Nonlinear Analysis Methods for the Simulation of Digital Wireless Communication Systems

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ABSTRACT

Adoption of digital modulation for many wireless communication systems has resulted in significant performance improvements over systems based on analog modulation. Concomitant with this have been changes in methods of characterizing and simulating system performance, particularly with respect to amplifier linearity. Digitally modulated signals are best represented by a power spectral density. In contrast, analog-modulated signals are adequately represented by discrete spectra. Consequently, many of the common discrete spectra nonlinear microwave analysis techniques are ill-suited to simulating systems characterized by signals with power spectral density representations. This study examines some of the common types of signals used in digital wireless communication systems, and looks at their representation and characterization. The applicability of various nonlinear analysis methods to microwave power amplifier simulation in the context of digital modulation is explored. © 1996 John Wiley & Sons, Inc.

INTRODUCTION

Most present and future wireless communication systems rely exclusively on digital modulation, in contrast to first-generation systems, which were based on analog modulation. Digital modulation offers increased channel capacity, improved transmission quality, secure communication, and the ability to provide other value-added services [1]. These next-generation systems, however, present new challenges to the microwave engineer with respect to representation and characterization of digitally modulated signals, and also with respect to nonlinear simulation of digital wireless communication systems. Digital modulation is fundamentally different from analog modulation: the former is characterized by a signal represented as a power spectral density (PSD), and the latter is usually characterized by a signal represented as discrete spectra. The microwave engineer is thus posed with a representation and characterization problem, which impacts the choice of CAE tools for the optimum tradeoff in simulation robustness, generality, speed of execution, and memory utilization. In this article, we consider the representation and characterization problem for digitally modulated signals, and within the framework of an adequate understanding, examine the various nonlinear simulation methods available. Time-domain, mixed-domain, frequency-domain, and measurement-based behavioral modeling methods are examined, with emphasis placed on simulation of microwave power amplifiers intended for use in portable digital wireless communication systems.

This article begins with a discussion of the representation and characterization of digitally modulated signals, which is necessary to establish which nonlinear analysis methods are amenable to excitation by signals described by PSD repre-

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sentations. A comprehensive introduction to characterization of linearity in the context of digital modulation is also presented. Time-domain analysis methods are considered first. Common variants, such as direct integration, discrete associated modeling (SPICE), and shooting methods are covered. Time-stepping and its impact on fast Fourier transformation is also considered. Next we look at hybrid methods, commonly referred to as harmonic balance methods. Time-invariant and time-variant approaches are considered, as are tone-spacing and fast Fourier transformation issues. Frequency-domain methods are presented next, with emphasis on the Volterra nonlinear transfer function approach. Last we discuss a behavioral modeling method based on AM–AM and AM–PM measurements.

REPRESENTATION AND CHARACTERIZATION OF DIGITALLY MODULATED SIGNALS

The Representation Problem

Spectral occupancy analysis is of fundamental importance in wireless communication system design as it is representative of spectral efficiency (revenue for the service provider) and time-domain peak-to-average ratio (talk-time for the subscriber). It is therefore necessary to deterministically represent and characterize digitally modulated signals in both the frequency- and time-domains. Classical spectral occupancy analysis of analog-modulated signals are done typically with discrete spectra methods. For example, with narrow-band frequency modulation Carson's rule can be used to predict spectral occupancy [2]. With amplitude modulation a time-domain description of the signal, based on knowledge of envelope statistics, allows the peak-to-average ratio to also be predicted [2]. However, with digital modulation, stochastic process theory must be utilized to deterministically represent and characterize the signal.

Bandpass digital and analog modulation alters some characteristic or characteristics of a carrier signal, such as amplitude, frequency, or phase. The fundamental difference between the two methods is that with digital modulation the modulated characteristic(s) have a number of prescribed discrete states; such discretization does not occur with analog modulation. Since a state change is unknown a priori, it must be modeled as a stochastic process. This process is preferably based on a data sequence that causes all possible states of the modulated signal to occur in equal proportion over a finite time interval. A time-domain representation of a signal modulated by this data sequence is not unique: the time-domain representation space is infinite. Only the time-domain statistics, typically characterized by the first and second moments, are constant. Having thus imposed ergodicity, and therefore, wide-sense stationarity, the frequency-domain representation of this signal is a function only of the time-domain statistics (and filtering, which is deterministic). With a given modulation format, ergodic data sequences, with equal first and second moments, will have identical PSD representations [3, 4]. It is therefore desirable that a digitally modulated signal used to characterize amplifier performance should have deterministic statistics independent of absolute time, for example, a maximal-length sequence [4].

A frequency-domain representation is derived by considering one bit of an infinitely long ran-

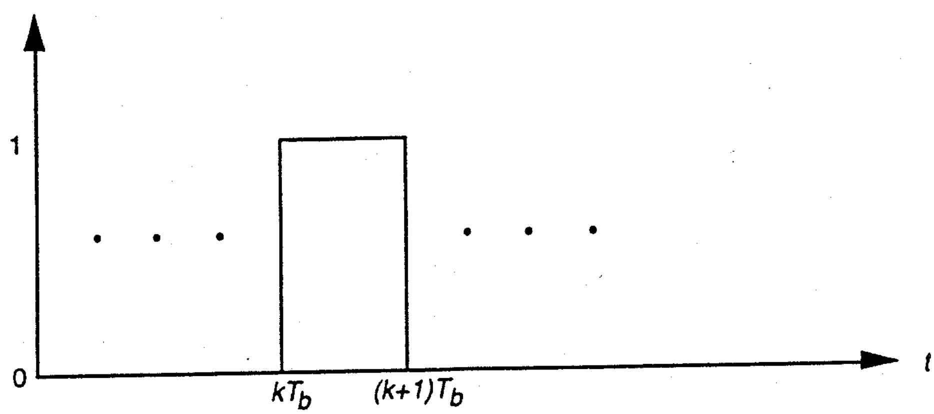

Figure 1. One bit of a random data sequence.

dom data sequence, shown in Figure 1. Since a bit transition is possible anywhere within the region kT_b to $(k+1)T_b$, where $k = 1, 2, 3 \ldots$, a deterministic representation is not possible. If, however, we assume the probability of a bit transition is equiprobable over the interval kT_b to $(k+1)T_b$, then the frequency-domain representation can be evaluated deterministically. Assuming that a binary "1" and "0" are mapped to +A and −A respectively, that is, a non-return-to-zero mapping, and that adjacent bits are uncorrelated, the PSD is:

$$S(f) = A^2 T_b \operatorname{sinc}^2 T_b f \qquad (1)$$

where T_b is the bit rate. Observe from eq. (1) that the ratio of the two adjacent sidelobes to the central lobe is approximately −13 dB. This characteristic is interpreted as adjacent-channel interference, and is typically controlled by a special class of band-limiting filters called Nyquist filters [4]. This band-limiting operation directly impacts the time-domain peak-to-average ratio of the modulated carrier, and must therefore be included in the representation of a digitally modulated signal. Note further that a correlated data sequence will skew the shape of the PSD, leading to a biased frequency-domain representation. This influences the adjacent sidelobe to central lobe ratio, and may impact linearity characterization, as demonstrated in the following section.

Eq. (1) is derived assuming an infinitely long data sequence. All data sequences are finite, and perhaps periodic (as is the case with maximal-length sequences), so that the PSD represented by eq. (1) is in general expressed as a finite summation of incommensurate tones approximating a PSD. If eq. (1) is generated by fast Fourier transforming a time-domain representation, the number of tones comprising the PSD will be related to the number of bins in the fast Fourier transform. To generate a time-domain representation we must consider, in addition to the data sequence, the form of modulation employed. Most digital wireless communication systems have adopted some form of phase-shift-keying (PSK) or amplitude-phase-shift-keying (QAM), with one exception being the European digital cellular standard, which is based on frequency-shift-keying (FSK) [5–9]. Our focus here is the North American digital cellular system, which uses $\pi/4$-DQPSK (differential quadrature phase-shift-keying) modulation. This modulation format consists of two orthogonal PSK signals with the data sequence encoded differentially. To generate a time-domain representation the generalized quadrature modulation equation:

$$s(t) = i(t)\cos[\omega_c t + \phi_i(t)] - q(t)\sin[\omega_c t + \phi_q(t)] \qquad (2)$$

is used, where $i(t)$ and $q(t)$ embody the particular modulation rule for amplitude, $\phi_i(t)$ and $\phi_q(t)$ embody the particular modulation rule for phase, and ω_c is the carrier radian frequency. A time-domain envelope representation following the NADC standard without Nyquist filtering is shown in Figure 2; the corresponding frequency-domain representation, using a 1024-point FFT, is shown in Figure 3. Observe in this example that the envelope of this signal is nearly constant, although Nyquist filtering will impart a 3.6-dB peak-average ratio. The methods presented here are general and apply to other modulation formats used in digital wireless communication systems, such as QAM and MSK. However, for the latter modulation a continuous-phase representation is more convenient than the quadrature modulation representation used here [4].

