
seen from the sources

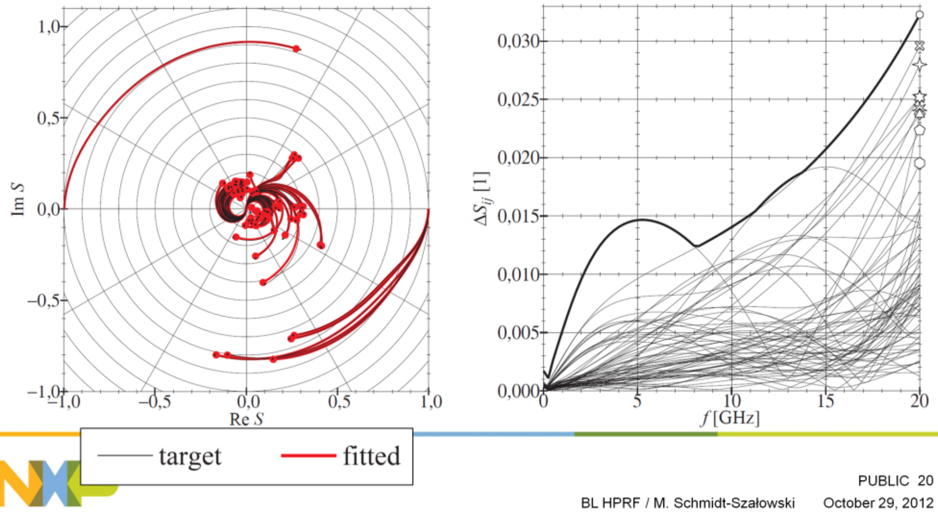
Capacitance of the IO's
seen by the sinks



Multiport S-parameters may contain so much information that it can be difficult to inspect them and interpret potential discrepancies. For this reason, it useful to project the model response in several different ways observing one feature at the time. These two figure show the diagonal elements of the inductance and capacitance matrix, that can be observed from the side of the sources and skinks, respectively. The fit is spot-on, also in the neighborhood of the pole.

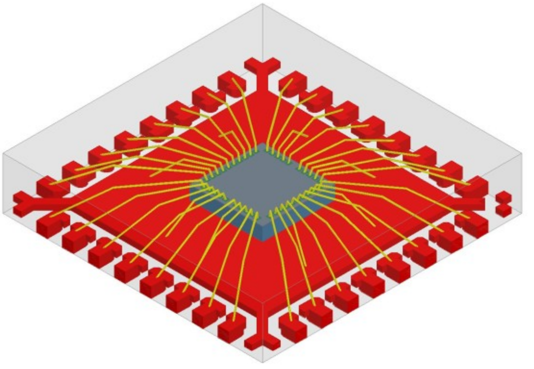
Example 1: SOT886 package

S-parameters | **Fitting errors in the S-parameters**



The model accuracy can be visualized also on the complex plane of the S-parameters.

