An Integrated Overview of CAD/CAE Tools and Their Use on Nonlinear Network Design

José Carlos Pedro
and
Nuno Borges Carvalho

Telecommunications Institute – University of Aveiro

Summary

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions
**Objectives**

1- Present an overview of the various CAD tools for RF/microwave nonlinear network design.

2 - Integrate some fields held in nonlinear circuit/system design, as:
   - nonlinear circuit and device modeling strategies,
   - nonlinear circuit/system analysis methods valid for RF design
   - circuit performance evaluation and design validation

**Summary**

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions
No closed-form solutions are possible!
CAD tools, are necessary. And so are accurate mathematical modeling of devices, circuits and signals.

Signals which are Envelope Modulated RF Carriers:

Aperiodic of Continuous Spectra and involving 2 Time-scales $\Rightarrow \nabla \ ?$

An approximate signal model is required!
Approximate Models for Envelope Modulated RF Carriers

Sinusoidal: $V(\omega_0)$

Two-Tone: $V_1(\omega_1) \quad V_2(\omega_2)$

Multi-Tone: $V(\omega)$

Time-Varying RF Carrier: $\sum_k V_k(\tau_1)e^{j\omega_0\tau_2}$

... or simply represent and simulate only the envelope! ...
Summary

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions

Methods for Nonlinear RF Circuit and System Simulation - 1

Nonlinear RF Network Modeling

Take, for example, this Nonlinear Network:

\[ i_S(t) \quad \begin{array}{c} \text{G} \\ \text{v}_O(t) \end{array} \quad q_{NL}[v_O(t)] \quad i_{NL}[v_O(t)] \]

It can be represented by the following ordinary differential nodal equation:

\[ G v_O(t) + \frac{d q_{NL}[v_O(t)]}{dt} + i_{NL}[v_O(t)] = i_S(t) \]
Time-Step Integration

\[ G v_O(t) + \frac{d q_{NL}[v_O(t)]}{dt} + i_{NL}[v_O(t)] = i_S(t) \]

This ODE is solved for \( v_O(t) \) transforming it into a difference equation, discretizing time in various instants \( t_k \) separated by dynamic steps \( h_k \):

\[ G v_O(t_k) + q_{NL} \left[ \frac{v_O(t_k)}{h_k} - q_{NL}[v_O(t_{k-1})] \right] + i_{NL}[v_O(t_k)] = i_S(t_k) \]

or

\[ h_k G v_O(t_k) + q_{NL}[v_O(t_k)] + h_k i_{NL}[v_O(t_k)] = h_k i_S(t_k) + q_{NL}[v_O(t_{k-1})] \]

which is then solved in a time-sep by time-step basis, for all \( v_O(t_k) \), beginning with the initial state \( v_O(t_0) \) until final time \( v_O(t_K) \).

---

Since this time-step integration scheme was conceived for transient response calculations it is:

1 - Inefficient: Steady-State response requires waiting until all transients died.

2 - Inadequate: Works on time-domain while most signal and circuit models, are represented in frequency-domain.

3 - Inaccurate: Impossibility of waiting for complete vanishing of long transients, as use of the DFT between domains introduce large errors.

Nevertheless, time-step integration is still one of the mostly used methods of nonlinear circuit and system simulation !!

It is the core method of all SPICE like or Simulink programs.
Methods for Nonlinear RF Circuit and System Simulation - 4

Time-Domain Calculation of Steady-State Responses
(Shooting – Newton)

To bypass the transient computation, the initial condition, \( v_O(t_0) \), must be selected such that after the excitation period \( T \) the same initial state is obtained:

\[
v_O(t_0 + T) = v_O(t_0) \quad \text{for} \quad i_S(t_0 + T) = i_S(t_0)
\]

The idea is to estimate the sensitivity of the final state, \( v_O(t_0 + T) \), to variations in the initial condition, \( v_O(t_0) \):

\[
S[v_O(t_0)] = \frac{\partial v_O(t_0 + T)}{\partial v_O(t_0)} \bigg|_{v_O(t_0)} = \frac{\Delta v_O(t_0 + T)}{\Delta v_O(t_0)}
\]

and then use this sensitivity to propose an educated guess for the correct initial condition \( v_O(t_0) \).