Representation of a digitally modulated signal requires an explicit form of the modulation rule, an expression for the modulation characteristic, and an expression for Nyquist filtering. Then, provided a maximal-length data sequence is used, the notions of ergodicity and wide-sense stationarity provide confidence that the time-domain representation of a digitally modulated signal is an accurate and repeatable representation of the data sequence. This ensures that amplifier performance characteristics, such as linearity, are not influenced by latent data sequence properties, with the resultant ambiguities in performance evaluation.

The Characterization Problem

In the preceding section it was demonstrated that digitally modulated signals are conveniently represented by continuous spectra, in contrast to analog-modulated signals, which are usually represented by discrete spectra. Concurrent with this representation difference are alternative methods of system performance characterization. Concepts such as bit error rate, adjacent-channel power, and transient spectrum regeneration are now the performance metrics for digitally modulated signals. Since the focus in this article is nonlinear simulation of microwave power amplifiers, atten-

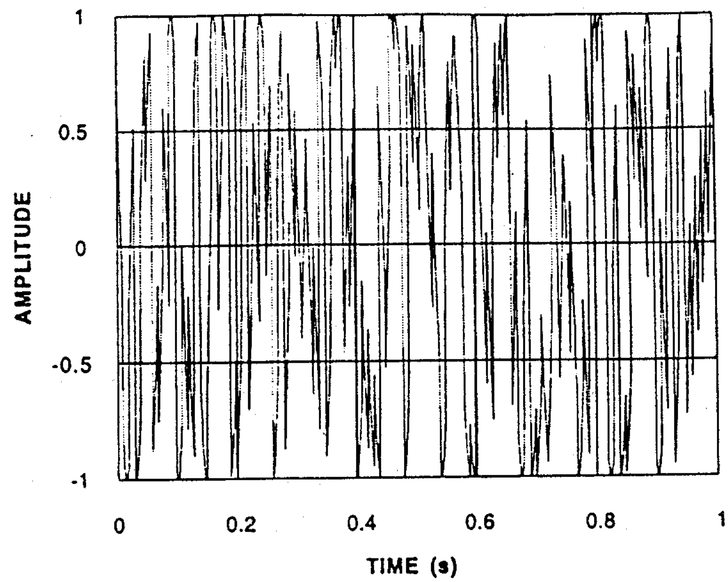

Figure 2. Time-domain representation of a $\pi/4$-DQPSK-modulated signal using an uncorrelated NRZ data sequence.

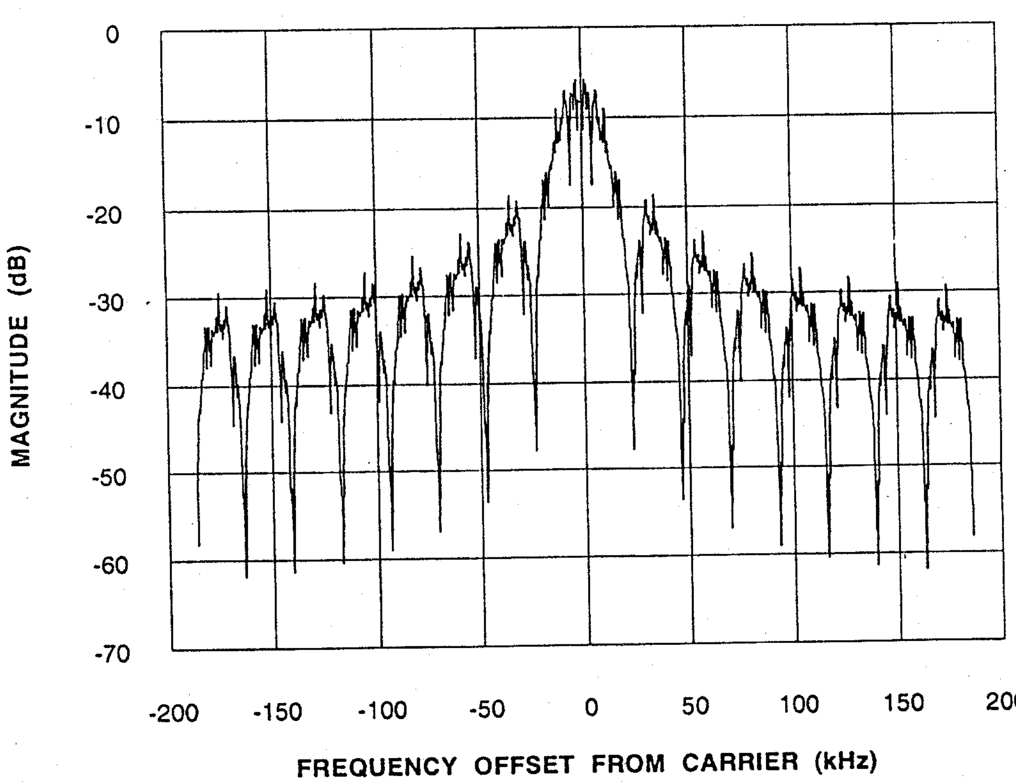

Figure 3. Frequency-domain representation of a $\pi/4$-DQPSK modulated signal using an uncorrelated NRZ data sequence.

tion will be placed on the latter two concepts. Adjacent-channel power arises from spectrum regeneration, the process wherein a band-limited digitally modulated signal passes through a nonlinearity and a portion of the band-limited spectrum regenerates. Amplifier linearity in the context of digital modulation is therefore most suitably characterized by measuring the degree of spectrum regeneration. This is done by comparing the power in the two adjacent channels to the power in the main channel: the upper and lower adjacent-channel power ratios (ACPR). Control of ACPR is a critical design issue since stringent specifications limiting ACPR directly impact the efficiency of a microwave power amplifier [5, 6, 9]. Also, note that upper and lower ACPR are in general not equal.

Using the North American digital cellular specification, a band-limited nonlinearly amplified finite sum approximation of the PSD (1) is shown in Figure 4 [5]. The lower channel ACPR is defined as:

$$ACPR_{lower} = \frac{\int_{f_3}^{f_4} |H(\alpha, T_s)|^2 S(f)\,df}{\int_{f_1}^{f_2} |H(\alpha, T_s)|^2 S(f)\,df} \tag{3}$$

where the denominator represents the power in the main channel, $S(f)$ is the PSD function at the output of the amplifier under test (which has been band-limited prior to amplification), and $H(\alpha, T_s)$ represents the remaining half of the band-limiting Nyquist filter. (The Nyquist filter most commonly used is the root-raised cosine filter. Filtering is equally split between the transmitter and receiver to optimize the SNR at the receiver.) The expressions for upper and lower alternate-channel power ratios are similar, with the limits of integration moved up or down by one channel spacing. In the last section it was mentioned that a correlated data sequence would result in an asymmetric PSD. From eq. (3) it is clear how such an asymmetry would impact the characterization of ACPR, and consequently, the linearity characterization of the amplifier. Therefore, it is vitally important that a maximal-length data sequence be used for simulation- and measurement-based adjacent- and alternate-channel power ratio characterization [6, 10].

Note from Figure 4 that the integration bandwidth for the main-channel and the adjacent-channel powers are wider than the channel spacing, which is 30 kHz. Consequently, some power in the main channel will always be present in the

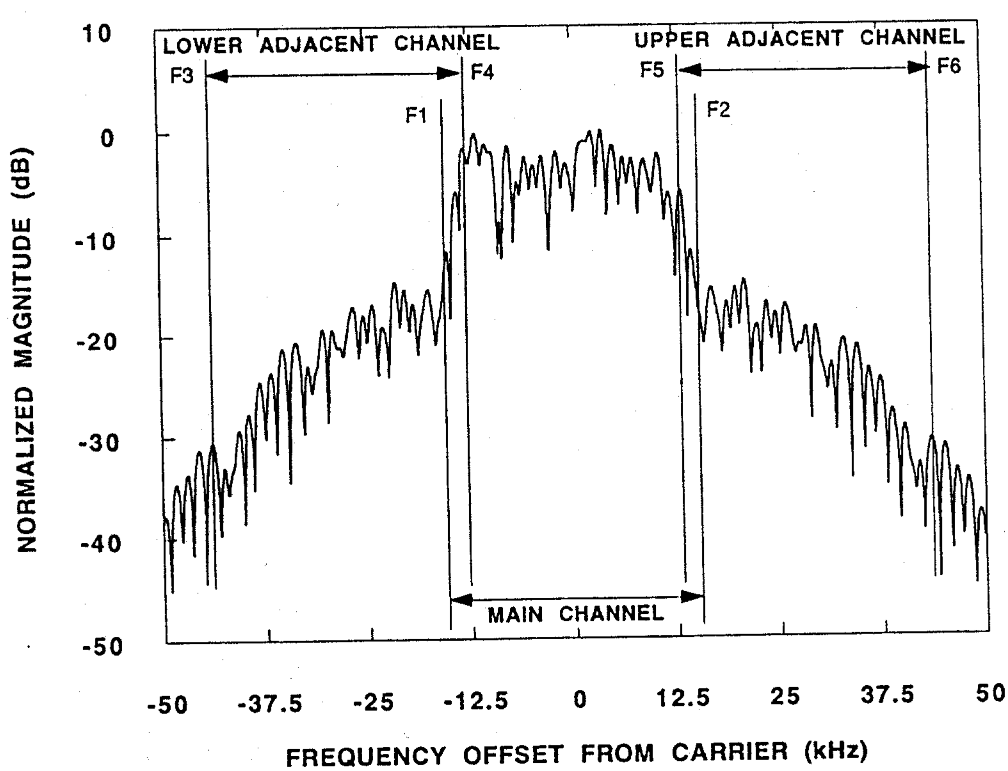

Figure 4. Definition of adjacent-channel and main channel integration limits.

two adjacent channels, independent of amplifier linearity. For the NADC system, this establishes an ideal minimum ACPR of −36.4 dBc. Overlap is a function of the channel spacing and Nyquist filter roll-off. Some systems, such as the Japanese digital cellular system, do not have overlap, the ideal minimum ACPR being established by thermal noise over the integration BW [6].

