Evaluating Co-Channel Distortion Ratio in Microwave Power Amplifiers

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***Abstract*—Laboratory results, obtained with a novel setup for a corrected co-channel distortion ratio, validate the idea that no matter the notch width, a conventional noise-power-ratio test produces optimistic small-signal in-band distortion measurements, when compared to a hypothetical continuous spectrum excitation test by the authors. This paper also generalizes previous memoryless mildly nonlinear behavior predictions to saturated and frequency-dependent regimes. Finally, a close agreement between measurement results and harmonic-balance simulated data provided an alternative means of corrected co-channel power-ratio evaluation.**

***Index Terms*—Amplifier distortion, bridge circuits, design automation, measurement.**

I. Introduction

MULTITONE intermodulation distortion ratio (M-IMR), adjacent channel power ratio (ACPR), and noise power ratio (NPR) evaluations are progressively replacing the ancient two-tone intermodulation (IMD) tests since they provide more convenient ways for in-band distortion characterization. This is a consequence of the fact that a multitone or band-limited noise is a much better representation of a general telecommunications signal than the two-tone stereotype. Also, since we are dealing with general nonlinear systems, the estimation of distortion performance is as much useful as their excitation in the test is closer to the ones expected in normal operation.

In this way, M-IMR and ACPR became interesting figures to measure the ratio between signal output power and sideband (or adjacent channel) distortion power, while the NPR has been accepted for many years as the standard for signal to in-band (or co-channel) distortion ratio evaluation.

However, in two recently published papers [1], [2], the authors have theoretically demonstrated the counterintuitive idea that no matter the NPR notch width, it will produce a nonnegligible impact on small-signal in-band distortion. Indeed, they predicted that the usual NPR figure-of-merit gives an optimistic misjudgment of 5.6 dB on those distortion components if the notch is located at the middle of the input bandwidth or up to 7 dB if co-channel distortion is to be tested at the bandwidth extremes. This implies that a corrected in-band distortion evaluation would impose an increase in output power backoff of a power amplifier (PA) by near 3 dB, for the same co-channel power ratio (CCPR) specifications.

Unfortunately, these interesting conclusions were derived from theoretical considerations only valid for unsaturated PAs or, more precisely, for general third-order nonlinear memoryless systems [2], driven by equal amplitude multitone signals. Therefore, any attempt to validate them experimentally under real conditions is worth the effort, as it would give a first estimate of the impact of the frequency independence, quasi-linearity, and driving spectrum flatness approximations involved, and lead to the proposal of a corrected CCPR evaluation setup.

The main objective of this paper is to address these issues. For that, a practical distortion measurement setup useful for corrected CCPR evaluation in microwave PAs will first be reviewed [3]. Distortion data obtained with these experiments will then be compared to the theoretical predictions and harmonic-balance (HB) simulation results in order to discuss the validity of the approximations made.

II. CCPR Experimental Setup

This section has a twofold purpose. First, a previously proposed CCPR measurement setup, appropriate to characterize the distortion of general bandpass nonlinear systems [3], will be reviewed. Second, the available accuracy and measurement bandwidth provided by that laboratory arrangement is discussed.

Since, in microwave and RF fields, the PA is the best illustration of this class of systems, in the following, we will take this part for the whole class, without loss of generality.

A. Proposed CCPR Setup Revision

Since conventional NPR tests require the elimination of the input signal from the observation bandwidth, they wipe out all signal correlated components [1], [2]. However, the evident solution of filling in the notch cannot also be accepted, as it would obviate the desired in-band distortion observation. Thus, the whole fundamental components are needed at the input to permit the generation of all possible nonlinear mixing products, but are undesirable at the output, where they behave as a dominant perturbation to the sought IMD signals.

To solve that dilemma, we suggested a distortion measurement setup [3], which is based on a bridge arrangement [4] or feedforward signal cancellation loop [5]. There, the device under test (DUT) is excited with its full input spectrum, while the linear components are then canceled out from the output,

