A Frequency-Domain, Volterra Series-Based Behavioral Simulation Tool for RF Systems

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Abstract

In this paper a new behavioral modeling approach for RF systems based is presented, based on a Volterra series input-output map representation. The modeling is done purely in the frequency domain, capturing the typical system level specifications for RF building blocks, independent of the implementation details. A harmonic balance simulation tool has been developed based on those models. The implementation focuses on deterministic effects such as distortion and frequency conversion. The behavioral simulator has been tested for various systems and results are presented.

1 Introduction

To accelerate the design of modern wireless communication systems it is crucial to perform quick explorations of system-level architectures using behavioral simulation and modeling. While the analog/RF part of the system typically presents the biggest challenges and is the bottleneck of the design process, designers typically rely on ad-hoc techniques and simple analytical models to perform high-level estimation. To accelerate the design of analog and mixed-signal systems, a top-down, constraint-driven design methodology for analog circuits has been proposed [1]. To expand the applicability of the methodology to RF, efficient behavioral simulation techniques are needed.

In RF systems, the typical figure of merit is the total signal-to-noise-and-distortion ratio (SNDR). Typically specifications are given in the frequency domain as the desired SNDR in the presence of certain environmental conditions, such as out-of-band signals or in-band, co-channel interference and intermodulation. Typical performance metrics include gain or attenuation as a function of frequency, second and third order distortion and noise figure (NF).

Once block-level characterization is obtained, either from specifications or from transistor-level simulations, analytical approximations [2] can be used to calculate the overall SNDR. Those approximations can be inaccurate when applied to systems with feedback or coupling, do not take into account phase and it is generally up to the designer to model correctly all the possible interactions between non-idealities. A more systematic and accurate approach could be to use macromodels in time-domain simulators. However, such models tend to add more implementation details than necessary and then again abstract only the necessary frequency domain performance. For example, the desired frequency-domain behavior of a filter has to be implemented with a specific pole-zero representation, that can result in very long simulation times. Frequency-dependent nonlinearities are also hard to model using this approach. Available harmonic-balance simulators also rely on time-domain models of blocks.

In this paper we present a new technique to model and simulate deterministic effects of RF systems, based on a purely frequency-domain representation. It uses Volterra series-based models and harmonic balance simulation. Unlike previous approaches [4], Volterra-based modeling is done only at the behavioral level using a look-up table approach. Harmonic balance is used to solve arbitrary interconnects of blocks. Using our models, the Jacobian of the system can be swiftly evaluated and solved since there is no dependancy on time-domain differential equation-based models.

2 Modeling of Deterministic Effects Using Volterra Series

In considering systems with weak nonlinearities only, one abstract approach that only uses high-level frequency-domain description, is to use the Volterra theory of nonlinear systems [3, 4]. Weakly nonlinear blocks can be described as N-ports with a Volterra input-output frequency domain mapping function [5]. In addition, RF system specifications are given with the assumption of sinusoidal inputs to approximate the modulated signal, thus the input can be considered to be a sum of complex exponential functions.

Under this assumption, the response of an N-port at port n is:

$$y_n = \sum_{m=1}^{M} y_n^{(m)} \tag{1}$$

where M is the order up to which nonlinearities are important. $y_n^{(m)}$ is the total response at port n of frequency mixes of m-th order given by:

$$y_n^{(m)} = \sum_{n_1,\ldots,n_m,k_1,\ldots,k_m} H_{nn_1\ldots n_m}^{(m)}(f_{k\,1},\ldots,f_{k\,m}) \cdot \\ x_{n\,1}(f_{k\,1})\ldots x_{n\,m}(f_{k\,m}) \cdot \\ e^{\jmath 2\pi[\sum_{i=1}^{m} f_{k\,i}]t} \qquad (2)$$

where $H_{nn_1,\ldots,n_m}^{(m)}(f_{k\,1},\ldots,f_{k\,m})$ is the value of the m-th order transfer function from ports $n_1,\ldots,n_m$ to port n for input frequencies $f_{k\,1},\ldots,f_{k\,m}$ and $x_{n\,m}(f_{k\,m})$ is the coefficient of the input harmonic $f_{k\,m}$ at port $x_{n\,m}$