Methods for Nonlinear RF Circuit and System Simulation - 5

Time-Domain Calculation of Steady-State Responses
(Shooting – Newton)

In the Shooting-Newton algorithm, this sensitivity, \( S[v_O(t_0)] \), is used to iteratively solve the boundary-value nonlinear equation:

\[
v_O(t_0 + T) = v_O(t_0) \quad \text{or} \quad v_O(t_0 + T) - v_O(t_0) = 0
\]

determining the new initial condition estimate, \( i+1v_O(t_0) \), from \( i'v_O(t_0) \), and the result of a time-step integration in a period, \( i'v_O(t_0 + T) \), by:

\[
i+1v_O(t_0) = i'v_O(t_0) - \left[ S[v_O(t_0)] \right]^{-1} \cdot \left[ i'v_O(t_0 + T) - i'v_O(t_0) \right]
\]

Shooting-Newton is an interesting alternative for RF steady-state simulation as \( S[v_O(t_0)] \) can be obtained along with time-step integration without any post-processing, and because it seems that the boundary-value equation \( v_O(t_0 + T) - v_O(t_0) = 0 \) is mildly nonlinear, although the circuit may be strongly nonlinear.
Methods for Nonlinear RF Circuit and System Simulation - 6

Frequency-Domain Methods

The RF approach for analyzing our circuit is to treat it in frequency-domain:

\[ i_s(t) \quad \frac{d}{dt} \quad v_o(t) \quad q_{NL}(v_o(t)) \quad i_{NL}(v_o(t)) \]

Both the excitation and the state-variables are represented as truncated Fourier series:

\[ i_s(t) = \sum_{k=-K}^{K} i_k e^{j k \omega_0 t} \quad \text{and} \quad v_o(t) = \sum_{k=-K}^{K} v_k e^{j k \omega_0 t} \]

which, substituted in the circuit’s time-domain ODE

\[ G v_o(t) + \frac{d}{dt} q_{NL}[v_o(t)] + i_{NL}[v_o(t)] = i_s(t) \]

gives:

\[ \sum_k G v_k e^{j k \omega_0 t} + \frac{d}{dt} \left[ q_{NL}\left( \sum_k v_k e^{j k \omega_0 t} \right) \right] + i_{NL}\left( \sum_k v_k e^{j k \omega_0 t} \right) = \sum_k i_k e^{j k \omega_0 t} \]

Because terms of Fourier series are orthogonal, this actually corresponds to a nonlinear system of \((2K+1)\) equations, one for each harmonic component \(k\).

In matrix form, that system is known as the Harmonic-Balance Equation.

\[ G V_o + j \Omega Q_{nl}(V_o) + I_{nl}(V_o) = I_s \]
**Methods for Nonlinear RF Circuit and System Simulation - 8**

**Frequency-Domain Methods**

\[ GV_o + j\Omega Q_{nl}(V_o) + I_{nl}(V_o) = I_s \]

There are two alternative ways of solving this HB equation for \( V_o \):

In Volterra Series it is assumed that the problem is only mildly nonlinear, and so that its solution can be approximated by the analytical solution of a similar problem in which the nonlinearities are approximated by low order Taylor series expansions.

In Harmonic-Newton, the full nonlinear HB equation is iteratively solved using a multi-dimensional Newton-Raphson algorithm.

---

**Methods for Nonlinear RF Circuit and System Simulation - 9**

**Volterra Series**

Approximating \( q_{nl}[v_o(t)] \) and \( i_{nl}[v_o(t)] \) by low order Taylor series expansions around the quiescent point, \((I_o, V_o)\), the nonlinear charge and current signal components can be given by:

\[ q_{nl}(v_o) = c_1 v_o(t) + c_2 v_o(t)^2 + c_3 v_o(t)^3 \]

and

\[ i_{nl}(v_o) = g_1 v_o(t) + g_2 v_o(t)^2 + g_3 v_o(t)^3 \]

which leads to a circuit solution \( v_o(t) \) of the form:

\[ v_o(t) = \sum_{n=1}^{\infty} \frac{1}{n!} \sum_{Q} I_{S_{\omega_1}} \cdots I_{S_{\omega_n}} H_{n}(\omega_{\omega_1}, \cdots, \omega_{\omega_n}) e^{j(\omega_{\omega_1} + \cdots + \omega_{\omega_n})t} \]

In Volterra Series the system becomes completely identified by its Nonlinear Transfer Functions.