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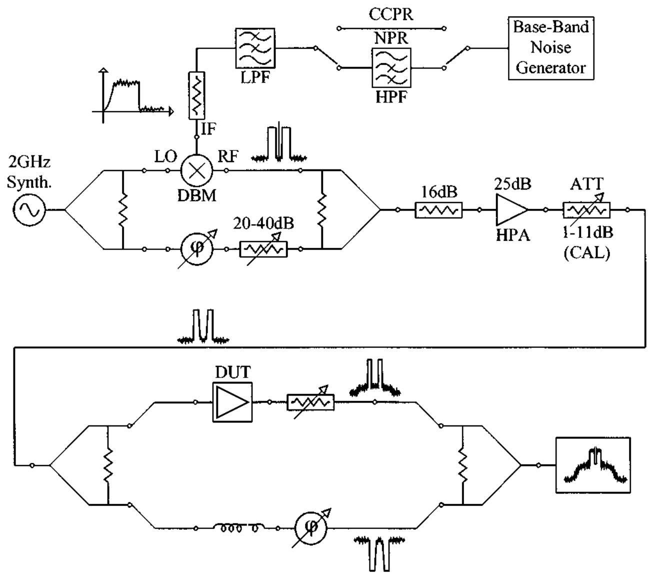

Fig. 1. Functional diagram of proposed distortion measurement setup [3].

using a scaled version of the stimulus provided by the bridge linear auxiliary arm.

For paper readability, a copy of that measurement setup is shown in Fig. 1.

This setup has at least three advantages for supporting it. The first one is that it is sufficiently simple to be built with components that are trivial in any microwave laboratory. The second is that the cancellation of the strong linear DUT's output components also prevent any possible distortion induced by the spectrum analyzer [4]. Finally, the third is that this arrangement has already been fully tested in all feedforward PA linearizers [5].

The similarity of this bridge to the signal cancellation loops included in feedforward linearizers is far from being accidental. In fact, a feedforward linearizer operates by first generating an error signal in a signal cancellation loop to afterwards correct the PA output in the distortion cancellation loop. Therefore, this error signal must be a replica of the amplifier-induced noise-plus-distortion perturbation. Thus, a visualization of this error signal constitutes a true distortion (plus noise) measurement system.

B. Available CCPR Measurement Accuracy

It should be noted that the setup of [3] really assumes that the DUT is an amplifier. If this is not the case, i.e., if the DUT does not contribute with a net power gain, but instead attenuates the signal, the variable attenuator present in the DUT's arm should be transferred to the auxiliary arm.

Perfect bridge balance requires that linear frequency characteristics of the DUT $\mathrm{H}_1(\omega)$ are such that it imposes a constant gain within the operation bandwidth, and that its phase lag is simply proportional to frequency. Although this can always be guaranteed close enough to the central frequency, it may not be met if the frequency deviates from this point. In those cases, a perfect linear components' cancellation cannot be achieved in the whole bandwidth, causing a degradation in the IMD measurement accuracy.

Since, in a normal microwave DUT, IMD power versus drive is expected to increase at a larger rate than fundamental output power, this accuracy impairment will only be perceptible at higher CCPRs. The limitations inherent to this setup will then be evident for quasi-linear DUTs and not for strong nonlinear ones. Due to the third-order nature of the IMD signals present at the output of these DUTs, and the first-order nature of the residual linear components, it can be expected that the distortion to perturbation ratio increases 2 dB per decibel of excitation level rise.

With these considerations in mind, an estimate of the maximum CCPR that may be measured with a certain prescribed error margin can be calculated by expressing the residual bridge signal amplitude as a function of the unbalances in phase and amplitude between the main and auxiliary bridge arms.

If the main and auxiliary arms have linear transfer functions of the form

$$\mathrm{H_m}(\omega) = \rho_\mathrm{m}(\omega)\mathrm{e}^{\mathrm{j}\phi\mathrm{m}(\omega)} \tag{1}$$

and

$$\mathrm{H_a}(\omega) = \rho_\mathrm{a}(\omega)\mathrm{e}^{\mathrm{j}\phi\mathrm{a}(\omega)} \tag{2}$$

respectively, then the uncanceled portion of the fundamental signal present at the bridge output will be proportional to

$$|\mathrm{H_b}(\omega)| = |\mathrm{H_m}(\omega) - \mathrm{H_a}(\omega)|. \tag{3}$$

At ω_0, the frequency at which the bridge was adjusted, $\mathrm{H_b}(\omega_0) = 0$. However, as ω deviates from this central value, the cancellation will be degraded, and a progressively higher residual fundamental will perturb co-channel IMD components. For evaluating the impact of this error signal on the measured data, we will assume that ω will suffer small deviations from ω_0, $\omega = \omega_0 + \Delta\omega$, so that first-order Taylor expansions of $\mathrm{H_m}(\omega)$ and $\mathrm{H_a}(\omega)$ are valid as follows:

$$\begin{aligned} H(\omega) &\approx H(\omega_0) + \frac{\delta H(\omega)}{\delta\rho(\omega)}\frac{\delta\rho(\omega)}{\delta\omega}\bigg|_{\omega_0}\Delta\omega + \frac{\delta H(\omega)}{\delta\phi(\omega)}\frac{\delta\phi(\omega)}{\delta\omega}\bigg|_{\omega_0}\Delta\omega \\ &= H(\omega_0)\left[1+\frac{1}{\rho(\omega_0)}\frac{\delta\rho(\omega)}{\delta\omega}\bigg|_{\omega_0}\Delta\omega + j\,\frac{\delta\phi(\omega)}{\delta\omega}\bigg|_{\omega_0}\Delta\omega\right] \end{aligned} \tag{4}$$

and $|\mathrm{H_b}(\omega)|$ becomes

$$\begin{aligned} &|H_b(\Delta\omega)| \\ &\quad= |H(\omega_0)|\left|\frac{1}{\rho_m(\omega_0)}\frac{\delta\rho_m(\omega)}{\delta\omega}\bigg|_{\omega_0} - \frac{1}{\rho_a(\omega_0)}\frac{\delta\rho_a(\omega)}{\delta\omega}\bigg|_{\omega_0}\right. \\ &\qquad\left. + j\left[\frac{\delta\phi_m(\omega)}{\delta\omega}\bigg|_{\omega_0} - \frac{\delta\phi_a(\omega)}{\delta\omega}\bigg|_{\omega_0}\right]\right| |\Delta\omega|. \end{aligned} \tag{5}$$

For our distortion bridge arrangement, where the auxiliary arm can approximately be modeled as a constant attenuation and a delay τ_L plus a constant phase φ, we have

$$\mathrm{H_a}(\omega) = \rho_\mathrm{m}(\omega_0)\mathrm{e}^{\mathrm{j}(\varphi - \omega\tau_\mathrm{L})} \tag{6}$$

where τ_{L} must be adjusted so that $\varphi - \omega_0\tau_{\mathrm{L}} = \phi_{\mathrm{m}}(\omega_0)$ and

$$|H_b(\Delta\omega)| = \rho_m(\omega_0)\left|\frac{1}{\rho_m(\omega_0)}\left.\frac{\delta\rho_m(\omega)}{\delta\omega}\right|_{\omega_0} + j\left[\left.\frac{\delta\phi_m(\omega)}{\delta\omega}\right|_{\omega_0} + \tau_L\right]\right||\Delta\omega|. \tag{7}$$

After balance, the bridge output fundamental signal power will be given by

$$|Y_1(\omega)|^2 = \frac{1}{2}|H_b(\Delta\omega)|^2|X(\omega)|^2 \tag{8}$$

where $|X(\omega)|^2$ stands for the bridge's total input signal level, while the intended IMD output power at the same frequency will be

$$|Y_3(\omega)|^2 = \frac{1}{2^3}\rho_m(\omega_0)^2\frac{|H_3(\omega,\omega_x,-\omega_x)|^2|X(\omega)|^6}{|H_1(\omega_0)|^2} \tag{9}$$

which corresponds to a measurement signal to perturbation ratio (SNR) of

$$\mathrm{SNR} \equiv \frac{|Y_3(\omega)|^2}{|Y_1(\omega)|^2} = \frac{1}{2^2}\rho_m(\omega_0)^2\frac{|H_3(\omega,\omega_x,-\omega_x)|^2}{|H_1(\omega_0)|^2|H_b(\Delta\omega)|^2}|X(\omega)|^4 \tag{10}$$

and $\mathrm{H}_3(\omega,\omega_x,-\omega_x)$ stands for the third-order nonlinear transfer function of the corresponding Volterra-series model.

If $|\mathrm{H}_1(\omega)|$ can be considered approximately constant in the vicinity of ω_0, then (10) can be written as

$$\mathrm{SNR} = \frac{1}{4}\frac{1}{\mathrm{CCPR}}\frac{\rho_m(\omega_0)^2}{|H_b(\Delta\omega)|^2} \tag{11}$$

which gives the setup CCPR measurement capability for a desired SNR accuracy, shown in (12), at the bottom of this page. Just to illustrate this accuracy evaluation, let us suppose we would like to determine the relative bandwidth in which CCPR measurements of at least 40 dB can be made, with no more than 2-dB error.

For this error margin

$$10\log\left[\frac{|Y_3(\omega)|^2}{|Y_3(\omega)+Y_1(\omega)|^2}\right] \le 2\text{ dB} \tag{13}$$

the ratio between desired IMD signals and residual fundamental power must be SNR$\geq$13.7 dB.