This representation cannot model strong nonlinear behavior, for example the behavior of a mixer. However, excluding strong compression, in a typical RF receiver the signals driving the strong nonlinearities are periodical. Thus, the *periodically time-varying* transfer function of a mixer, $H(t,f)$, can be expanded to a Fourier series [6]:

$$H(t,f) = \sum_{l=-\infty}^{\infty} H^l(f) \cdot e^{\jmath 2\pi l f_0} \qquad (3)$$

where f_0 is the frequency of the signal driving the nonlinearity and $H^l(f)$ is the l-th *sideband* transfer function of the *linear, time-varying system.*

While a mixer is a strongly nonlinear module, the input-output behavior from RF to IF is typically *weakly nonlinear* and thus can be represented with a Volterra input-output map as in (2). So if each coefficient of (3) is expanded to a Volterra series, both periodical strong nonlinearities and weak nonlinearities can be modeled. Similarly to (3), (2) becomes:

$$y_n^{(m)} = \sum_{l,n_1,\ldots,n_m,k_1,\ldots,k_m} H_{nn_1\ldots n_m}^{(m),l}(f_{k\,1},\ldots,f_{k\,m}) \cdot \\ x_{n\,1}(f_{k\,1})\ldots x_{n\,m}(f_{k\,m}) \cdot \\ e^{\jmath 2\pi[\sum_{i=1}^{m} f_{k\,i}+lf_0]t} \qquad (4)$$

By using this representation, most RF building blocks can be modeled in the frequency domain. For example, the model for a mixer taking into account third order distortion, reverse leakage gain, offset, conversion gain, RF leakage and 2nd spurious response could be the following (at a specific frequency f_j):

$$\begin{aligned} 0 &= -I_{in}(f_j) + A_{i_r}(f_j)I_{out}(f_j) + \frac{V_{in}(f_j)}{Z_{in}(f_j)} \\ V_{os}(f_j) &= -V_{out}(f_j) + Z_{out}(f_j)I_{out}(f_j) + {A_v}^{(1),0}(f_j)V_{in}(f_j) + \\ &\sum_{i:f_i \pm f_{LO}=f_j} {A_v}^{(1),\pm 1}(f_i)V_{in}(f_i) + \sum_{i:f_i \pm 2f_{LO}=f_j} {A_v}^{(1),\pm 2}V_{in}(f_i) + \\ &\sum_{i_1,i_2,i_3:f_{i_1}+f_{i_2}+f_{i_3}\pm f_{LO}=f_j} {A_v}^{(3),\pm 1}(f_{i_1},f_{i_2},f_{i_3})V_{in}(f_{i_1})V_{in}(f_{i_2})V_{in}(f_{i_3}) \end{aligned} \qquad (5)$$

where ${A_v}^{(m),l}$ is the sideband l, order m, of the voltage gain. Figure 1 depicts the model. Instead of currents and voltages also s-parameters can be used as in [5].

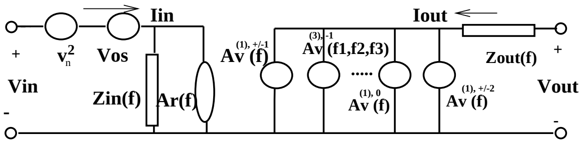

Figure 1: Mixer Behavioral Model

This representation is compatible with the frequency domain *cyclostationary* noise analysis [7]. If frequency domain noise spectral density functions are used to model the noise performance of the RF blocks, the sideband transfer functions defined in 3 can be used to compute the overall system noise performance. Thus we can have a unified representation of both deterministic and stochastic non-idealities.