Volterra series can thus be applied to either Circuit or System Analysis and Design!
Volterra Series

\[ v_o(t) = H_1(\omega) i(t) + H_2(\omega_1, \omega) v_o(t) + H_3(\omega_1, \omega_2, \omega_3) v_o(t) + \cdots \]

Harmonic-Newton

Alternatively, the full nonlinear harmonic-balance equation

\[ G V_o + j\Omega Q_{nl}(V_o) + I_{il}(V_o) = I_s \]

can be solved for \( V_o \) using a \((2K+1)\)-dimensional Newton-Raphson algorithm.

When the HB equation is put in the form:

\[ F(V_o) = G V_o + j\Omega Q_{nl}(V_o) + I_{il}(V_o) - I_s = 0 \]

the solver creates a succession of solution estimates \( 0 V_o, 1 V_o, \ldots, V_o, i+1 V_o, \ldots, f V_o \)

\[ \text{until } \left| F(V_o) \right| < \varepsilon \]

\[ i+1 V_o = V_o - \left[ \frac{dF(V_o)}{dV_o} \right]^{-1} F(V_o) \]

\[ \text{until } \left| F(V_o) \right| < \varepsilon \]
Harmonic-Newton

The Harmonic-Newton is the most used method for microwave circuit simulation.

In the Piecewise Harmonic-Newton implementation, the network

\[ i_s(t) \quad G \quad v_0(t) \quad q_{nl}[v_0(t)] \quad i_{NL}[v_0(t)] \]

Methods for Nonlinear RF Circuit and System Simulation - 12

Harmonic-Newton

The Harmonic-Newton is the most used method for microwave circuit simulation.

In the Piecewise Harmonic-Newton implementation, the network is divided into two sub-circuits, one nonlinear, memoryless, and another linear, dynamic, which, with the excitation, allow the construction of a nodal form of the HB equation:

\[
F(V_\omega) = I_{cl}(V_\omega) + I_{cnl}(V_\omega) - I_s(\omega) = 0
\]

where

\[ I_{cl}(V_\omega) = Y_{cl}(\omega)V_\omega(\omega) \quad \text{and} \quad I_{cnl}(V_\omega) = DFT^{\dagger}_{CNL}[DFT^{-1}[V_\omega(\omega)]] \]
Multi-Rate Techniques

When the excitation is an RF carrier $\cos(\omega_0 t)$ modulated by a base-band envelope $v_e(t)$, which is uncorrelated with that carrier, the circuit behaves as if it had a stimulus dependent on two different time-scales $\tau_1$ and $\tau_2$:

$$i_S(\tau_1, \tau_2) = V_e(\tau_1) \cos(\omega_0 \tau_2)$$

The circuit’s ODE, in $t$:

$$G v_O(t) + \frac{\partial q_{NL}[v_O(t)]}{\partial t} + i_{NL}[v_O(t)] = i_S(t)$$

becomes a Multi-Rate Partial Differential Equation, MPDE, in $(\tau_1, \tau_2)$:

$$G v_O(\tau_1, \tau_2) + \frac{\partial q_{NL}[v_O(\tau_1, \tau_2)]}{\partial \tau_1} + \frac{\partial q_{NL}[v_O(\tau_1, \tau_2)]}{\partial \tau_2} + i_{NL}[v_O(\tau_1, \tau_2)] = i_S(\tau_1, \tau_2)$$

which again can be solved in a bi-dimensional time-domain for $v_O(\tau_1, \tau_2)$, or bi-dimensional frequency-domain for $V_o(k_1 \Omega_0, k_2 \omega_0)$.

If both the envelope and the carrier are periodic, then it is probably better to solve the MPDE in frequency-domain. In this case, the state variable would become:

$$v_O(\tau_1, \tau_2) = \sum_{k_1} \sum_{k_2} V_{k_1, k_2} e^{j k_1 \Omega_0 \tau_1} e^{j k_2 \omega_0 \tau_2}$$

which substituted in the MPDE:

$$G v_O(\tau_1, \tau_2) + \frac{\partial q_{NL}[v_O(\tau_1, \tau_2)]}{\partial \tau_1} + \frac{\partial q_{NL}[v_O(\tau_1, \tau_2)]}{\partial \tau_2} + i_{NL}[v_O(\tau_1, \tau_2)] = i_S(\tau_1, \tau_2)$$

would lead to the following bi-dimensional HB equation, the basis for most of the multi-tone nonlinear simulation methods.

