PULSED-RF AND TRANSIENT ANALYSIS OF NONLINEAR MICROWAVE CIRCUITS BY HARMONIC-BALANCE TECHNIQUES

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ABSTRACT

The paper introduces a harmonic-balance approach to the simulation of nonlinear microwave circuits excited by pulsed RF and/or DC sources. The same technique also provides a transient analysis capability for circuits including passive components that can only be characterized in the frequency domain. Making use of Fourier expansions of the modulating signals, the problem is reduced to the analysis of the nonlinear circuit under multitone excitation. The resulting job is very demanding from the numerical viewpoint, and requires the harmonic-balance simulator to possess a number of advanced capabilities, which are discussed in detail. The pulsed-RF analysis of a microstrip power amplifier matched by radial stubs is presented as an example of application.

INTRODUCTION

The simulation of nonlinear microwave circuits, such as power amplifiers or mixers, operated under pulsed-RF conditions is an intriguing problem from the CAD viewpoint. In principle one could obviously resort to time-domain simulators [1], which can handle any signal waveform without restriction. However, the well-known limitations of time-domain techniques in the treatment of passive circuits virtually restrict this kind of approach to simple topologies only containing lumped elements and elementary types of distributed components. On the contrary, many practical microwave circuits contain passive integrated components which can only be characterized in the frequency domain by electromagnetic methods, especially at high frequencies (e.g., [2]). For these cases a nonlinear analysis in pulsed-RF conditions is still an open problem.

When the pulses form a periodic sequence, a pulsed-RF regime is a special form of steady-state regime. It is thus intuitive that numerical techniques explicitly aimed at steady-state analysis, such as harmonic balance (HB), should represent a possible way to do the job. The purpose of this work is to proof that this kind of analysis is, indeed, feasible making use of modern HB simulators with multitone analysis capabilities [3]. In the paper the pulsed-RF operation is treated as a particular form of quasi-periodic regime, and a suitable truncation criterion for its two-tone intermodulation spectrum is proposed. It is shown that convergence and CPU time problems can be overcome by resorting to advanced techniques such as parametric modeling and sparse-matrix methods. A convenient approximation of the modulating pulse waveform is introduced in order to keep such key parameters as the pulse rise time under control in the HB analysis. The power envelope of the signal waveforms is computed in order to simplify the visual interpretation of the numerical results.

For pulse durations long enough with respect to the RF period, the analysis approach presented in the paper also provides a direct way of performing a transient analysis by harmonic-balance methods. It is worth mentioning that in the case of lumped-element topologies that can be treated in the time domain, the results of our HB technique are consistent with those provided by classic time-domain simulators such as SPICE.

MULTITONE ANALYSIS APPROACH

Let us consider a nonlinear microwave circuit excited by an RF sinusoidal source of angular frequency ω_0 (*carrier*) modulated by a periodic signal s(t) of period $2\pi/\omega_s$. For the applications of interest in this paper, s(t) is ideally a sequence of rectangular pulses, and we shall address later the problem of finding a convenient approximation of this waveform. For the time being, we shall only assume that s(t) is defined by the Fourier expansion

$$s(t) = \sum_{p=-N}^{N} S_p \exp(jp\omega_s t) \qquad (1)$$

where $S_{-p} = S_p^*$. If the unmodulated input signal is represented by

$$v(t) = \mathrm{Re}\left[V_0 \exp(j\omega_0 t)\right] \qquad (2)$$

the modulated excitation becomes

$$u(t) = v(t)\, s(t) = \mathrm{Re}\left\{ \sum_{p=-N}^{N} V_0 S_p \exp\left[j(\omega_0 + p\omega_s)t\right] \right\} \qquad (3)$$

From a conceptual viewpoint, we can think of (3) as being the output of an ideal amplitude modulator whose two inputs are fed by (1) and (2). Thus a generic nonlinear circuit excited by (3) can be redrawn as an augmented circuit obtained by connecting the ideal modulator to the input port of the original one, excited by two periodic sources of frequencies ω_0, ω_s. In general, the analysis of a nonlinear circuit under pulsed-RF conditions can thus be treated as a two-tone intermodulation (IM) analysis problem. In the mixer case, an RF signal of the form (3) is superimposed on the local-oscillator regime, so that by a similar argument the analysis can be reduced to a three-tone IM problem.