3 Harmonic Balance High-Level Simulation

The models discussed in Section 2 were used to implement a frequency domain behavioral simulator, VHSIM. The simulator focuses on deterministic effects. It takes as inputs an arbitrary interconnection of RF blocks with performance parameters in the frequency domain and real harmonic sources, and finds the frequency spectrum of currents and voltages at all nodes. Performance parameters can be either obtained from the specifications of the block in the frequency domain or extracted from transistor-level simulators.

For linear parameters, the results of AC circuit simulations can be included. The coefficients of linear time-varying components (Equation 3) can be extracted by simulators using periodical steady-state analysis as in [8]. Linear interpolation is used to find the value of the parameter at the desired frequency. Second and third order nonlinearities are described by the magnitude and phase of the appropriate Volterra coefficients at different input frequency mixes. For example, the third order nonlinearity of an amplifier can be obtained by simulating the intermodulation performance at different input frequencies. Defining all $H^{(3)}(f_1,f_2,f_3)$ to be equal for all perturbations of f_1,f_2,f_3, third order intermodulation at $2f_1 - f_2$, can be trivially mapped onto the value of the third-order Volterra coefficient at $H^{(3)}(f_1,f_1,-f_2)$. To find the value of a kernel at an arbitrary frequency mix, we use the sum of the input frequencies as a variable to do linear interpolation at the Volterra input-output map look-up table. If many values are found that correspond to different input frequencies with the same sum, f_1 is used to perform interpolation among the selected values and so on.

The simulation algorithm used is *harmonic balance.* If f_i are the frequencies of the K input *harmonic* sources and f_{LO_i} the frequencies of the L periodical strong nonlinearities, the output spectrum contains all possible f_j such as:

$$f_j = \sum_{i=1}^{K} m_i f_i + \sum_{i=1}^{L} l_i f_{LO_i},$$

$$m_i, l_i : \sum_{i=1}^{K} |m_i| \leq M_{max}, \sum_{i=1}^{L} |l_i| \leq L_{max}, f_j \leq f_{max} \quad (6)$$

where M_{max} is the maximum order nonlinear product, L_{max} the maximum side-band sum and f_{max} the maximum important frequency assumed present at the system. Currently, up to third-order nonlinear models for blocks have been considered, but the method is expandable to higher orders at the expense of complexity.

The simulator assembles the sparse-matrix equations for all possible frequencies f_j determined in (6):

$$AI(f_j) = 0 \quad (7)$$

$$V(f_j) - A^T E(f_j) = 0 \quad (8)$$

$$\sum_{f_i + q_l f_{LO_l} = f_j} K_I^{(1)q_l f_{LO_l}}(f_i) I(f_i) + K_V^{(1)q_l f_{LO_l}}(f_i) V(f_i) + \ldots$$
$$\sum_{f_{i_1} + f_{i_2} + f_{i_3} + q_l f_{LO_l} = f_j} K_V^{(3)q_l f_{LO_l}}(f_{i_1}, f_{i_2}, f_{i_3}) \cdot$$
$$V(f_{i_1}) \otimes V(f_{i_2}) \otimes V(f_{i_3}) = S(f_j) \quad (9)$$

where V, I, E are the vectors of branch currents, branch voltages and node voltages respectively, $K_{V,I}^{(i)q_l f_{LO_l}}$ are assembled from the block-level models as in (5), A is the connectivity matrix and "$\otimes$" is the Kronecker product of vectors. Equations (7), (8) are Kirchoff's current and voltage laws respectively and (9) is the generalized nonlinear branch equation in the frequency domain. Assuming all excitations to be real, we need only consider the solutions at $f_j \geq 0$. If J is the number of possible non-negative output frequencies determined in (6), N the number of nodes and B the number of branches, (7)-(9) is a nonlinear, complex system of $J \cdot (N + 2B)$ equations in the frequency domain. Defining $X(f)^T = [I(f_1) \ldots I(f_J) \ldots V(f_1) \ldots E(f_J)]$, (7)-(9) can be re-written as:

$$F(X(f), X(-f)) = 0 \quad (10)$$

Since for real system with real excitations $X(-f) = X^*(f)$, (10) becomes:

$$F(X(f), X^*(f)) = 0 \quad (11)$$

If $X = X_R + \jmath X_I$, $F = F_R + \jmath F_I$ defining $\overline{X}^T = \left[{X_R}^T {X_I}^T\right]$ and $\overline{F}^T = \left[{F_R}^T {F_I}^T\right]$, (11) can be converted into a real system of equations that can be solved with the Newton-Raphson iteration:

$$\overline{X}^{n+1} = \overline{X}^n - \frac{\partial \overline{F}(\overline{X}^n)}{\partial \overline{X}^n}^{-1} \cdot \overline{F}(\overline{X}^n) \quad (12)$$

where

$$\frac{\partial \overline{F}(\overline{X})}{\partial \overline{X}} = \left[\begin{array}{cc} \frac{\partial F_R}{\partial X_R} & \frac{\partial F_R}{\partial X_I} \\ \frac{\partial F_I}{\partial X_R} & \frac{\partial F_I}{\partial X_I} \end{array} \right] \quad (13)$$

At every step of the iteration the linear system is very sparse and can be solved efficiently using sparse matrix techniques. A typical flow of the overall behavioral simulation methodology is shown in Figure 2.

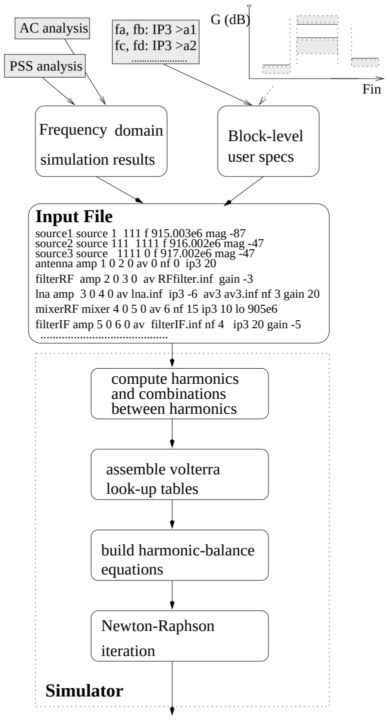

Figure 2: Behavioral Simulator Flow Diagram

Unlike previous work [4], values for the Volterra coefficients*are extracted directly from specifications or extracted performances.* Since no costly differential equation-based models have to be used, the computation time of the Jacobian for the *harmonic balance* system of equations is greatly reduced compared to standard harmonic-balance implementations. The Volterra coefficients are used purely at the *behavioral level*, not a the device level, which results in small nonlinear algebric systems that can be solved very efficiently.

4 Results

The super-heterodyne receiver of Figure 3 was used to test the behavioral simulator VHSIM. The super-heterodyne receiver is intended to operate in the $902 - 928MHz$ band and selects a $60KHz$ channel. It has a sensitivity of $-90dBm$, intermodulation protection of $47dB$ and in-band blocking of -37dB meeting a 10^{-3} BER (10dB SNDR)specification. The results were compared to behavioral models in SPECTREHDL, the high-level language used with SPECTRERF [9]. In simple cases, we also verified the results with exhaustive analytical calculations. In both simulators performance metrics such as filtering, second and third order intercept points (IP_2, IP_3), DC offsets, RF leakages, coupling and frequency translation were modeled. In SPECTRE-HDL, nonlinearities were modeled with polynomials and filters were mapped to 3-5th order Chebyscheff implementations.