Another important characteristic is that of transient spectrum regeneration. This arises primarily in systems based on time division multiple access (TDMA), such as the NADC system and the PDC system [5, 6]. Since these systems are segregated in time (and frequency), the carrier is periodically keyed on and off. Transient spectrum regeneration results in a decrease in the ACPR due to transient energy of the bursted signal. The degree of reduction of ACPR due to the transient spectrum regeneration is strictly a function of the time-domain ramp function. Ramp functions such as the Gaussian function are normally used due to their strongly confined spectral content [11]. The ability to discriminate between steady-state and transient spectrum regeneration is useful in establishing the proper ramp characteristics, and thus requires a simulation method capable of supporting time-dependent frequency-domain nonlinear analysis.

Characterization of digitally modulated signals requires the ability to represent PSDs; the ability to include the effect of band-limiting; and, in some cases, the ability to look at bursted signals in the frequency domain as a function of time. Additionally, the ability to generate digitally modulated signals, as discussed in the previous section, is necessary. With a clear understanding of the most basic elements of the representation and characterization problem, we can now consider the amenability of various nonlinear analysis methods for the simulation of digital wireless communication systems.

TIME-DOMAIN ANALYSIS METHODS

Simulation of nonlinear microwave circuits presents many problems that usually preclude, or at least limit, the use of certain analysis techniques. The requirements placed on nonlinear analysis methods amenable to representation and characterization of digitally modulated signals impose additional functional requirements, further limiting the number of methods that are usable. Time-domain analysis methods fall in the latter category. Although time-domain methods have the unrivaled ability to perform transient analysis, many characteristics of interest in both the nonlinear simulation of microwave power amplifiers and the characterization of digitally modulated signals are best handled in the frequency domain. In this section we will consider the three most common types of time-domain analysis methods: direct integration; associated discrete modeling (SPICE); and the shooting method.

Direct Integration Method

The direct integration method is perhaps the simplest nonlinear analysis method available [12, 13]. The network equations are usually written in modified nodal admittance matrix form to yield a nonlinear implicit equation of the form:

$$\mathbf{f}(\dot{\mathbf{x}}, \mathbf{x}, \mathbf{w}, t) = 0 \tag{4}$$

where $\dot{\mathbf{x}}$ is a vector of voltage and current time-derivatives, $\mathbf{x}$ is a vector of voltage and current variables, and $\mathbf{w}$ is a vector of nonlinear charge and flux terms [12, 14]. Solution of eq. (4) for all t is usually found using a predictor-corrector method, involving transformation of eq. (4) into a nonlinear algebraic equation using a backward difference formula. The resulting nonlinear algebraic equations are then solved using Newton's method.

Several properties of microwave power amplifiers and digitally modulated signals make apparent the limitations of the direct integration method. These properties include: circuit time constants differing by several orders of magnitude; a modulation envelope that is slowly varying with respect to the carrier frequency; and circuit elements that are difficult to describe in the time domain (e.g., transmission lines). Consequences of these properties are, respectively: a time step controlled by the largest time constant for integration stability, but a long simulation time required to reach steady-state; special preprocessing requirements for adequate fast Fourier transformation of the slowly varying envelope; and simplification of physical equations, or advanced methods, for time-domain characterization of frequency-dependent elements [14, 15].

Adaptive time-stepping algorithms coupled with variable-order integration routines can reduce the effect of widely separated time constants [12, 16]. These methods present difficulties for subsequent frequency-domain characterization, however. Since the FFT requires uniform time samples, interpolating polynomials usually are used to transform the nonuniform time samples of the adaptive time-stepping algorithm to uniform time samples suitable for the FFT. This process inherently reduces the dynamic range of the FFT due to interpolation noise. More importantly, the effect of additive numerical noise due to long simulation times, necessary for suitable frequency resolution for characterization of ACPR, has not been adequately resolved. Additive numerical noise therefore manifests itself as tradeoff in FFT dynamic range and frequency resolution. Currently, state-of-the-art FFT algorithms based on interpolation polynomials have 80 dB to 100 dB of dynamic range, this being unacceptable for some digital wireless communication systems [7, 17]. A frequency resolution of 100 Hz is estimated as the minimum necessary for adequate characterization of ACPR for the NADC and PDC standards thus placing a lower bound on simulation time (e.g. window width).

A further complication can be seen in the generation of digitally modulated signals in the time domain. With emphasis on microwave power amplifier nonlinear simulation, generation of a frequency-shifted digitally modulated source is appropriate. We thus obviate the need for implementing eq. (2) directly in the simulation, instead implementing it numerically as an ideal vector modulator. For a phase-modulated signal such as $\pi/4$-DQPSK, eq. (2) can be written in discrete form, with filtering, as:

$$\begin{aligned} x(t_n) = {} & \left\{h(t_m) * \exp\left[j\left({}^i\theta_m + {}^i\theta_{m-1}\right)\right]\right\}\exp(j\omega t_n) \\ & - \{h(t_m) * \exp[j({}^q\theta_m + {}^q\theta_{m-1})]\} \\ & \times \exp(j\omega t_n + \pi/2) \end{aligned} \tag{5}$$

where m is the modulation time index, n is the carrier index, $h(t_m)$ is the impulse response of the band-limiting filter, "$*$" represents convolution, and i and q represent the in-phase and quadrature components of the modulating source. Note that incorporation of adaptive time stepping with eq. (5) represents an incompatibility with the convolution operation, which usually requires uniform time stepping. Similar expressions can be easily derived for other modulation formats as well, such as continuous-phase modulation for the European digital cellular system (GSM) [8].

Associated Discrete Modeling (SPICE)

SPICE, and related commercial derivatives, are the most common time-domain nonlinear simulation tools in use today [18, 19]. SPICE is similar to the direct integration method except that the order of integration and time-discretization are reversed. The problem of solving a set of coupled nonlinear differential equations, for example, eq. (4), is thus reduced to a finite-difference problem requiring solution of a set of linear algebraic equations of the form:

$$\mathbf{M}({}^j\mathbf{x})^{j+1}\mathbf{x} = {}^j\mathbf{y} \tag{6}$$

where $\mathbf{M}$ is the modified nodal admittance matrix representing the network equation, $\mathbf{x}$ is a vector of unknown node voltages and branch currents, $\mathbf{y}$ is a vector of sources, and j is the Newton iterate. This equation is solved once for each time step [12, 13]. Note that this method replaces all elements with equivalent circuit models consisting of only a conductance and current source, in contrast to the direct integration method, which uses an implicit element formulation. A significant advantage of this method is that convergence can be controlled locally, resulting in superior convergence properties.

Since only the network formulation differs, the associated discrete model method suffers all of the basic limitations of the direct integration method. This would not be readily apparent since eq. (6) appears to be independent of time step. The time step appears, however, in the finite-difference description of each element, and thus impacts directly the convergence properties of eq. (6). Most versions of SPICE, therefore, adopt adaptive time stepping, requiring interpolation polynomials for the FFT. Similarly, frequency-domain elements are modeled in the same fashion as with the direct integration method.

The Shooting Method

The shooting method potentially avoids the issue of the long simulation time associated with the previous methods by attempting to find the steady-state solution directly [20, 21]. The state equations, eq. (4), are placed into explicit form and integrated once per iteration (in contrast to the previous methods) so that an initial condition $\mathbf{x}(t_0)$ is found such as:

$$\int_0^T \mathbf{f}[\mathbf{x}(t_0), \tau]\, d\tau \le \varepsilon \tag{7}$$

is satisfied to an arbitrary tolerance, ε. Since the definition of periodicity is $\mathbf{x}(t) = \mathbf{x}(t + T)$, a gradient method may be employed to find $\mathbf{x}(t_0) = \mathbf{x}(T)$. Note that this method is faster than the previous methods if the number of iterations to find the steady-state initial condition is less than the number of periods necessary to reach steady state.