$$G V_o(\Omega, \omega) + I_{nl}[V_o(\Omega, \omega)] + j \Omega Q_{nl}[V_o(\Omega, \omega)] + I_f(\Omega, \omega) = 0$$
In most practical cases, however, the envelope is aperiodic, and so it is better to solve the MPDE in the frequency-domain, \( \omega \), for the carrier, but in the time-domain, \( \tau \), for the envelope. In this case, the state variable would become:

\[
y_{o}(\tau_1, \tau_2) = \sum_{k_2} V_{k_2}(\tau_1) e^{j k_2 \omega_0 \tau_2}
\]

which substituted in the MPDE would lead to the following \( \tau \) time-varying HB equation:

\[
G V_o(\tau_1) + \frac{\partial Q_{nl}[V_o(\tau_1)]}{\partial \tau_1} + j \Omega Q_{nl}[V_o(\tau_1)] + I_{nl}[V_o(\tau_1)] - I_s(\tau_1) = 0
\]

This time-varying HB equation is now discretized in \( \tau \)-time

\[
h_k G V_o(\tau_{1k}) + \frac{Q_{nl}[V_o(\tau_{1k})] - Q_{nl}[V_o(\tau_{1k-1})]}{h_k} + j h_k \Omega Q_{nl}[V_o(\tau_{1k})] + I_{nl}[V_o(\tau_{1k})] - I_s(\tau_{1k}) = 0
\]

which allows the determination of the envelope transient solution for each of the \( V_{o\ell}(\tau) \) harmonics, solving:

\[
h_k G V_o(\tau_{1k}) + Q_{nl}[V_o(\tau_{1k})] + j h_k \Omega Q_{nl}[V_o(\tau_{1k})] + h_k I_{nl}[V_o(\tau_{1k})] =
\]

\[
= h_k I_s(\tau_{1k}) + Q_{nl}[V_o(\tau_{1k-1})]
\]

This method, known as Envelope Transient Harmonic-Balance, constitutes a serious step towards a true nonlinear system envelope simulator.
In summary, this envelope transient harmonic-balance can be described by:

1º - Represent the excitation as an RF carrier modulated by a base-band envelope:

\[ i_S(t) = i_{Sc}(\tau_2) \]

\[ i_S(\omega_0) \]

or, in-frequency:

\[ i_S(\Omega) \]

\[ i_S(\Omega_0) \]

2º - Discretize the base-band envelope at instants \( \tau_{1k} \), create the excitation at those instants \( I_s(\tau_{1k}) \), and calculate the solution, \( V_o(\tau_{1k}) \), by an Harmonic-Newton solver.

3º - Interpolate the obtained \( V_o(\tau_{i}) \) variations.
4º - Finally, reconstruct the solution in frequency, \( V_o(\omega) \), and time-domains \( v_o(t) \).

**Summary**

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions
Correct design of nonlinear networks demands for a precise description of the whole circuit, which includes the nonlinear active device, but also matching and bias networks.

Frequency modeling of the matching circuits is usually undertaken for the fundamental and, when the goal is efficiency or distortion, for the harmonics.

Bias networks are usually neglected. Typical microwave/millimeter wave network analyzers are only capable of measuring S parameters above 40MHz, while envelopes span up to a few hundreds of KHz or MHz, at most!
A simple Volterra model shows the impact of all in-band and out-of-band terminating impedances.

\[ H_1(\omega) = \frac{-G_{m_1}}{1 + G_m Z_1(\omega)} \]  
\[ H_2(-\omega_1, \omega_2) = \frac{-1}{2 + 2G_m Z_2(-\omega_1 + \omega_2)} \left[ 2G_m + G_m Z(-\omega_1)H_1(-\omega_1) + G_m Z(\omega_1)H_1(\omega_1) + 2G_m, H_1(-\omega_2)Z(-\omega_2)H_1(\omega_2)Z(\omega_2) \right] \]

Base-band, and/or, bias networks are usually ignored by the designer, although they can be of paramount importance in nonlinear network design. They control signal’s envelope dynamics, e.g. manifested as IMD asymmetries!
Base-band, and/or, bias networks are usually ignored by the designer, although they can be of paramount importance in nonlinear network design. They control signal’s envelope dynamics, e.g. manifested as IMD asymmetries!
Nonlinear Device and Circuit Modeling - 4

Base-band, and/or, bias networks are usually ignored by the designer, although they can be of paramount importance in nonlinear network design. They control signal’s envelope dynamics, e.g. manifested as IMD asymmetries!