If the DUT can be modeled as an amplifier with a simple second-order bandpass network with a quality factor of $\mathrm{Q}=5$ plus an inversion, and if the phase unbalance can be assumed to dominate $|\mathrm{H_b}(\Delta\omega)|$, then

$$\left.\frac{\delta\phi_m(\omega)}{\delta\omega}\right|_{\omega_0} = -\frac{2Q}{\omega_0}. \tag{14}$$

As $\phi_{\mathrm{m}}(\omega_0)$ is π, this phase shift can be reproduced in the auxiliary arm by an odd number of half-wavelengths or

$$\tau_L = \frac{(2k+1)\pi}{\omega_0}. \tag{15}$$

If the best value of k is selected, i.e., $k=1$,

$$|H_b(\Delta\omega)| = 0.58\frac{|\Delta\omega|}{\omega_0} \tag{16}$$

which results in a maximum relative bandwidth $2|\Delta\omega|/\omega_0$ of nearly 0.4%.

This frequency-deviation value is representative of the relatively narrow measurement bandwidths offered by this type of signal cancellation loop. In the present case, the bandwidth is quite sensitive to the desired accuracy and maximum CCPR value and, thus, it can be anticipated that the setup has its small-signal CCPR application domain restricted to relatively narrow-bandwidth amplifiers. Also, some care is required for the design of the auxiliary arm so that it matches as close as possible the DUT's frequency characteristics. Setups intended for DUTs with significantly different (or temperature variant) linear characteristics can be improved using some sort of adaptive feedforward loop [6].

III. Small-Signal NPR and CCPR Experimental Results

For illustrating the use of this CCPR setup, a MESFET-based C-band amplifier was used as the DUT to be tested.

In optimized conditions, the implemented setup was capable of more than 55-dB cancellation within the tested 500-kHz channel bandwidths. That allowed easy CCPR measurements as high as 45 dBc.

Fig. 2 shows measurement results of a conventional NPR test obtained with the above presented setup when the DUT is driven in its small-signal regime. For comparison purposes, the figure also includes corresponding HB simulations performed with an in-house developed HB simulator [7], [8].

The set of traces of highest power correspond to the measured and simulated DUT's output, obtained for a driving level 19 dB higher than the one used for loop balance (Pin_0). Thus, they are the response that should be expected from a conventional NPR setup.

The lowest amplitude traces correspond to the measured and simulated bridge output after linear components' cancellation. Therefore, they constitute the wanted distortion.

$$\mathrm{CCPR_{Max}} = \frac{1}{4}\frac{\rho_m(\omega_0)^2}{\mathrm{SNR_{min}}\left|\frac{1}{\rho_m(\omega_0)}\left.\frac{\delta\rho_m(\omega)}{\delta\omega}\right|_{\omega_0} + j\left[\left.\frac{\delta\phi_m(\omega)}{\delta\omega}\right|_{\omega_0} + \tau_L\right]\right|^2|\Delta\omega|^2} \tag{12}$$

