Numerically Efficient Design of Highly Linear Microwave Power Amplifiers

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Abstract — **A recently proposed approach for the prediction of power amplifier (PA) intermodulation distortion (IMD) on the basis of *single-tone* Harmonic Balance simulations is used in this paper for the definition of a numerically efficient design procedure of highly linear microwave power amplifiers. In particular, it is shown how standard design goals on third or fifth order intermodulation products, usually involving time-consuming two-tone HB analyses, are more simply but with sufficient accuracy taken into account by means of faster single-tone HB analyses in the iterative design process. The proposed procedure is compared with standard approaches in the design case of single-transistor and multiple paralleled-transistors MMIC amplifiers, showing dramatic reductions in computation time.**

I. INTRODUCTION

Modern broadband digital radio systems require very high-linearity microwave power amplifiers in order to adopt spectrally-efficient non-constant envelope modulation schemes. In such a context, power amplifier mild non-linearity evaluation becomes a very important issue in order to meet specifications during the design phase.

In the paper we consider the design of large-signal amplifiers with maximum power, linearity and minimum-guaranteed gain constraints. In order to take into account standard design goals on third (fifth, …) order intermodulation products, time-consuming two-tone HB analyses are needed. Since the design process is carried out by means of a non-linear optimisation search for the near-optimal complex source and load device terminations which satisfy all the design goals, a large number of two-tone analyses must be carried out on an iterative basis. This leads typically to time-consuming and computationally inefficient design processes, especially when dealing with power amplifiers using a large number of paralleled or cascaded transistors. Instead, we propose a new design procedure, based on analytical results for IMD prediction presented in [1], which can dramatically reduce the requested optimisation time by adopting single-tone HB analyses.

II. INTERMODULATION DISTORTION PREDICTION THROUGH SINGLE-TONE HB ANALYSIS

Theoretical results presented in [1] are briefly recalled here. Let us consider a two-tone sinusoidal excitation at frequencies: $\omega_1=\omega_0-\Delta\omega/2$, $\omega_2=\omega_0+\Delta\omega/2$, such that: $\Delta\omega<<\omega_0$. The instantaneous incident wave $a_{in}(t)$ at the input port of the power amplifier (scalar, 50Ω normalization) can be conveniently expressed as:

$$a_{in}(t)=\mathrm{Re}\left\{\overline{a}(t)\cdot e^{j\omega_0 t}\right\} \quad (1)$$

where:

$$\overline{a}(t)=\overline{a}_1 e^{j(\omega_1-\omega_0)\cdot t}+\overline{a}_2 e^{j(\omega_2-\omega_0)\cdot t} \quad (2)$$

represents the complex incident wave at the input port of the power amplifier. Thus, the two-tone excitation test can be seen as a modulated signal, both in amplitude and phase, having complex modulation envelope $\overline{a}(t)$. The instantaneous reflected wave at the output port of the power amplifier can be analogously written as:

$$b_{out}(t)=\mathrm{Re}\left\{\overline{b}(t)\cdot e^{j\omega_0 t}\right\} \quad (3)$$

where $\overline{b}(t)$ is the complex reflected wave at the output port of the power amplifier. Since $\Delta\omega<<\omega_0$, the complex quantities $\overline{a}(t)$ and $\overline{b}(t)$ are only slowly time-varying. We adopt a description of the power amplifier, such as:

$$\overline{b}(t)=\overline{H}\left\{\omega_0,\left|\overline{a}(t)\right|\right\}\cdot\overline{a}(t) \quad (4)$$

where the $\overline{H}\{\}$ is a complex describing function depending on ω_0 (frequency dependence in the small range: $\omega_0-\Delta\omega/2<\omega<\omega_0+\Delta\omega/2$ is neglected) and on the absolute value of $\overline{a}(t)$ since time-invariance is implicitly assumed. By considering the in-band signal components only (out-of-band signal components can be easily filtered out), the function $\overline{H}\{\cdot\}$ in (4) can be assumed to be even with respect to $\left|\overline{a}(t)\right|$, as it could be rigorously proved by means of Volterra analysis[1] [2]-[3]. Thus, we assume:

$$\begin{aligned}\overline{b}(t)&=\overline{H}\left\{\omega_0,\left|\overline{a}(t)\right|^2\right\}\cdot\overline{a}(t)=\\&=\left[H_R\left\{\omega_0,\left|\overline{a}(t)\right|^2\right\}+jH_I\left\{\omega_0,\left|\overline{a}(t)\right|^2\right\}\right]\cdot\overline{a}(t)\end{aligned} \quad (5)$$

where the real $H_R\{\cdot\}$ and imaginary part $H_I\{\cdot\}$of $\overline{H}\{\cdot\}$ have also been highlighted. By developing H_R and H_I in power series and omitting the explicit dependence on time, we obtain:

$$\begin{aligned}\overline{b}=&\Big[\Big(H_R^{SS}(\omega_0)+\alpha(\omega_0)\cdot\left|\overline{a}\right|^2+\gamma(\omega_0)\cdot\left|\overline{a}\right|^4+\ldots\Big)+\\&+j\Big(H_I^{SS}(\omega_0)+\beta(\omega_0)\cdot\left|\overline{a}\right|^2+\delta(\omega_0)\cdot\left|\overline{a}\right|^4+\ldots\Big)\Big]\cdot\overline{a}\end{aligned} \quad (6)$$

[1] On the other hand, in case of not-even dependence in (4), no small signal solution would be definable.

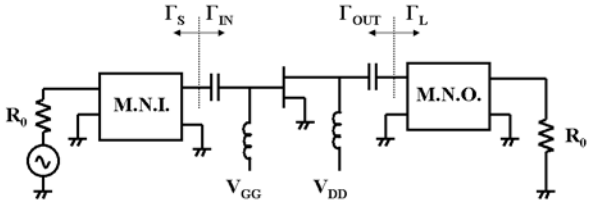

Fig. 1. Single-transistor PA. Ideal bias- and passive lossless matching- networks may be conveniently considered during the early phase of the design process.

where H_R^{SS} , H_I^{SS} are the real and imaginary parts of the small-signal component of $\overline{H}\{\cdot\}$, and α, β, γ, δ, … are suitable scalar polynomial coefficients to be determined. By substituting the complex incident wave (2) into (6), after simple algebraic manipulation, the reflected wave at the output port of the power amplifier can be written as:

$$\begin{aligned}\overline{b} &= \overline{A}\cdot e^{j(\omega_1-\omega_0)\cdot t} + \overline{B}\cdot e^{j(\omega_2-\omega_0)\cdot t} + \\ &+\overline{C}\cdot e^{j(2\omega_1-\omega_2-\omega_0)\cdot t} + \overline{D}\cdot e^{j(2\omega_2-\omega_1-\omega_0)\cdot t} + ...\end{aligned} \tag{7}$$

where $\overline{A}$, $\overline{B}$, $\overline{C}$, $\overline{D}$ are complex quantities depending on H_R^{SS}, H_I^{SS}, on the polynomial coefficients α, β, γ, δ,… and on the input signal levels $\overline{a}_1$, $\overline{a}_2$. By considering two tone having the same amplitude, i.e., $|\overline{a}_1| = |\overline{a}_2| = \hat{a}$, and assuming: $\overline{a}_1 = \hat{a}$ (zero phase, without loss of generality); $\overline{a}_2 = \hat{a}\cdot e^{j\angle\overline{a}_2}$, in the simplest case of a second-order series expansion we obtain [1]:

$$\begin{aligned}|\overline{A}| &= |\overline{B}| \cong \hat{a}\cdot\left|H_R^{SS} + jH_I^{SS}\right| \\ |\overline{C}| &= |\overline{D}| = \hat{a}^3\cdot|\alpha + j\beta|\end{aligned} \tag{8}$$

where the approximation in the first of (8) holds for sufficiently low signal levels such as those involved in IP3 evaluation tests. From (8) it is straightforward to derive the power of the spectral components at the output of the power amplifier. In particular:

$$\begin{aligned}P_1 &= \frac{1}{2}\cdot|\overline{A}|^2 = \frac{1}{2}\hat{a}^2\cdot\left[(H_R{}^{SS})^2 + (H_I{}^{SS})^2\right] \\ P_3 &= \frac{1}{2}\cdot|\overline{C}|^2 = \frac{1}{2}\hat{a}^6\cdot\left(\alpha^2+\beta^2\right)\end{aligned} \tag{9}$$

represent the power associated with the fundamental and the third-order IM products. Thus, the prediction of IP3 can be carried out, by means of (9), through the well-known relationship:

$$IP3^{[dB]} = \frac{3P_1^{[dBm]} - P_3^{[dBm]}}{2}, \tag{10}$$

provided that the two scalar coefficient α and β are first identified.