Nonlinear Device and Circuit Modeling - 5

Real telecommunication signals involve envelopes of dense spectrum. Bias networks must be accurately represented from DC to Envelope Bw.

All this bandwidth is accounted for.
Nonlinear Device and Circuit Modeling - 6

Nonlinear Device Model Formulation

Any transistor, which is a memoryless nonlinearity embedded in a linear dynamic network becomes:

- a complex dynamic nonlinear device if tested on its terminals,

![Dynamic Nonlinear Two-Port Diagram]

Nonlinear Device and Circuit Modeling - 7

Nonlinear Device Model Formulation

Any transistor, which is a memoryless nonlinearity embedded in a linear dynamic network becomes:

- a complex dynamic nonlinear device if tested on its terminals,

- or a memoryless nonlinearity if deembedded from its linear dynamic elements.

![Dynamic Linear Four-Port Diagram]
Nonlinear Device and Circuit Modeling - 8

Nonlinear Device Model Formulation

For illustrative purposes, we will now discuss a Nonlinear MESFET Model obtained using the above procedure.

First step is to adopt a certain equivalent circuit topology, identifying there the nonlinear elements and the embedding dynamic network.

Nonlinear Device Model Formulation

In the MESFET case, we used one-dimensional physics simulations of the device to get detailed information on the $i_{DS}(v_{GS}, v_{DS})$ and $C_{gs}(v_{GS})$ functions.

The adopted functions are valid in a large range of $v_{GS}$ and $v_{DS}$, are continuous and present continuous derivatives, and approximate reasonably well the measurements at least up to the 3rd derivative.

\[ i_{DS}(v_{GS}, v_{DS}) = \beta \mu + \ln\left(e^\mu + e^{-\mu}\right) \tanh(\alpha v_{DS}) \]

They are thus consistent with measured (0 order) DC, (1st order) S-par, and (2nd and 3rd order) harmonic distortion.
**Nonlinear Device and Circuit Modeling - 10**

**Nonlinear Device Model Extraction**

The linear equivalent circuit was extracted from small-signal frequency swept S-parameter data, using the direct method of Dambrine et al, and then adjusted by careful linear optimization.

\[ i_{DS}(v_{GS}, v_{DS}) \text{ nonlinear model parameters were extracted by comparing measured } i_{DS}, \text{ and all its first 9 derivatives up to order 3 in respect to } v_{GS} \text{ and } v_{DS}. \]

\[ i_{DS}(v_{GS}, v_{DS}) = I_{DS} + Gm v_{gs} + Gd v_{ds} + Gm2 v_{gs}^2 + Gmd v_{gs} v_{ds} + Gd2 v_{ds}^2 + Gm3 v_{gs}^3 + Gm2d v_{gs}^2 v_{ds} + Gmd2 v_{gs} v_{ds}^2 + Gd3 v_{ds}^3 \]

with the ones predicted by the adopted functional description.

---

**Nonlinear Device and Circuit Modeling - 11**

The derivatives’ extraction procedure is based on intermodulation measurements when the device is excited at its input and output by one tone at \( \omega_1 \), and another one at \( \omega_2 \).
Nonlinear Device and Circuit Modeling - 12

As seen in the figures, $i_{DS}(V_{GS})$ derivatives predicted by the model are in very good agreement with correspondent measured data.

Measured and modeled drain source current, $i_{DS}$ and its derivatives, $G_m$, $G_{m2}$ and $G_{m3}$

Nonlinear Device and Circuit Modeling - 13

The same set-up can also be used for extracting $C_{GS}(V_{GS})$ coefficients. But now, higher frequencies must be applied.
$C_{gs}(v_{GS})$ and its derivatives predicted by the model are again in a reasonable good agreement with correspondent measured data.

Measured and modeled gate source capacitance, and its derivatives $C_{gs}$, $C_{g2}$, and $C_{g3}$

An alternative way to this device and circuit level modeling is System or Black-Box Behavioral Modeling

Conceived for the efficient simulation of large telecommunication systems, it tries to find input/output transfer functions that accurately describe the observed responses of the system.