As in any application of the harmonic-balance technique, a discrete spectrum must be assumed for all signal waveforms. This is equivalent to establishing a truncation criterion for the multidimensional Fourier expansions used to represent all time-dependent quantities. A suitable criterion for the present case is to take into account all IM products of the carrier ω_0 and of a number of near-carrier sidebands $\omega_0 + p\omega_s$ with $|p| \le P$. If a generic IM product is denoted by Ω_k, this criterion leads to the

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spectrum

$$\begin{aligned} &\Omega_k = |k\omega_0 + k_s p\omega_s| \\ &0 \le k \le M \\ &1 \le k_s \le K_s = \text{int}\left(\frac{k+M}{2}\right) \\ &|p| \le P \end{aligned} \tag{4}$$

where M and P are empirically established. A number of tests on numerical accuracy have shown that an excellent choice is to take P = N, and to set M to the highest harmonic number that has to be considered for a continuous-wave (CW) analysis of the same circuit. A reduced spectrum defined by P = N and $k_s \equiv 1$ is sufficient in many practical cases. For mixer analysis a spectrum defined according to the above criteria is repeated on each side of every local-oscillator harmonic of interest.

For computational purposes, a general-purpose HB simulator [3] has been modified to accept modulated sources besides conventional CW sources as standard excitations. The Fourier coefficients of the modulating signal are computed once for all according to the discussion presented in the next section, and are stored in the computer memory. The RF source is simply defined by means of the complex amplitude V_0 and the indication that the source is modulated. Then, according to (3), the program automatically connects in series to the RF input port 2P + 1 sinusoidal sources of complex amplitudes V_0S_p and frequencies $\omega_0 + p\omega_s$ ($|p| \le P$), respectively. At this stage the multitone harmonic-balance analysis can proceed in the usual way.

It is noteworthy that the program also allows the modulation defined by (1) to be applied to the bias sources. The two cases do not differ conceptually nor computationally, and in particular the intermodulation spectrum is always defined as above. In this case the program connects in series to the bias port of interest a DC source E_0S_0 and P sinusoidal sources of complex amplitudes $2E_0S_p$ at frequencies $p\omega_s$ ($1 \le p \le P$), where E_0 is the unmodulated bias voltage. The pulsed bias can also be offset in order to include circuits periodically switched between two different bias levels. This facility allows the simulation of a number of interesting events such as turnon transients in amplifiers and oscillators, and tuning transients in VCO's.

WAVEFORM REPRESENTATION

The most obvious approach to the approximation of the modulating waveform would be to use the truncated Fourier series for a sequence of ideal rectangular pulses. However, this solution is not acceptable for practical purposes because it does not allow the pulse rise time to be controlled, and because of the large ripples occurring in the approximated waveform. Since the circuit response may exhibit ripples produced by distortion, it is essential that any spurious fluctuations occurring in the input signal be much smaller in magnitude, otherwise the results could be misinterpreted.

A possible way of overcoming these difficulties is to make use of a non-ideal pulse waveform. The abrupt rise and fall at the beginning and at the end of the pulse are replaced by smooth transitions (e.g., of sinusoidal shape). The pulse rise and fall times (t_r, t_f) can thus be accurately defined, e.g., as the times taken by the signal envelope to rise or fall between the 10% and 90% amplitude points [4]. The other parameters relevant to the definition of the pulse shape are the pulse width τ and the pulse repetition time T. The duty cycle is defined as τ/T. The ratios t_r/T, t_f/T and the duty cycle determine the number N of harmonics required to approximate the pulse shape with a prescribed maximum ripple, and thus have a critical effect on the numerical efficiency of the simulation. For best results, the train of pulses is first Fourier expanded in the regular way, then the coefficients of (1) are numerically optimized to minimize the ripple. As an example, the sum of (1) with N = 50 is plotted in fig. 1 for a sequence of pulses having a duty cycle of 30% and rise and fall times equal to 5% of T. In this case, the maximum deviation from the exact waveform is about 8×10^{-4}.