Although this method avoids the problem of additive numerical noise due to long simulation times, it still suffers from nonuniform time stepping. Additionally, digitally modulated signals possess slowly varying envelopes which are only approximately periodic. Elements modeled in the frequency-domain would be handled in a manner similar to the previous methods, a major drawback being that the initial conditions must also be established at every point along any transmission lines present.

Assessment of Time-Domain Methods

The ability of the direct integration and the associated discrete model methods to represent arbitrary signals conceptually facilitates representation of digitally modulated signals. However, the ratio of the minimum time-step to the sequence length for a typical maximal-length sequence imposes a severe memory penalty on methods already requiring a large of amount of memory. As an example, a typical sequence length for the NADC format is 10 ms, although the time step would be on the order of 100 ps to capture the fundamental and harmonics of the 850 MHz carrier [5]. Representation of arbitrary ramp functions and bursted analysis is possible, albeit with a similarly severe memory penalty. Each method suffers from frequency-domain characterization limitations, with limitations manifested as a tradeoff in dynamic range and frequency resolution. Finally, some researchers describe a fundamental approximation error present in the SPICE algorithm, due to what amounts to a z-domain approximation to the frequency-domain characteristics of the circuit [22].

MIXED-DOMAIN ANALYSIS METHODS: HARMONIC BALANCE

Microwave circuits in general, and microwave power amplifiers in particular, exhibit properties that are best handled with steady-state frequency-domain analysis methods with quasiperiodic excitations. As well, characterization of digitally modulated signals is best handled in the frequency-domain. However, representation of nonlinearities and of digitally modulated signals are best expressed by algebraic time-domain expressions. Therefore, a method to combine analyses in both domains would possess distinct advantages over strictly time-domain methods. The harmonic balance method is one such technique. The harmonic balance method is inherently efficient for steady-state frequency-domain analysis since a solution form (phasor magnitude and angle) is imposed a priori, in contrast to time-domain methods, which use no similar assumption [23]. Evaluation of the nonlinear device characteristics in the time-domain and the subsequent representation and characterization of digitally modulated signals in the frequency-domain is accomplished using the fast Fourier transform. In contrast to time-domain methods with nonuniform time samples, most implementations of the harmonic balance method incorporate precisely spaced time and frequency sample points. This attribute provides a dynamic range in excess of 200 dB, and is ideally suited to the characterization problem [24].

This section presents two types of harmonic balance methods: the time-invariant phasor method and the time-variant phasor method. The former is currently available in several CAE tools [25–27]. A preliminary commercial version of the latter was released in late 1995 [25, 28, 29]. Both methods can conceptually be used to simulate digital wireless communication systems, although the time-variant harmonic balance method has significant advantages with respect to frequency-domain characterization accuracy, circuit and simulation generality, and memory utilization.

Harmonic Balance Using Time-Invariant Phasors

The harmonic balance technique in its most common form is referred to as frequency-domain time-invariant harmonic balance since the assumed solution form is based on phasors (the time-domain form of harmonic balance imposes an assumed solution form on a sequence of time samples) [30–34]. Although most commercial harmonic balance algorithms use the modified nodal admittance matrix network formulation, the idea of harmonic balance is most easily demonstrated using the approach of Nakhla and Vlach [32]. This approach is based on partitioning of the linear and nonlinear portions of a circuit as shown in Figure 5. The $I_{s,j}$ represent excitation sources; $I_{m,k}$ and i_k represent frequency-domain and time-domain interfacial currents, respectively; and $V_{m,k}$ and v_k represent frequency-domain and time-domain node voltages, respectively. Indices m and k are harmonic number and node number, respectively. A frequency-domain solution at the kth node up the the nth harmonic of the form:

$$\mathbf{V}_k(j\omega) = \text{real}\left\{ \sum_{m=0}^{n} V_m \exp(j\omega m + \phi_m) \right\} \quad (8)$$

is assumed, with the unknown variables being the amplitude and phase of each frequency component at each interfacial node. Note that phasor amplitude and phase at each frequency is fixed with respect to time, hence the name time-invariant harmonic balance. The solution, then, is a finite sum of commensurate discrete spectra.

Conservation of current and charge in the frequency-domain at each of the interfacial and source nodes of Figure 5 with $\mathbf{V}$ a vector of node voltages gives:

$$\mathbf{F(V)} = \mathbf{I(V)} + \Omega\mathbf{Q(V)} + \mathbf{YV} + \mathbf{I}_s = \mathbf{0} \quad (9)$$

where the first two terms on the right side are calculated as:

$$\mathbf{F}_{NL}(\mathbf{V}) = \mathfrak{F}\mathbf{i}(\mathfrak{F}^{-1}\mathbf{V}) + \Omega\mathfrak{F}\mathbf{q}(\mathfrak{F}^{-1}\mathbf{V}) \quad (10)$$

where $\mathfrak{F}$ is the Fourier transform operator and Ω is a diagonal matrix of $j\omega m$ terms representing frequency-domain differentiation. Solution is carried out in the frequency domain, with nonlinearities (expressed in the time-domain) transformed to the frequency domain using an FFT with precisely spaced time samples. With a Newton–Raphson type solution method eq. (9) can be solved iteratively as:

$$^{j+1}\mathbf{V} = {}^{j}\mathbf{V} - \mathbf{J}^{-1} \cdot \mathbf{F}(^{j}\mathbf{V}) \quad (11)$$

where $\mathbf{J}$ is the network Jacobian function [24, 33, 34].

As with time-domain methods, a representation of the digitally modulated signal under consideration is necessary. Two approaches are possible: a time-domain representation, or a frequency-domain representation. The time-domain representation would consist of an expression similar to eq. (5), with the digitally modulated source then embedded within the nonlinear network of Figure 5. Alternatively, a frequency-domain representation of the form

$$Y(j\omega) = \sum_{m=-n}^{n} H(j\omega m)X(j\omega m) \quad (12)$$

is possible where $H(j\omega m)$ is the transfer function of the band-limiting Nyquist filter and $X(j\omega m)$ represents the spectrum of the modulation format under consideration. Note, however, that $X(j\omega m)$ will ultimately be based on time-domain data, since a random process is most easily specified by a time-domain data sequence. Once $X(j\omega m)$ is generated from the time-domain data sequence, then at the interfacial nodes of Figure 5 the frequency-domain representation of $Y(j\omega)$ is represented as a finite sum of sinusoids. It is important to observe that the resolution of the PSD approximation in eq. (12) is controlled by the maximum order of the harmonic balance analysis.

Generation of a digitally modulated signal in HP MDS is based on direct implementation of a vector modulator, with a $\pi/4$-QPSK encoder excited by a user-defined bit sequence. A two-tone

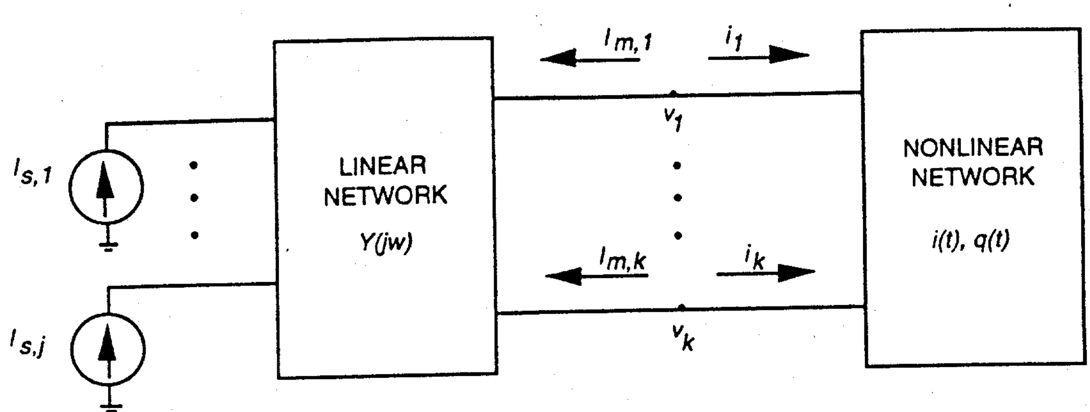

Figure 5. Partition representation of the time-invariant harmonic balance method [32].

simulation is used with one tone up-conversion frequency of the vector modulator. The second tone is representative of the instantaneous phase between the in-phase and quadrature components, which is simply frequency, and thus can be modeled as discrete spectra corresponding to each phase transition. The maximum number of tones representable, and thus the accuracy of the approximation to the PSD with a finite sum of sinusoids, is limited by the maximum order of the simulation. Figure 6 shows the frequency-domain representation of the vector modulator for the NADC format with maximum order set to 32 and a bit sequence of 32 bits. Figure 7 shows the same signal after passing through a simple memoryless unilateral cubic nonlinearity. Solution of this simple nonlinearity required over 45 megabytes of RAM and nearly 10 minutes of compute time on a Sun Sparc 20. Observe the spectrum regeneration occurring in the adjacent channels. Convergence in the presence of strong nonlinearity, for example, simulation of compression, was not possible. Consequently, simulation of practical nonlinearities is not possible with this method; however, simulation of weak nonlinearities should be possible.