At the present time it seems we can model “exactly” mildly nonlinear dynamic systems by Volterra series,

and we could also model strong nonlinear memoryless systems,

but we can not yet represent strong nonlinear dynamic systems.

Presently an important effort is being paid to solve this challenging problem!
Summary

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions

Design Validation - 1

The simplest approach to nonlinear network design validation is testing under a sinusoidal excitation:

\[ y(t) = A \cos(\omega t) \]

the output is nothing but the sum of a DC component, the fundamental and its harmonics:
Only gain compression (AM/AM) and phase shift (AM/PM) versus input power, can be verified.

Exclusively accounts for short memory effects at the fundamental frequency.

Can not account for the envelope or signal bandwidth behavior!

Testing for simple envelope effects requires, at least, a two-tone excitation:

\[
x(t) = A_1 \cos(\omega_1 t) + A_2 \cos(\omega_2 t) = 2A_1 \cos\left(\frac{\omega_2 - \omega_1}{2} t\right)\cos\left(\frac{\omega_2 + \omega_1}{2} t\right)
\]

\[
y(t) = a_2 A_2^2 + a_2 A_2^2 \cos[\omega_2 - \omega_1, 0]
+ \left[ a_2 A_2^2 + \frac{9}{4} a_2 A_2^2 \right] \sin(\omega_1 t) + \left[ a_2 A_2^2 + \frac{9}{4} a_2 A_2^2 \right] \sin(\omega_2 t)
+ \frac{a_2 A_2^2}{2} \cos(2\omega_1 t) - a_2 A_2^2 \cos(2\omega_2 t) - \frac{a_2 A_2^2}{2} \cos(2\omega_2 t)
+ \frac{3}{4} a_2 A_2^4 \sin[(2\omega_1 - \omega_2) t] - \frac{3}{4} a_2 A_2^4 \sin[(2\omega_1 + \omega_2) t]
+ \frac{3}{4} a_2 A_2^4 \sin[(2\omega_1 - \omega_2) t] - \frac{3}{4} a_2 A_2^4 \sin[(2\omega_1 + \omega_2) t] - \frac{a_2 A_2^4}{4} \sin(3\omega_1 t)
- \frac{a_2 A_2^4}{4} \sin(3\omega_1 t)
\]
Since the envelope of a two-tone signal is a pure sinusoid, these tests are actually sensing the nonlinear network’s long-term memory effects only at specific frequency spots.
Special care should be paid to the phase arrangements of multi-tone signals as they define envelope peak-power, and thus strongly affect nonlinear response.

Time waveform representation of two signals composed of 10 evenly spaced and equal amplitude tones, but with different phase arrangements.

Best results are obtained using the signals which the system is intended to handle. But, these signals are sometimes substituted by RF band-limited gaussian noise.
To illustrate those design validation ideas let us compare simulation and measured results obtained with a practical MESFET microwave amplifier.

This amplifier was modeled as previously explained, and then tested under Two-Tone excitation. A perfect match between HB simulations and measurements is evident.
Finally, the amplifier was tested for a Band-Limited White Noise excitation. The stimulus’ power spectral density was:

![Power spectral density function of the noise stimulus.](image1)

Measured and HB simulated in-band output results of ACPR and CCPR are depicted in the figure. Once again, a very good agreement was obtained between both data types.

![Power spectral density function of the amplifier’s in-band output.](image2)

This proves the enormous benefits of the present CAD tools in nonlinear microwave network design.
**Summary**

1 – Objectives

2 – The Nonlinear RF Network CAD Problem

3 – Methods for Nonlinear RF Circuit and System Simulation

4 – Nonlinear Device and Circuit Modeling

5 – Design Validation

6 – Conclusions

**Conclusions - 1**

1 – Although nonlinear network design is a problem of complex nature, it is rapidly advancing towards a mature state.

2 – Because both short-term and long-term memory effects are involved, simulating nonlinear networks subject to telecommunication signals requires efficient mixed-mode multi-rate techniques.
Conclusions - 2

3 – But, it also requires specific electron device and terminating networks modeling, valid, not only near the excitation frequency, as at the base-band and the first few harmonics.

4 – Because the response of a nonlinear network to a certain input can not be determined from the response to any other excitation, it must be tested with signals as close as possible to the expected real ones. However, it seems that reasonable design analysis and validation can already be obtained from band-limited noise approximations.