With the spectrum defined by (4), the Fourier expansion of the current through a load resistor R has the form

$$\begin{aligned} i(t) = &\ \text{Re}\left\{\sum_{p=0}^{Q} I_{0p}\exp(jp\omega_s t)\right\} \\ &+ \text{Re}\left\{\sum_{k=1}^{M}\sum_{p=-Q}^{Q} I_{kp}\exp\left[j(k\omega_0 + p\omega_s)t\right]\right\} \end{aligned} \tag{5}$$

where the current harmonics are denoted by I_{kp}, and $Q=K_sP$. In the commonly encountered case of a modulating frequency much smaller than the carrier (e.g., $\omega_s \le 0.01\omega_0$), the quantity of primary interest is the power envelope P(t) of the load current, defined as the time average of the instantaneous power $Ri^2(t)$ in a period of the carrier. From (5) by straightforward calculations we get

$$P(t) \approx R\left[a_0^2(t) + \frac{1}{2}\sum_{k=1}^{M} a_k^2(t)\right] \tag{6}$$

where

$$\begin{aligned} a_0(t) &= \text{Re}\left[\sum_{p=0}^{Q} I_{0p}\exp(jp\omega_s t)\right] \\ a_k(t) &= \left|\sum_{p=-Q}^{Q} I_{kp}\exp(jp\omega_s t)\right| \end{aligned} \tag{7}$$

The use of (6) through (7) allows the circuit response under pulsed RF conditions to be displayed with a considerable saving of CPU time with respect to a straightforward use of (5).

REQUIREMENTS FOR A HARMONIC-BALANCE SIMULATOR

The pulsed-RF analysis outlined in the previous sections is a rather ill-conditioned job from the numerical viewpoint. The main numerical difficulties arise from the following aspects of the problem.

1) The number of spectral components to be balanced is usually large. The use of a few hundreds of frequencies is customary for an accurate representation of the signal waveforms, and this often leads to nonlinear solving systems with several thousands of scalar unknowns.

2) The problem is strongly nonlinear because of the simultaneous occurrence of heavily driven devices (such as saturated FET's, or BJT's in class-C operation) and multitone excitation.

As a consequence, the harmonic-balance simulator in use must possess a number of advanced features in order to be able to carry out successfully this kind of analysis with practically acceptable efficiency. The following requirements are of primary importance.

1) The Jacobian matrix for a Newton-iteration based solution must be computed exactly to ensure highest numerical efficiency and maximum accuracy [3].

2) An efficient use of sparse-matrix techniques is mandatory in view of the storage and factorization of Jacobian matrices with sizes of several thousands [5, 6].

3) The power-handling capabilities of the HB simulator should be brought up to the highest possible level without resorting to time-consuming techniques such as source stepping. The use of parametric modeling coupled with norm reduction represents an optimum choice in this respect [7].

Fortunately, a number of advances in harmonic-balance techniques have recently made available general-purpose HB simulators with all the required capabilities. With these programs, typical pulsed-RF analyses can be performed in a matter of minutes at the workstation level, as shown in the next section. Vectorized versions of the same programs running on supercomputers can solve very large-size problems in a few seconds.

NUMERICAL RESULTS

As an example of application, let us consider a hybrid FET amplifier having the topology schematically illustrated in fig. 2. The FET is a 600 μm device described by a modified Materka and Kacprzak model [8], and radial microstrip stubs are used both in the RF matching sections and in the bias circuit. The amplifier has a saturated power output (at the 4 dB compression level) of +25.3 dBm with 6 dB of associated gain across a 1 GHz band centered around 10 GHz. It is fed by a 10 GHz sinusoidal source modulated by a periodic sequence of rectangular pulses having a pulse repetition frequency of 10 MHz, rise and fall times of 5 ns, and a duty cycle of 30%. The modulating waveform is shown in fig. 1. The analysis is carried out with the spectrum defined by (4) with $M = 4$, $k_s \equiv 1$, and $P = 50$. The required CPU time on a SUN SPARCstation 2 is about 350 seconds in saturation, and about 200 seconds at the 1 dB compression point.