Addressing the characterization problem using harmonic balance is straightforward since the output is in the frequency domain. Two approaches are possible: carrier-level processing, or baseband processing. For NADC ACPR measurements, multiplication of the proper spectral components by the remaining half of the band-limiting filter transfer function would yield the correct ACPR. Note that this process is essentially a

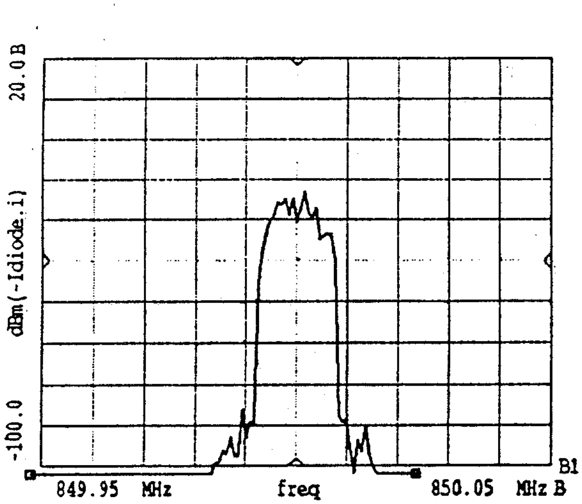

Figure 6. PSD of a $\pi/4$-QPSK signal using time-invariant harmonic balance.

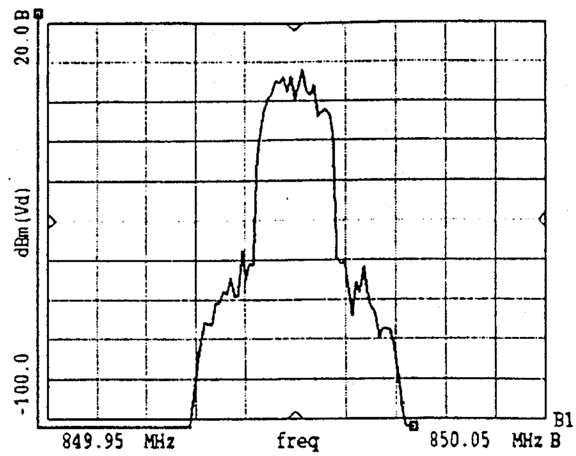

Figure 7. PSD of signal of Figure 6 after passing through a zero feedback memoryless cubic nonlinearity.

discrete integration implementation of eq. (3). The main channel power is determined in a similar fashion. The latter method is identical to the former except that the output of the nonlinear system is downconverted (using a vector demodulator) to baseband. Since this method would involve additional simulation time, the carrier-level approach is preferred. Burst analysis, for simulation of transient spectrum regeneration, would be prohibitive due to the number harmonics needed to model the rapid transitions in the time-domain.

Harmonic Balance Using Time-Variant Phasors

Harmonic balance using time-variant phasors represents the state of the art in current nonlinear analysis research, with a commercial version having been released in late 1995 [25, 28, 29]. Time-variant harmonic balance is ideally suited to the representation and characterization problem. In contrast to time-variant harmonic balance, in which the assumed phasor solution was time-invariant, we instead assume a solution of the form:

$$\mathbf{V}_k(j\omega) = \text{real}\left\{ \sum_{m=0}^{n} V_m(t)\exp[jm\,\omega(t) + \phi_m(t)] \right\} \tag{13}$$

where, in general, the amplitude, frequency, and phase of each term are allowed to vary with respect to time. If $V_m(t)$ varies slowly with respect to the carrier frequency we are in essence solving for the envelope of the signal at each node without the requisite memory requirements of time-invariant harmonic balance, or the frequency-domain dynamic range and resolution problems of time-domain methods. Taking the Fourier transform of each summation term in eq. (13) results in a highly resolved PSD approximation of the digitally modulated signal, not an ill-conditioned approximation, as with time-invariant harmonic balance.

Since a highly resolved frequency-domain representation based on the signal envelope is used, representation of an uncorrelated data sequence is possible without the memory penalty and convergence problems associated with time-variant harmonic balance. As well, all nonlinear device models that have been implemented in time-invariant harmonic balance, for example, Curtice quadratic/cubic and Gummel–Poon, can be used, with simulation of amplifier compression easily handled. This method is also capable of simulating ACPR source- and load-pull contours, as well as time-dependent frequency-domain analysis, for example, burst simulation.

Assessment of Mixed-Domain Methods

Several practical limitations in time-invariant harmonic balance exists. That time-invariant harmonic balance converges only for weak nonlinearities is easily explained. Spectral decomposition of a digitally modulated signal ideally would result in a spanning set equal to a linear combination of the signal-space basis vectors. The short data sequences required with the present method are correlated, resulting in a spanning set consisting of (some) linearly dependent vectors. These vectors subsequently result in the Jacobian being ill-conditioned in eq. (11), leading to problems in solving the network matrix equation. Next, the direct relationship between the length of the data sequence and memory requirements for approximation of the PSD of the signal with discrete spectra limit sequence length independent of convergence problems. With the HP MDS a 32-bit limitation exists for the data sequence; longer sequences would require additional tones, and hence additional memory (other commercial implementations of harmonic balance would have similar memory limitations). Finally, although band-limiting Nyquist filtering is incorporated easily, many of the common filter specifications require roll-off of 100 dB over 100 Hz [5–9]. Hence, tone spacing becomes critical as movement of a single tone in or out of the filter passband directly impacts the prediction of ACPR. A minimum tone spacing of 468 Hz was achieved with the present method, being too large for accurate characterization of ACPR.

Time-variant harmonic balance addresses each of the problems listed above. Since a highly resolved PSD approximation based on the carrier envelope is the assumed solution form, uncorrelated data sequences can be used without a memory penalty, resulting in improved convergence. This method can use any of the nonlinear device models that have been developed for use in time-invariant harmonic balance, ensuring generality. Moreover, band-limiting roll-off error would not be an issue. A frequency-domain resolution of 50 Hz is possible, thereby circumventing entirely the issue of ACPR ambiguity due to tone spacing.

FREQUENCY-DOMAIN ANALYSIS METHODS

Frequency-domain nonlinear analysis methods avoid many of the problems inherent to time-domain and mixed-domain nonlinear analysis methods, as well as being ideally suited to the representation and characterization of digitally modulated signals. Since all operations are performed in the frequency-domain, Fourier transformation is unnecessary. Frequency resolution and dynamic range limitations, due to long simulation times, are therefore reduced. Consequently, it is possible to achieve dynamic ranges in excess of 400 dB with frequency resolution suitable for characterization of ACPR [35, 36]. As well, phenomena usually described in the frequency-domain, such as transmission line coupling and dispersion, are easily incorporated. A significant advantage of some frequency-domain nonlinear analysis methods, such as the Volterra nonlinear transfer function method, is that closed-form expressions can be developed, thus providing insight otherwise not available with strictly numerical methods such as the time-domain and mixed-domain methods previously described [37–39]. That frequency-domain nonlinear methods are not widely adopted, however, is due to the considerable effort expended to develop appropriate frequency-domain models for nonlinear elements.

Many types of frequency-domain nonlinear analysis methods exist, ranging from simplified power series analysis methods to iterative spectral-domain techniques [35–43]. The fundamental premise of frequency-domain nonlinear analysis methods is that input–output descriptions can be derived for a wide class of nonlinear elements, circuits, and systems. For example, the Volterra nonlinear transfer function method relates discrete input spectra to discrete output spectra up to an arbitrary order when algebraic nonlinear current and charge expressions are available [36–39, 44–46]. If a narrow-band assumption is made, representative of most digital wireless communication systems, the discrete spectra can be replaced by a high-resolution PSD approximation, thus providing a simple solution to the representation and characterization problem.