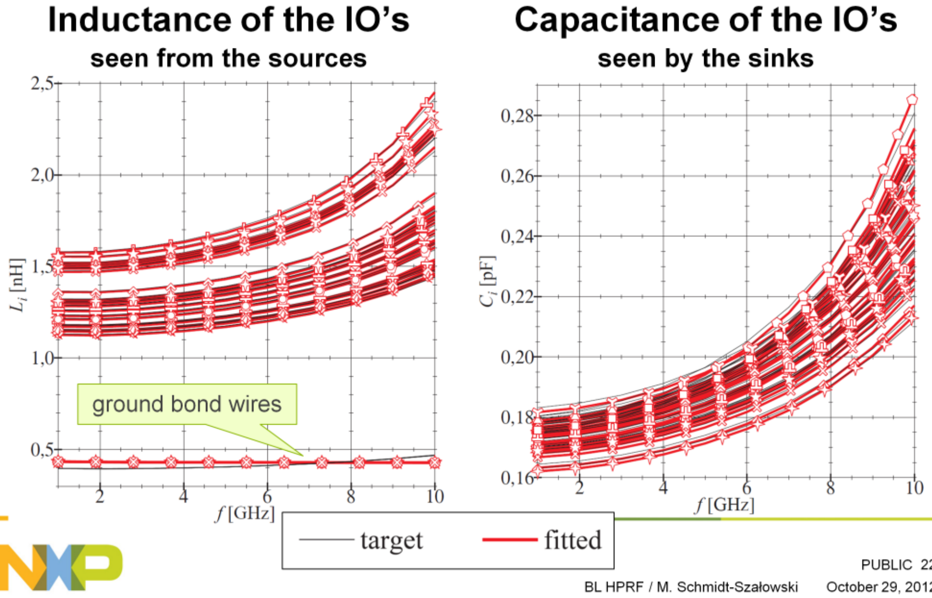
Example 2: HVQFN32 package



- Characteristics
  - 72 ports (= 40 sources + 32 sinks)
  - 1442 unknowns
  - frequency range DC to 10 GHz
  - extraction time ca. 120 min

This is a relatively large package for low-frequency applications.

Example 2: HVQFN32 package

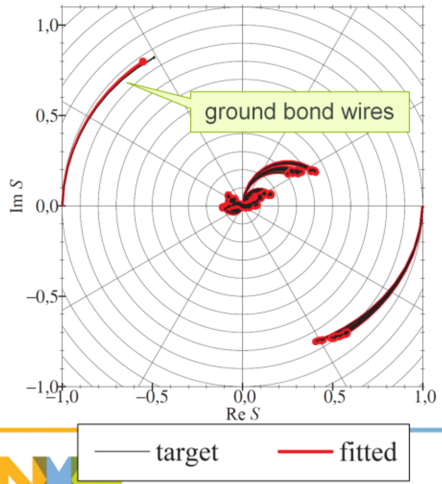


The package is far too large to incorporate all the possible mutual couplings. The number of unknown can be however significantly reduced using sparse matrix techniques. Omission of less relevant couplings leads to only insignificant loss of accuracy.

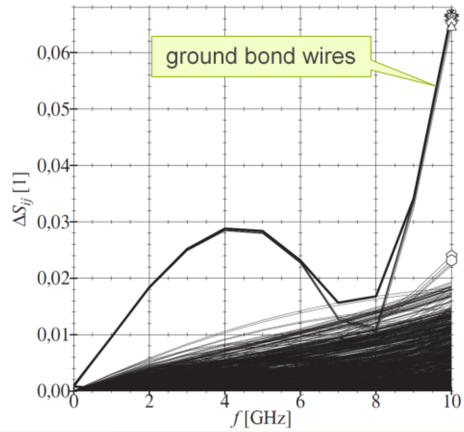
The biggest mismatch is caused by the relative long ground bond wires (downbons) whose capacitance to ground has not been modeled.

Example 2: HVQFN32 package

S-parameters



Fitting errors in the S-parameters



The over-all fit of the model is excellent, except for minor discrepancies related to the downbonds. But even then, the accuracy of the model is more then sufficient in the given application.

Summary

- Good trade-off between physical and mathematical modeling methods:
  - Owing to the built-in loss models low-order model is sufficient.
  - Passivity, causality, and reciprocity guaranteed by construction.
  - Fully automated.
- The generated netlist can be used in all popular types of analog circuit analysis:
  - linear and non-linear
  - DC, time domain, frequency domain
  - noise
- About 50 packages successfully extracted in last two years.
- Good candidate for a commercial EDA tool.

This modeling approaches addresses several issues that turned out to be a road block for the commercial modeling tools. The physical boundaries of the model enforce the required properties like, reciprocity, passivity, causality, leakage-free DC, ability to be extrapolated in frequency domain, and support for noise analysis.

At the same time the model comes with built-in loss models. The extraction routine does no need to extend the order of the model in order to capture the skin effect or dielectric relaxation.