To this aim, a power-swept single-tone Harmonic-Balance simulation can be carried out, provided that a suitably accurate dynamic model of the transistor under mild non-linear operation is available. Simple analytical minimization of the square discrepancies between the power expansion (6) of the describing function $\overline{H}\{\cdot\}$, possibly truncated to the second-order, and the analogous single-tone HB-simulated function provides simple and reliable identification of the α, β coefficients through simple closed-form analytical expressions. In particular, in the case of second-order series truncation, only two single-tone HB at suitably selected input power levels are sufficient for the coefficients identification, as will be discussed in the next section.

Finally, by also taking into account higher-order terms in the polynomial expansion in (6), even more accurate identification of the low-order coefficients is obtained. Moreover, higher order IM products can also be evaluated in this case. For instance, by taking into account the fourth-order terms in (6) the fifth-order IM products (and IP5) could also be predicted.

III. Highly Linear Microwave Power Amplifier Design

The above outlined theoretical results are used here in order to define a numerically efficient optimisation loop for the design of highly linear microwave power amplifiers. Let us now first consider the new design procedure in the case of the single-transistor PA shown in Fig.1.

As outlined in the flow-chart of Fig. 2, the proposed procedure is based on an optimisation loop in the domain of the Γ_S, Γ_L terminations at the operation frequency f_0, according to conventional design procedures. In particular, near-optimal terminations are sought for with the aim of achieving maximum third order intercept point (IP3), provided that a minimum transducer power gain

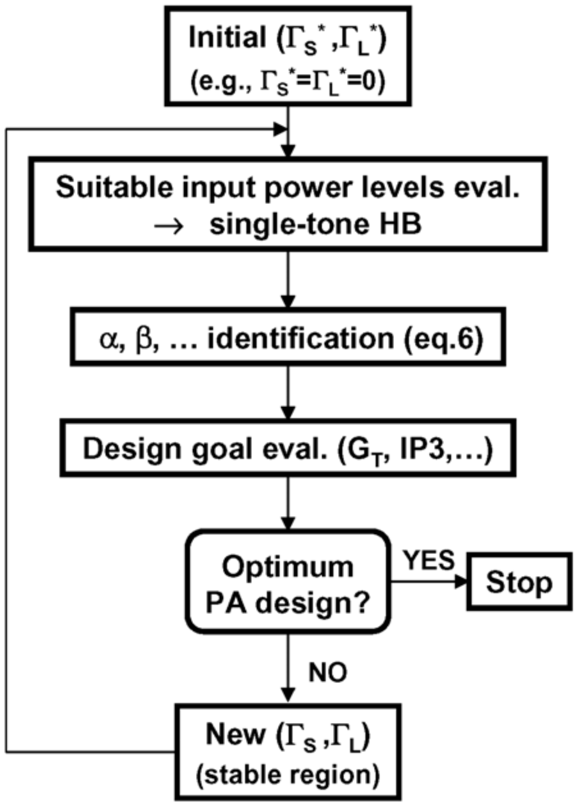

Fig. 2. Flow chart of the proposed PA design procedure for highly linear microwave PAs.

constraint is satisfied[2] ($G_T > G_{Tmin}$). Moreover, in order to guarantee stability of the amplifier stage, the search domain for Γ_S, Γ_L is bounded by values giving: $|\Gamma_{IN}|<1$; $|\Gamma_{OUT}|<1$ at f_0 (at different frequencies stability is guaranteed by 50Ω terminations , as discussed later in this section).

The real novelty of the proposed procedure over other conventional design approaches is that, instead of using two-tone analyses in the core of the design loop, a much more numerically efficient single-tone HB is adopted here for the IP3 goal evaluation through (9)-(10). In particular, the identification of the α, β, … coefficients is carried out at each iteration loop (i.e., at each guess of Γ_S, Γ_L) through closed-form analytical expressions, providing the best fit of (6) to single-tone HB results carried out at suitably selected levels of input available power.