In fig. 3 the power envelopes of the input and output pulses are compared for two peak input power levels, namely, +15.5 dBm (1 dB gain compression), and +19.3 dBm (4 dB compression). Note that the group delay of the amplifier is only 0.43 ns, and thus would not be appreciable with the time scale of fig. 3. Thus, in order to check the ability of the algorithm to accurately control the phases of the harmonics, an ideal transmission line 3 m long was added at the output port of the amplifier. The figure correctly shows a 10 ns group delay between the input and output waveforms. Finally, fig. 4 shows a detail of the power envelopes with an expanded vertical scale, for the input pulse and for the output pulses at the 1 dB compression point and in saturation. The progressive degradation of the pulse waveform as the peak input power is increased is clearly evident from the figure.

Note that it would be virtually impossible to carry out the same transient analysis by time-domain techniques, because of the radial microstrip stubs for which an accurate characterization is only available in the frequency domain [9].

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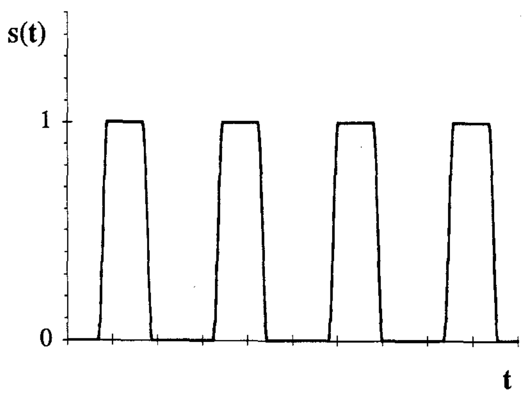

Fig. 1 Approximation of a sequence of rectangular pulses with finite rise and fall times.

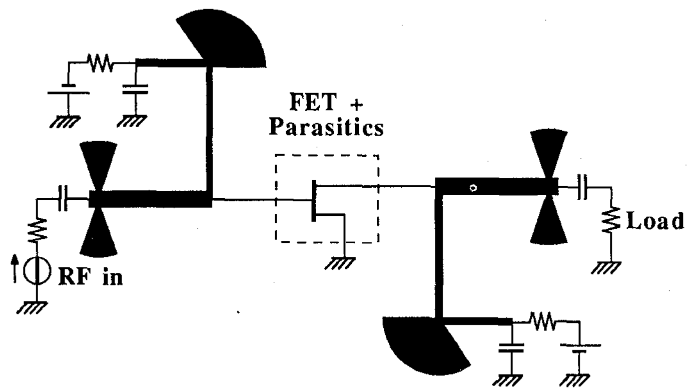

Fig. 2 Schematic topology of a microstrip power amplifier.