Volterra Nonlinear Transfer Function Method with Power Spectral Density Representation of Input Signal

The Volterra nonlinear transfer function method is based upon development of a nonlinear input–output description, in contrast to a state-variable or node-voltage description, of a nonlinear system represented by current-voltage and charge-voltage frequency-dependent expressions. Wiener demonstrated that, provided a circuit is weakly nonlinear and does not possess hysteresis, an input-output description of order k defined by the k-dimensional impulse response [47]:

$$g_k(t) = \int h_1(\tau_1)x(t-\tau_1)\,d\tau_1 + \iint h_2(\tau_1,\tau_2)x(t-\tau_1)x(t-\tau_2)\,d\tau_1\,d\tau_2 + \iiint h_3(\tau_1,\tau_2,\tau_3)x(t-\tau_1)x(t-\tau_2) \times x(t-\tau_3)\,d\tau_1\,d\tau_2\,d\tau_3 + \cdots \qquad (14)$$

is possible, where the integrals are over all time. The total nonlinear response up to order l is:

$$y(t) = \sum_{k=1}^{l} g_k(t) \qquad (15)$$

where l is usually limited to 3 because of analytic complexity. For steady-state quasiperiodic analysis we can analytically Fourier transform eq. (14) to obtain the kth order Volterra nonlinear transfer function, giving a nonlinear frequency-domain input–output description. It is in this form that this method finds widest applicability in microwave engineering. Assuming quasiperiodicity allows the input excitation to be represented as a finite summation of sinusoids, so that the kth order output, $g_k(t)$, consisting of nonlinearities of up to order $k \le l$, can be expressed as:

$$g_k(t) = 2^{-k} \sum_{c1=-d}^{d} \sum_{c2=-d}^{d} \cdots \times \sum_{ck=-d}^{d} H_k(\omega_{c1}, \omega_{c2}, \ldots, \omega_{ck}) \cdot \exp[j(\omega_{c1} + \omega_{c2} + \cdots + \omega_{ck})t] \qquad (16)$$

where H_k is the kth order Volterra nonlinear transfer function and it has been assumed that each of the d sinusoids has unit amplitude [45]. Note that the summations in eq. (16) explicitly formulate all possible linear combinations up to order k of the input sinusoids. These linear combinations constitute the rectification, linear, harmonic, cross-modulation, intermodulation, desensitization, and gain enhancement/compression terms. The Volterra nonlinear transfer function simply describes the magnitude and phase of power transfer from the input to the output, much as conventional linear transfer function analysis does for linear systems.

Development of the Volterra nonlinear transfer function can be done analytically, experimentally, or numerically [36–39, 48, 49]. The method developed here is based on analytical derivation of the Volterra nonlinear transfer function of a GaAs MESFET [36]. Under the assumption of weak nonlinearity and narrow-band modulation, the method is capable of predicting the contribution to aggregate adjacent-channel power of each nonlinearity in a GaAs MESFET. Using the method of nonlinear currents closed-form expressions for the set of Volterra nonlinear transfer functions describing a GaAs MESFET equivalent circuit can be derived up to order 3. Each transfer function consists of current-voltage and capacitance-voltage expansions up to degree 3.

Representation of the digitally modulated input signal is done using eq. (5) with subsequent Fourier transformation to generate a PSD. Again the modulation rule and filtering were based on the NADC standard [5]. A maximal-length sequence of 1024 bits was used. Noting that multiplication in the time-domain maps to convolution in the frequency domain, we can characterize upper adjacent channel power directly as:

$$P_{adj,upper} = \int_{f_3}^{f_4} |H_3(f, \alpha, T_s)|^2 S(f) * S(f) * S(f)\, df \quad (17)$$

where $H_3(f, \alpha, T_s)$ is the third-order Volterra nonlinear transfer function, $S(f)$ is the NADC PSD, "$*$" denotes convolution, and the limits of integration are defined in Figure 4. A similar expression can be derived for the main-channel power assuming that it consists of linear terms only, appropriate for weak nonlinearities. Thus:

$$P_{main} = \int_{f_1}^{f_3} |H_1(f)|^2 S(f)\, df \quad (18)$$

where the limits of integration are also defined in Figure 4.

Figure 8 shows the results of the simulation to characterize spectrum regeneration due only to nonlinear transconductance; the input PSD is also shown. The simulated ACPR is −26.0 dBc. A frequency resolution of 50 Hz was obtained; in comparison, time-invariant harmonic balance achieved a resolution of 468 Hz. To remove the effect of filter overlap, the receive Nyquist filter was not used and the integration bandwidth was reduced to 30.00 kHz from 32.81 kHz. The ideal ACPR is then approximately −50 dBc instead of −36.4 dBc, allowing the low-level nonlinear effects of C_{ds} and g_{ds} to be observed. Incorporation of the filter response is trivial, however, with only the addition of the Nyquist transfer function magnitude squared response in eqs. (17) and (18).

Assessment of Frequency-Domain Methods

The primary benefit of the Volterra nonlinear transfer function method is that the input and output signals are described explicitly in the frequency-domain, obviating the need for Fourier transformation. This eliminates associated problems such as additive numerical and interpolation noise. Since the input and output signals are represented directly as a PSD, memory requirements are small compared to time-invariant harmonic balance. The length of the bit sequence is independent of the memory needed to represent the input and output PSDs; therefore, maximal-length sequences can be used without a memory penalty. Moreover, iteration is unnecessary since closed-form expressions result, facilitating a rapid solution. Finally, tone spacing, a significant problem with time-invariant harmonic balance for the characterization of ACPR, is circumvented completely with the Volterra nonlinear transfer function method.

Limitations are significant, however, as the overhead involved in deriving the Volterra non-

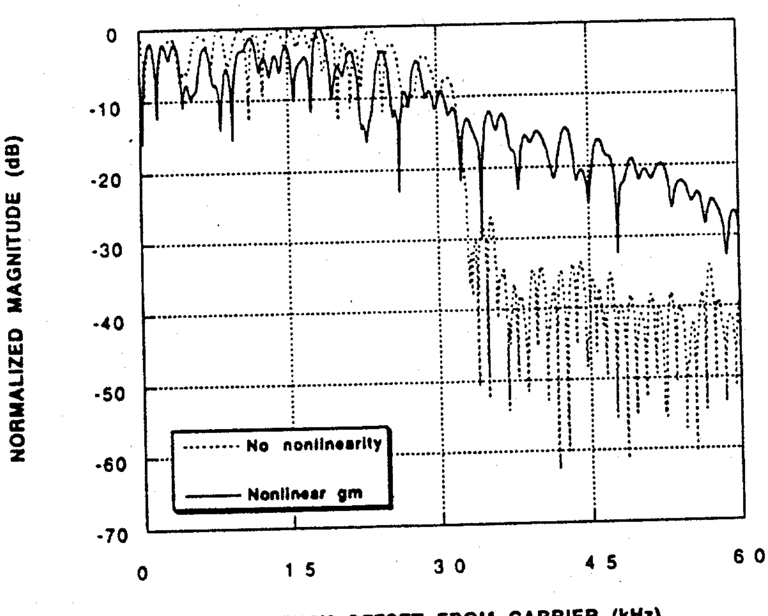

Figure 8. Simulation of GaAs MESFET spectrum regeneration based on the Volterra nonlinear transfer function method [36].

linear transfer functions, which are application specific, is considerable. As well, analysis is usually limited to order three or less, although a numerical Volterra nonlinear transfer function of order 40 was developed for simulating a diode circuit [50]. That higher order transfer functions are necessary to model even mild nonlinearities is a consequence of the relatively gradual convergence of the Volterra series representation [15]. Other frequency-domain methods, such as spectral balance, are not ideally suited for simulation of digital wireless communication systems, being frequency-domain equivalents of harmonic balance.

MEASUREMENT-BASED BEHAVIORAL MODEL METHOD

Provided certain conditions are met, measurement-based behavioral modeling methods have the potential to provide a convenient alternative to the time-variant harmonic balance method. Fortunately, these conditions are not restrictive, and in many instances it is possible to accurately describe a nonlinear system with input–output descriptions that are experimentally derived. This is in contrast to analytical behavioral modeling methods, such as the Volterra nonlinear transfer function method, which require explicit development of the input–output description. For example, steady-state characterization of a microwave power amplifier operating in compression is possible by using measured frequency-domain data to generate a nonlinear input–output description. The behavioral modeling approach has the significant advantage that very little understanding of the internal characteristics of the nonlinear system is necessary. Moreover, since feedback effects are incorporated implicitly within the input–output description, the solution is noniterative. And, although the input–output description may be either in the time domain or frequency domain, Fourier transformation is not an issue since uniform time samples are used. Finally, the method is amenable to PSD representation and characterization of a digitally modulated signal with a bit sequence of arbitrary length, thus ensuring that a maximal-length data sequence can be used.

A Measurement-Based Behavioral Model of a Microwave Power Amplifier

A measurement-based behavioral model must have an input–output description capable of accurately describing the nature of the circuit. As well, the input–output description must consist of parameters that are easily measured and/or calculated. One approach, now considered, is to represent a microwave power amplifier by a complex transfer function based on measured AM–AM and AM–PM data. Since the present method does not model the instantaneous signal, but instead the envelope, a slowly varying assumption must be made. This is usually not a restriction for most digital wireless communication systems, although certain devices, for example, GaAs MESFETs, may exhibit dispersive effects that may not be accurately modeled by the slowly varying envelope assumption.

Consider eq. (2), which upon expansion yields:

$$s(t) = m(t)\cos\phi(t)\cos(\omega_c t) - m(t)\sin\phi(t)\sin(\omega_c t) \qquad (19)$$

Making the slowly varying envelope assumption allows the energy, E, of the signal to be calculated using the low-pass equivalent representation of eq. (19) so that:

$$E = \frac{1}{2}\int_{-\infty}^{\infty} |m(t)|^2\, dt \qquad (20)$$

A consequence of this condition is that the power characteristics of a complex signal are completely described by its envelope, that is, $m(t)$. Now, the output of a low-pass equivalent complex transfer function is described completely given the envelope characteristics only of the input signal. This procedure is conceptually compared to the previous methods in Figure 9. Note that this process essentially converts a circuit description at the carrier frequency to a low-pass equivalent circuit description, and thus can be interpreted as a band-pass to low-pass mapping operation.