The criterion for the choice of these power levels must be carefully chosen, since the accuracy in predicting the IP3 by means of (9)-(10) relies on a proper selection of the input power range (i.e., $|\overline{a}|^2$), where the describing function $\overline{H}\{\cdot\}$ may be well approximated by means of the power series expansion (6) with assigned order. In our examples we always adopted a truncation of the series to second order terms, obtaining greater numerical efficiency in the design procedure (only two single-tone HB required at suitable input power levels for each loop iteration) without excessive loss in the IP3 prediction accuracy.

The two power levels used for the identification of the α and β coefficients may be chosen as: 1) a very low non-critical input power level, corresponding to a practically-linear behaviour of the amplifier; 2) an "upper" power level (i.e., larger and involving non-linear operation), which depends on the actual values of the source and load coefficients Γ_S, Γ_L. Since the "upper" power level must be updated at every design loop cycle, a simplified numerically-efficient criterion should be adopted for its evaluation.

To this aim, at the very beginning of the design procedure (i.e., before starting the loop outlined in Fig.2), the IP3 corresponding to the initial guess Γ_S^*, Γ_L^* of the source and load terminations is evaluated by means of a two-tone Harmonic Balance simulation at very low input power levels. Then, the "upper" power level corresponding to Γ_S^*, Γ_L^* is assumed as the maximum input power level P_{IN}^* (corresponding to an output power P_{OUT}^*), which still provides accurate single-tone-HB-based prediction of IP3 through (9)-(10) (i.e., IP3 practically coincident with the corresponding value based on two-tone-HB). Moreover, since this "upper" power is dependent on the particular choice of Γ_S, Γ_L, the value is updated at every loop iteration by assuming: $P_{IN}(\Gamma_S,\Gamma_L)=P_{OUT}^*/G_T(\Gamma_S,\Gamma_L)$, where G_T represents the small-signal transducer gain easily evaluated by means of well-known closed-form equations [4].

[2] Since: $2\cdot P_1 = 2\cdot IP3 + D$, where $D = P_3 - P_1$, maximisation of IP3 provides minimum distortion at any requested output power or, equivalently, maximum output power at any acceptable distortion.

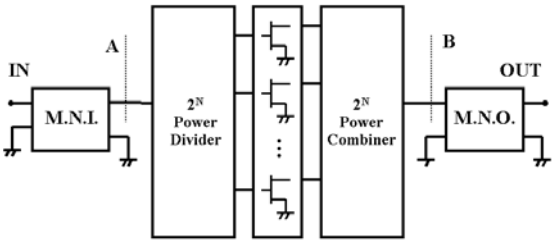

Fig. 3. Paralleled multi-transistor PA. The proposed design procedure allows for dramatic reduction in design time in this case.

This assumption is justified by considering that, due to the mild non-linearity involved, the small-signal transducer gain is not too far from the actual value under large-signal operation. Moreover, since the design procedure leads to values of Γ_S, Γ_L corresponding to near-maximum IP3, lower levels of gain compression are obtained at equals output power levels.

Ideal passive lossless bias- and matching- networks are assumed in the first stage of the proposed design procedure. These implement the required Γ_S, Γ_L reflection coefficients at the fundamental frequency f_0, while, as it happens in conventional active load-pull systems, (R_0 =) 50Ω terminations are provided at any out-of-band harmonic frequency. The implementation of such ideal networks is straightforward in modern commercial CAD tools, by means of black-box linear two-ports components, where each scattering-matrix parameter may be defined, either analytically or on a look-up-table basis, as a function of frequency.

An advantage of using ideal bias- and matching-networks in the early stage of the amplifier design process is that the particular topology of the actual networks is not of concern at all. On the other hand, actual topologies must be eventually considered in order to realistically take into account also the (minor) contributions to third-order IM products of the network impedances at out-of-band frequencies: $\Delta f = f_2 - f_1$, $2f_1$, $2f_2$, … [5]-[6]. To this aim, the near optimum Γ_S, Γ_L terminations identified through the proposed simplified may be eventually considered as a "good" initial guess for final circuit optimisation, where ideal bias- and matching- networks are replaced with actual ones and a two-tone HB analysis is used.