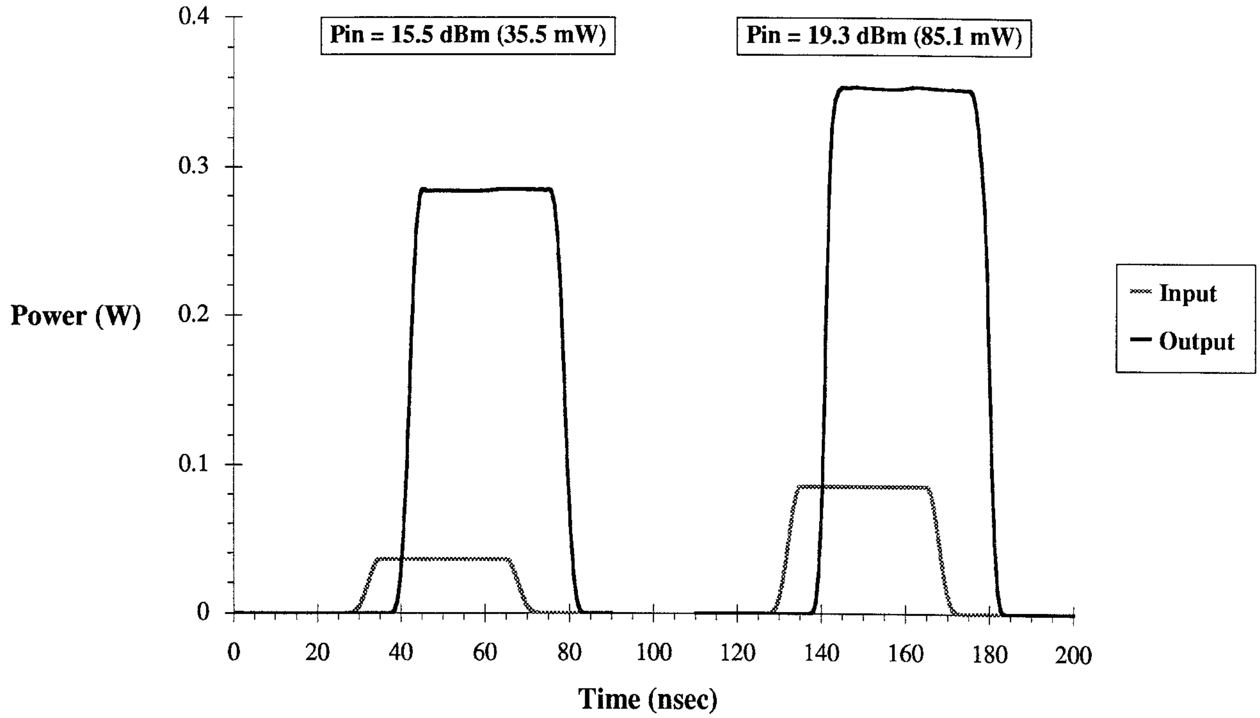

Fig. 3 Power envelopes of the input and output pulses.

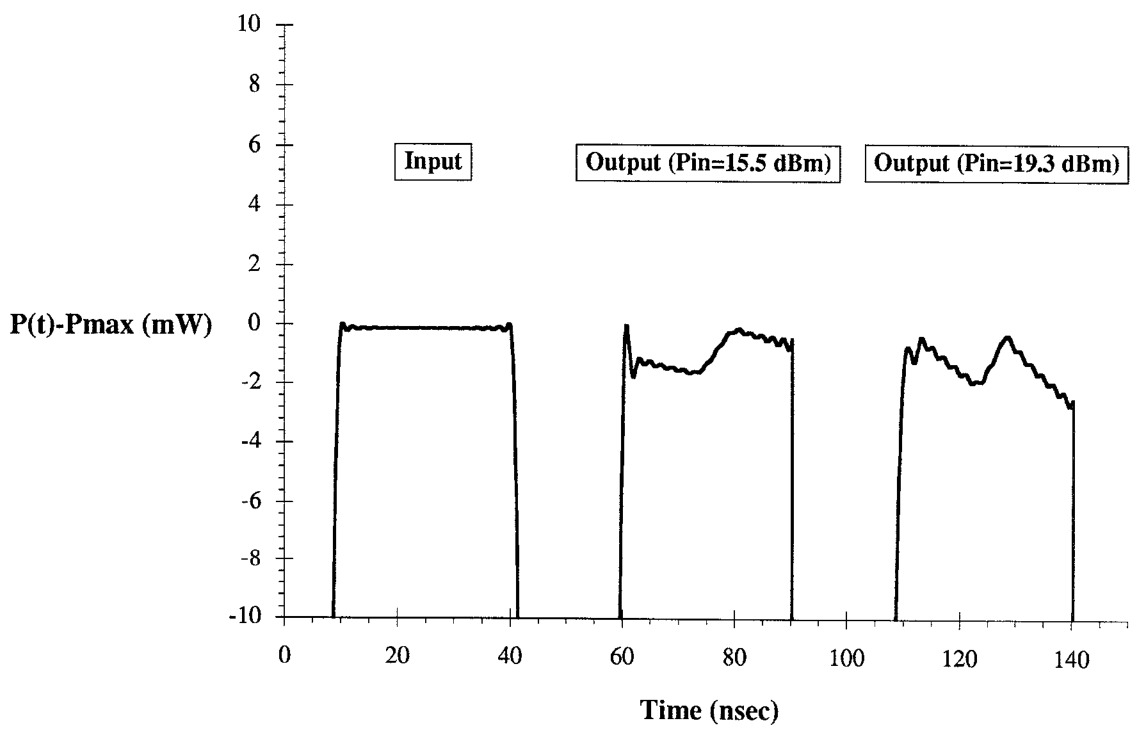

Fig. 4 Details of the power envelopes of the input and output pulses.