The results of the frequency domain behavioral simulator were found to be consistent with the ones from

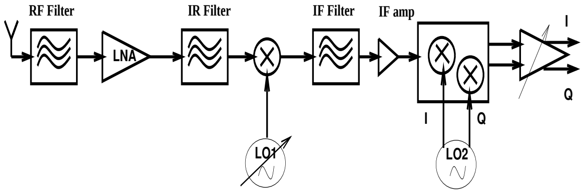

Figure 3: Super-Heterodyne Receiver

		VHSIM	SPECTRERF	
			PSS	*PAC*
2-tones	CPU (sec)	7	1722	220
2-LOs	max error (dB)	0.1	0	1.5
3-tones	CPU (sec)	67.8	1722	220
2-LOs	max error(dB)	0.3	0	3.2

Table 1: Super-heterodyne Receiver Simulation Comparison

SPECTRERF with a huge improvement in simulation time. A three-tone test was used to check the signal-to-interference ratio in the presence of 2 intermodulating signals. For SPECTRERF the frequencies used were: $LO_1 = 905MHz, LO_2 = 10Mz, F_1 = 916MHz, F_2 = 917MHz, F_3 = 917MHz$. Unlike VHSIM that is affected only by the total number of harmonics, simulation time in SPECTRERF greatly depends on the frequency spacing and the order of the filters used. For SPECTRE to converge, we had to increase the spacing of the tones and relax the specifications of the filters. The results are summarized in Table 1. VHSIM simulations were run on a DEC Alpha-Server 600MHz and all CPU times are normalized to that. The maximum error refers to the total SNDR and the result of a periodic steady-state analysis is used for reference. A typical VHSIM output is shown in Figure 4. Figure 5 shows the simulation time in VHSIM for different number of input tones, while changing f_{max}(Equation 6). Table 2 summarizes the performance of the simulator for 4 and 5 input tones. N_1, N_2, are the number of input and LO frequencies, F_{max} the maximum upper frequency, J_{TOT} the total number of frequency combinations, J the number of output frequencies used, $\#N_z$ the percentage of non-zero Jacobian elements and N the total number of branches and nodes.

5 Conclusions

A frequency-domain, "black box" representation of deterministic effects of RF systems has been described. The modeling is based on the Volterra theory of nonlinear systems. The behavioral harmonic balance simulator developed based on those models, is significantly faster than time-domain macro-modeling and can accurately simulate deterministic effects such as intermodulation, compression, frequency translation. The simulator can be used both in system design to partition high-level specifications of RF systems to blocks, as well as in system verification given extracted high level, frequency domain performance parameters. System-level verification results compare accurately available behavioral simulation techniques with great improvement in simulation time. Furthermore more complicated interactions can be accurately modeled, such as frequency-dependent nonlinearities. Currently we are working in comparing a full transistor simulation of a receiver with a behavioral simulation.

N_1, N_2,	F_{max}	CPU (s)	J_{TOT}	J	N	$\#N_z\%$
2,2	2.1e9	7.1	1241	168	8604	0.05
2,2	6e9	17	1241	253	12144	0.03
3,2	1.1e9	9.1	16807	217	11067	0.03
3,2	2.1e9	67	16807	428	19788	0.02

Table 2: VHSIM Performance Table

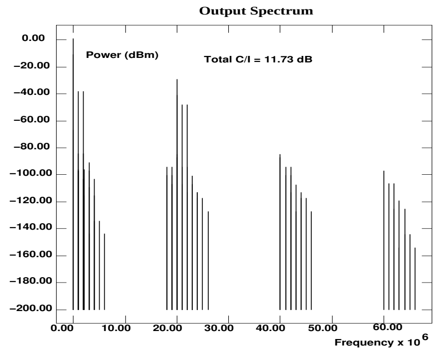

Figure 4: Simulator Output

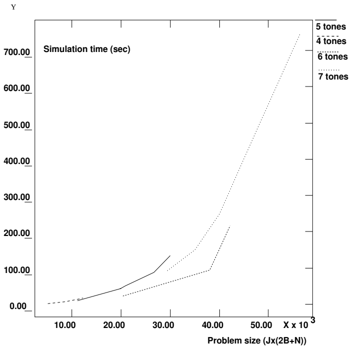

Figure 5: Simulation Speed

Acknowledgments

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