The present method is that implemented in OmniSys from HP-EEsof [52–54]. Consider a complex low-pass equivalent input–output description:

$$h(t) = f[m(t)] \cdot \exp\{j\theta[m(t)]\} \qquad (21)$$

where $f[m(t)]$ and $\theta[m(t)]$ are nonlinear AM–AM and AM–PM descriptions, respectively. An in-phase and quadrature time-invariant memoryless decomposition of eq. (21) is possible if it is assumed that the statistics of the resulting PSDs

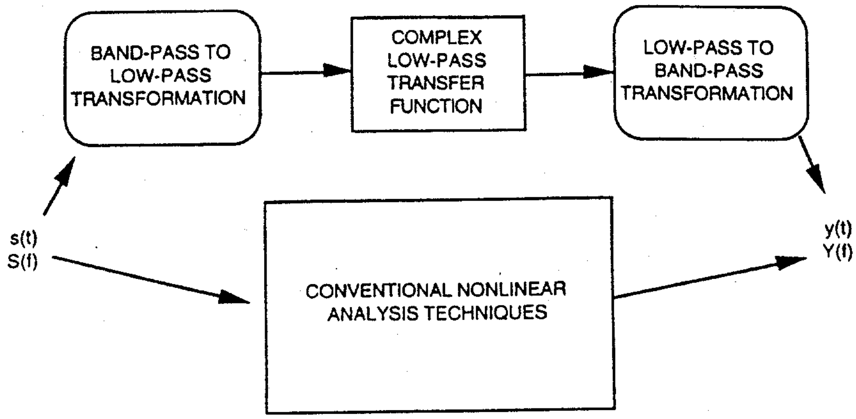

Figure 9. Comparison of conventional nonlinear analysis methods with slowly varying envelope behavioral model.

are independent. This quadrature decomposition takes the form [53, 54]:

$$f_i(t) = f[m(t)]\cos\theta[m(t)] \quad (22a)$$

$$f_q(t) = f[m(t)]\sin\theta[m(t)] \quad (22b)$$

Here, $f[m(t)]$ and $\theta[m(t)]$ are expressed as nonorthogonal polynomials of degree k:

$$f[m(t)] = a_0 + a_1 m(t) + \cdots + a_k[m(t)]^k \quad (23a)$$

$$\theta[m(t)] = b_0 + b_1 m(t) + \cdots + b_k[m(t)]^k \quad (23b)$$

where coefficients $a_0, \ldots, a_k$ and $b_0, \ldots, b_k$ are generated by a least-squares fit to measured AM–AM and AM–PM data. Note that implicit to the narrow-band assumption is that harmonics cannot be characterized, as they fall out of the region of applicability. Therefore, coefficients $a_0, \ldots, a_k$ and $b_0, \ldots, b_k$ in the OmniSys simulator are restricted to odd-degree only, with the assumption that in-band intermodulation is due only to odd-degree terms with k typically up to degree 19. (Note that, in general, this assumption is not true, i.e., an even-order nonlinearity with feedback can generate in-band IM products.)

The circuit simulated was based on a 20-mm LDMOS device, with AM–AM and AM–PM transfer characteristics measured using a Hewlett-Packard 8753B Network Analyzer. Bias was class AB with a fundamental frequency of 850 MHz with modulation following the NADC format. Figure 10 shows a plot of measured and simulated AM–AM and AM–PM data. Figure 11 shows measured and simulated ACPR as a function of input power. Reasonable agreement be-

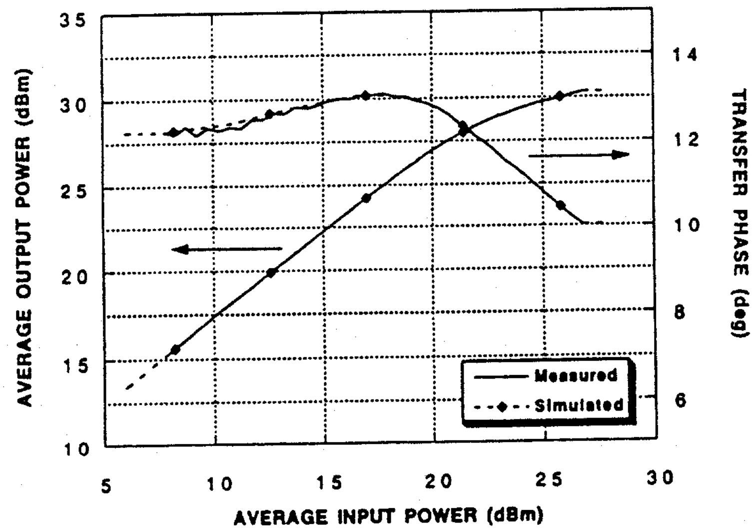

Figure 10. Measured and simulated AM–AM and AM–PM curves using OmniSys [52].

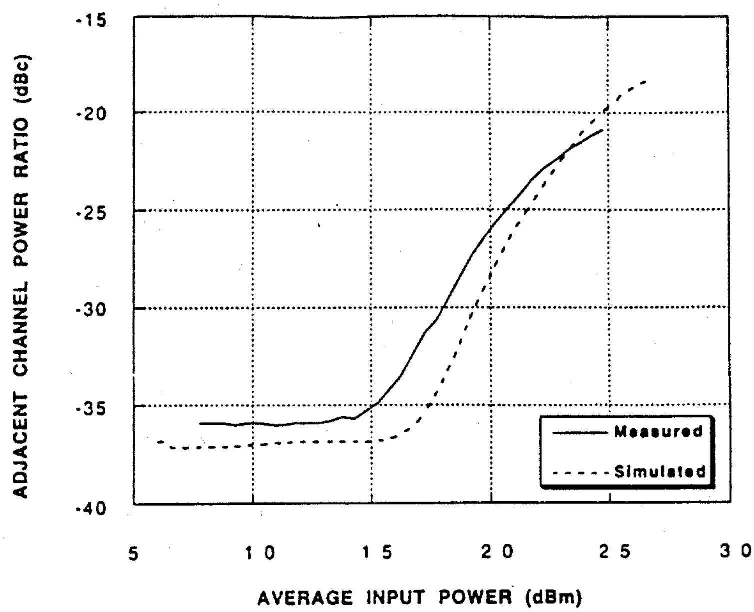

Figure 11. Measured and simulated ACPR.

tween measured and simulated ACPR was obtained over a wide dynamic range. Frequency resolution comparable to the Volterra nonlinear transfer function method was obtained. The alternate-channel power ratio was considered as well, although results were not as encouraging, with an error of over 100 dB observed. Overall, the behavioral model method works well, and is useful for quantifying ACPR of nonlinear systems.

Assessment of Behavioral Model Methods

The primary advantages of the behavioral model method lie in its rapid model extraction procedure and simple simulation method. The present method and the time-variant harmonic balance method are the only two methods available for predicting the ACPR of a microwave power amplifier operating in compression, although the former is not capable of burst analysis, simulation of source- and load-pull contours, bias-dependent analysis, or harmonic analysis. Other benefits include data sequence length independent of memory requirements, short simulation time, and freedom from tone-spacing problems. Note as well that, although Fourier transformation is necessary, uniform time samples are used ensuring excellent dynamic range.

Some limitations are evident however. The present behavioral model method is in general limited to qualitative steady-state modeling only. Moreover, simulation of harmonics is not possible. Although agreement within 3 dB for ACPR was obtained, the alternate-channel power ratio was not in good agreement over the entire dynamic range, in one instance being in excess of −100 dBc. The reasons for this are simple. Since the least squares fit is forced to work on a 19th degree nonorthogonal polynomial, there results a potential for lower-degree fitting terms not matching measured differential terms necessary for the accurate representation of ACPR. The spacing of the measurement points will tend also to have an impact on the accuracy of the fit. Simulation using nonorthogonal polynomials can be unstable, resulting in oscillatory behavior, thereby further impacting the ability to predict ACPR [55]. One method to overcome this limitation is to use a low-order complex fitting function based on a Fourier–Bessel expansion. Variants of this method have been used by two groups to predict two-tone intermodulation distortion and ACPR from AM–AM and AM–PM measurements, respectively, [56, 57]. Utilization of orthogonal functions will result in better performance over nonorthogonal polynomials of the same degree due to the basis functions better representing the nonlinear behavior of a microwave power amplifier.

DISCUSSION AND CONCLUSION

This article has presented a rigorous examination of the most basic nonlinear analysis methods in the context of simulations of digital wireless communication systems. The representation and characterization problem was outlined, with a clear description given of the differences between digitally modulated signals and analog-modulated signals. While analog-modulated signals can be represented as discrete spectra, digitally modulated signals must be represented as a power spectral density. Alternative characterization methods used to quantify system performance were described. The concept of adjacent-channel power was introduced as a metric for quantifying the linearity of microwave power amplifiers used in digital digital wireless communication systems.