The outlined design procedure may be efficiently exploited especially in the design of multi-transistor PAs, such as in the case of distributed amplifiers or the paralleled 2^N-transistor PA final stage shown in Fig.3. In this case, after designing the power dividing/combining networks according to minimum space requirements and other possible constraints [7], the design procedure may be simply applied at sections A and B shown in the figure. Dramatic reduction in simulation times can be obtained in this case, as shown in the next section.

IV. Design Procedure Validation

The evaluation of the IP3 prediction accuracy through (9)-(10) was first considered. To this aim, all the simulation setups needed for the identification of the α

	New design procedure (1-tone HB)		Conventional design (2-tone HB)	
	PA 1x	**PA 8x**	**PA 1x**	**PA 8x**
IP3 [dBm]	48.6	52.0	48.5	52.5
G_T (small signal) [dB]	15.9	15.9	15.9	15.9
$P_{OUT}@G_{T1dB}$ [dBm]	23.7	31.8	23.8	31.9
Optimum Γ_S	0.445 /-3.9°	0.689 /-142.9°	0.494 /-3.5°	0.551 /-163.5°
Optimum Γ_L	0.242 /-165.9°	0.841 /-154.9°	0.248 /-171.4°	0.867/-155.1°
Design time [s]	50	230	336	13738

Tab. I. Comparison between new and conventional power amplifier design procedures. As expected, almost the same optimum amplifier performance can be obtained with dramatic reduction in computation times.

	Envelope Simulation			Single Tone HB Eqs. (9)-(10)	Two-Tone HB
	T_{STEP}=100ns T_{STOP}=120µs	T_{STEP}=110ns T_{STOP}=120µs	T_{STEP}=100ns T_{STOP}=110µs		
IP3	48.2 dBm (~ok)	32.4 dBm (!!)	47.5 dBm (!)	48.6 dBm	48.6 dBm
Sim. time (norm.)	100%	93%	92%	23%	37%

Tab. II. Intermodulation prediction by using (9)-(10) in comparison with other simulation methods (two-tone HB and envelope) with the optimum load and source impedances: Γ_S = 0.445 /-3.9°, Γ_L = 0.242 /-165.9°. The accuracy of envelope simulations is strongly dependent on "tuning" of the T_{STEP} and T_{STOP} parameters.

and β coefficients were implemented in the framework of the Agilent ADS CAD tool for microwave circuit design. A 0.15µm PHEMT (W = 600 µm) in MMIC technology was chosen for all the validations. Moreover, the transistor was described by means if an Agilent EEHEMT non-linear dynamic foundry model.

In order to quantify the achievable prediction accuracy, IP3 values computed on the basis of single-tone HB analyses through (9)-(10) for different source and load terminations were compared with corresponding values based on two-tone HB analyses (2 MHz tone spacing). The IP3 prediction accuracy was always better than 0.3 dB.

The validation of the new PA design procedure was then addressed. To this aim, the design flow outlined in Fig.2 was also fully implemented in the CAD tool, where both a single (1x) and a paralleled eight-transistors (8x) class-A, MMIC power amplifiers were designed using the proposed procedure at the frequency of 5 GHz. The obtained results were then compared, with respect to circuit performance and simulation time, to conventional designs based on two-tone HB analyses (i.e., designs based on the same flow-chart in Fig.2, but using two-tone HB analyses for the evaluation of the IP3 goal).

The analysis of the comparison data presented in Tab.1, clearly shows that the same near-optimum amplifier performance can be obtained by adopting the new procedure with dramatic reduction in the simulation time, especially in the case of multi-transistor circuits. In particular, the (8x) PA design time by using the new procedure only takes a few minutes, while many hours are required with conventional approaches.

Finally, a test was made in order to evaluate the IP3 computation times by means of envelope simulation. Comparative results are shown in Tab. II for the (1x) PA adopting the near-optimum source and load terminations. These results show that, at mild non-linearity envelope simulation is even slower than two-tone analysis and it also requires quite critical adjustment of the T_{STEP} and T_{STOP} simulation parameters.

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