Four basic nonlinear analysis methods were examined: time-domain, mixed-domain, frequency-domain, and behavioral models. Several attributes were deemed necessary to adequately address the representation and characterization problem. These were: the ability to generate a variety of complex signals with maximal-length data sequences; the ability to implement band-limiting Nyquist filters; and the ability to examine signals in the frequency domain with at least 100-dB dynamic range and resolution of 100 Hz. Additional desirable, but not necessary, attributes include the ability to perform burst-signal analysis and the ability to look at frequency-domain parameters as a function of time.

Time-domain methods, although possessing the unrivaled ability to perform transient analysis, suffer from several problems. Since adaptive time-stepping algorithms are commonly employed, interpolation is required for Fourier transformation. Consequently, dynamic ranges is limited to approximately 100 dB using state-of-the-art methods. Time-domain methods also require a large amount of memory to solve the stiff network equations for microwave circuits, a typical ratio of time-step to data sequence length being in excess of 10^6. A long simulation time is associated with this problem as well. Finally, elements amenable to frequency-domain characterization, for example, lossy transmission lines, are difficult to implement with time-domain methods. Present time-domain methods appear ill-suited to the simulation of digital digital wireless communication systems.

Mixed-domain methods overcome many of the problems inherent to time-domain methods, providing for characterization of elements and parameters in the domain in which they are best represented. The first type of harmonic balance considered assumed a solution form consisting of a finite sum of phasors—hence it is referred to as time-invariant harmonic balance. Implementation of maximal-length data sequences is prohibitive due to excessive memory and computation requirements. Consequently, only correlated data sequences can be used, resulting in an ill-conditioned Jacobian and leading to convergence problems. As well, ambiguity in the characterization of ACPR results from the coarse tone-spacings due to the necessarily short data sequence. Present forms of time-invariant harmonic balance are incapable of predicting ACPR of a microwave power amplifier in compression, although analysis of weakly nonlinear circuits may be possible.

Time-variant harmonic balance was introduced as a novel method for simulating strongly nonlinear circuits excited by digitally modulated signals. This method, based on a robust PSD approximation of a digitally modulated signal, circumvents all of the issues associated with time-invariant harmonic balance. Arbitrary PSD resolution coupled with good memory efficiency is possible. This characteristic resolves the issue of tone spacing and ACPR ambiguity due to filtering associated with time-variant harmonic balance. Moreover, this method can use any of the standard nonlinear device models that have been coded for use in time-invariant harmonic balance, thus ensuring generality. Two exclusive properties of time-variant harmonic balance are its ability to simulate ACPR load- and source-pull contours on microwave transistors and burst analysis.

Frequency-domain methods are distinguished by their ability to work strictly in the frequency-domain, thus avoiding all problems associated with Fourier transformation. Consequently, memory requirements are nearly independent of data sequence length. And, since some methods work directly with PSDs, characterization of ACPR is trivial. Dynamic ranges in excess of 400 dB is possible with frequency resolution of 100 Hz easily achieved. The Volterra nonlinear transfer function method presented demonstrated that frequency-domain methods are fast, require little memory, and provide insight that is not possible with nonlinear analysis methods requiring iteration. Although frequency-domain methods appear to work well, they are usually limited to low-order analysis, precluding simulation of ACPR of a microwave power amplifier in compression. Additionally, significant overhead is required in developing frequency-domain elements best characterized in the time-domain.

Under certain nonrestrictive conditions, a behavioral model can be used to simulate digital digital wireless communication systems. This method retains many of the benefits of frequency-domain methods, such as the ability to represent data sequences of arbitrary length, without a memory penalty, and short simulation times. Since uniform time samples are used, dynamic range in excess of 100 dB and frequency resolution better than 100 Hz are possible. Some limitations of the empirical behavioral model method include not being able to easily simulate the effect of frequency and bias variations, simulating harmonic levels, and simulating the effect of source and load changes. Inclusion of these would necessitate extraction of another behavioral model representing the new conditions, or using a time-domain model, coupled with conventional harmonic balance, to simulate AM–AM and AM–PM curves.

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BIOGRAPHIES

John F. Sevic was born in Lansing, Michigan, in 1965. John received the BS degree with honors in 1988 and the MS degree in 1989, both in electrical engineering. John is currently pursuing the PhD degree in electrical engineering, part-time, at Arizona State University. His research there focuses on investigation of alternative harmonic-balance basis functions for analysis of digitally modulated aperiodic signals. He joined Motorola Cellular Infrastructure Group, Arlington Heights, Illinois, in 1989, as an RF engineer. During that time he was engaged in the design of high-linearity RF power amplifiers for TDMA-based digital cellular infrastructure. John is currently a senior staff scientist at the Motorola Semiconductor Product Sector, Phoenix, Arizona, where he is responsible for developing linearity characterization and simulation techniques for digital wireless communication systems. He developed the first large-signal automated load-pull system for measurement of adjacent- and alternate-channel power ratio contours. He has also developed many novel techniques for characterizing and representing digitally modulated signals, including a method to analytically correlate IM and adjacent-channel power. John is a member of the IEEE MTT and ED Societies, and Eta Kappa Nu. He is the author of six publications; has given two workshops for the MTT-S, and has given a short-course on digital modulation for ARFTG. John holds two patents for his work in high-efficiency RF-power amplifier techniques. John is a registered professional engineer in the state of Illinois. (Photo not available.)

Michael B. Steer received his BE and PhD in electrical engineering from the University of Queensland, Brisbane, Australia, in 1978 and 1983, respectively. Currently he is director of the Electronics Research Laboratory and associate professor of electrical and computer engineering at North Carolina State University. He expertise in teaching and research involves circuit design methodology. From a teaching perspective he has taught courses at the sophomore through advanced graduate level in circuit design including basic circuit design, analog-integrated circuit design, RF and microwave circuit design, solid-state devices, and computer-aided circuit analysis. He teaches video-based courses on computer-aided circuit analysis and on RF and microwave circuit design which are broadcast nationally be the National Technological University. Research has been directed at developing RF and microwave design methodologies. Throughout his career this work has been closely tied to the development of microwave circuits and solving fundamental problems in both high-speed digital and microwave circuit implementations. His dissertation project focuses on the design of parametric amplifiers which use a single reactive deviced in a reflection amplifier to achieve amplification of microwave signals. The result of this work was the development of the first method capable of simulating multitone signals in a reasonably general nonlinear microwave circuit. Early years were devoted to refining this nonlinear analysis technique culminating in a general method for performing the simulation of nonlinear microwave circuits and systems in the frequency domain. This was in addition to contributions to generic microwave circuit simulation technology. An outgrowth of the microwave circuit design methodology work was the electrical performance modeling of packaging for high-speed digital circuits including the characterization of interconnects in printed circuit boards, in multichip modules and in subtenth micron-integrated circuits. Modeling is of great importance in microwave engineering and the modeling contributions at RF and microwave frequencies led to the development of behavioral models of digital drivers and receivers. Dr. Steer is the librarian of the industry-based IBIS consortium which provides a forum for developing behavioral models. A converter written by his group to automatically develop behavioral models from a SPICE netlist is being used by upwards of 50 companies and has been incorporated in several commercial computer-aided engineering programs. This work also led to the development of novel area-efficient measurement techniques. Currently his interests are in the computer-aided engineering of quasioptical power combining systems; the implementation of a two-dimensional quasioptical power combining system; high-efficiency, low-cost RF technologies for wireless applications; and computer-aided engineering of mixed digital, analog, and microwave circuits. Dr. Steer is a senior member of the Institute of Electrical and Electronic Engineers and is active in the Microwave Theory and Techniques (MTT) Society. In the MTT society he serves on the technical committees on Field Theory and on Computer Aided Design. He has organized many workshops and taught many short courses on signal integrity, wireless, and RF design. He is a 1987 Presidential Young Investigator and he has authored or coauthored more than 110 papers on topics related to RF and microwave design methodology. He has worked on projects sponsored by the National Science Foundation, the Army Research Office, SEMATECH, Air Force Office of Scientific Research, Advanced Research Projects Agency, BNR, DEC, IBM, Analog Devices, Compact Software, and Scientific Research Associates.

Anthony M. Pavio is currently the manager of Strategic Product Development for the Commercial Semiconductor Products Division of Motorola whose staff joined in 1994. Prior to Motorola (1979–1994), Dr. Pavio was the technical director of the Microwave Technology Products Division of Texas Instruments. From 1974 to 1979, he was a senior microwave design engineer at Rockwell International, and from 1972 to 1974 he worked for Raytheon Company. Dr. Pavio's design experiences includes broadband microwave components, such as mixers, oscillators, hybrids, stripline arrays, monopulse feeds, traveling-wave amplifiers, waveguide antennas, microwave and HF power amplifiers, and stripline/coaxial filters. He has published more than 60 technical papers, a text book, and holds more than 20 patents. Dr. Pavio received an MSE degree in electro-physics from the University of Connecticut in 1972 and a PhD in electrical engineering from Southern Methodist University in 1982.