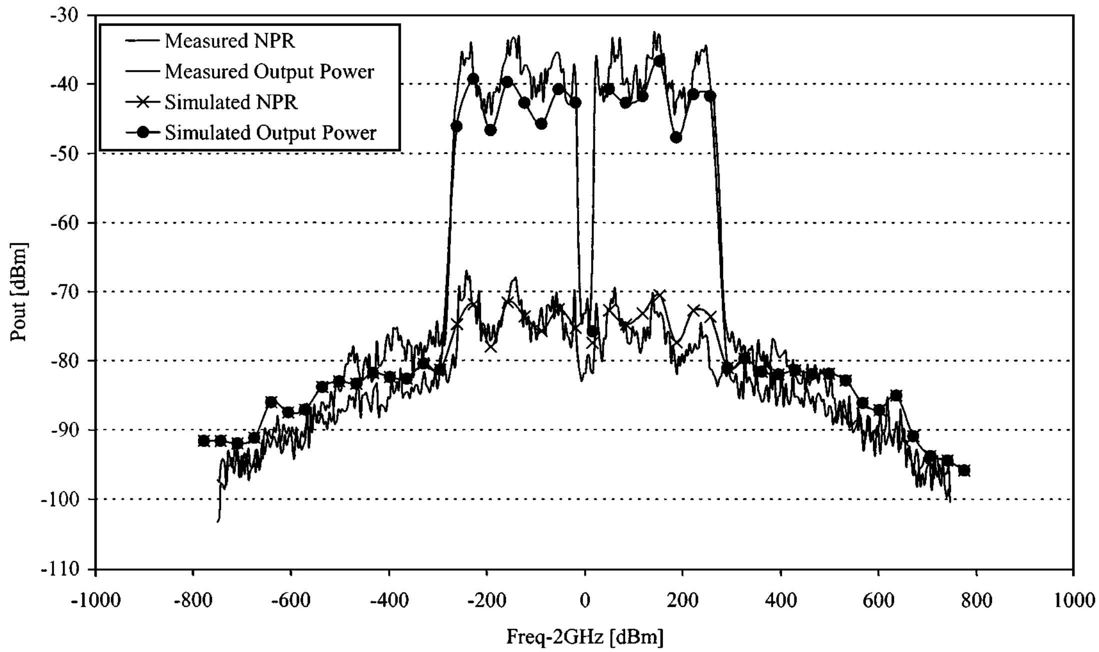

Fig. 2. Conventional NPR test results as measured with the proposed setup and simulated with our HB simulator.

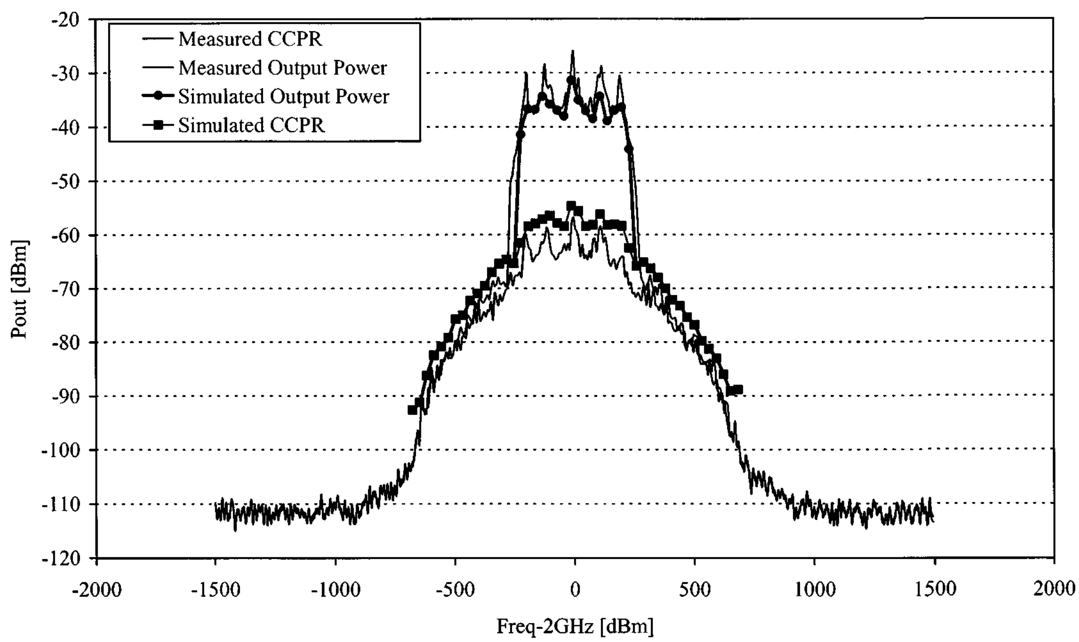

Fig. 3. Corrected CCPR test results as measured with the proposed setup and simulated with our HB simulator.

The results of the corresponding corrected co-channel distortion measurement and simulation are shown in Fig. 3.

The highest level traces are again the DUT's output at a drive level 19 dB above Pin_0, while the lowest ones are the bridge output after loop compensation. The distortion reveals itself when the linear components are cancelled. This is the true co-channel and adjacent distortion that would be obtained if the PA were operated under such a continuous spectrum excitation.

The inexistence of the valley on the in-band distortion previously observed in Fig. 2 is the proof of the misleading in-band evaluation predicted by [1] and [2] on conventional NPR tests. Moreover, a closer look onto these two diagrams reveals that the differences in distortion level observed at the bandwidth center are about 5 dB, and the differences between these results present at the channel edges are nearly 8 dB. These are the corrections of 5.6 and 7 dB theoretically predicted in [1] and [2] for the center and extremes of the operative bandwidth, respectively.

These results clearly validate the theoretical statement that NPR measurements underestimate co-channel IMD, and that a new distortion measurement setup like the one now described is needed.

In the HB simulator, the amplifier's active device nonlinearities were represented by a large-signal MESFET model especially conceived to predict IMD performance [9]. The input

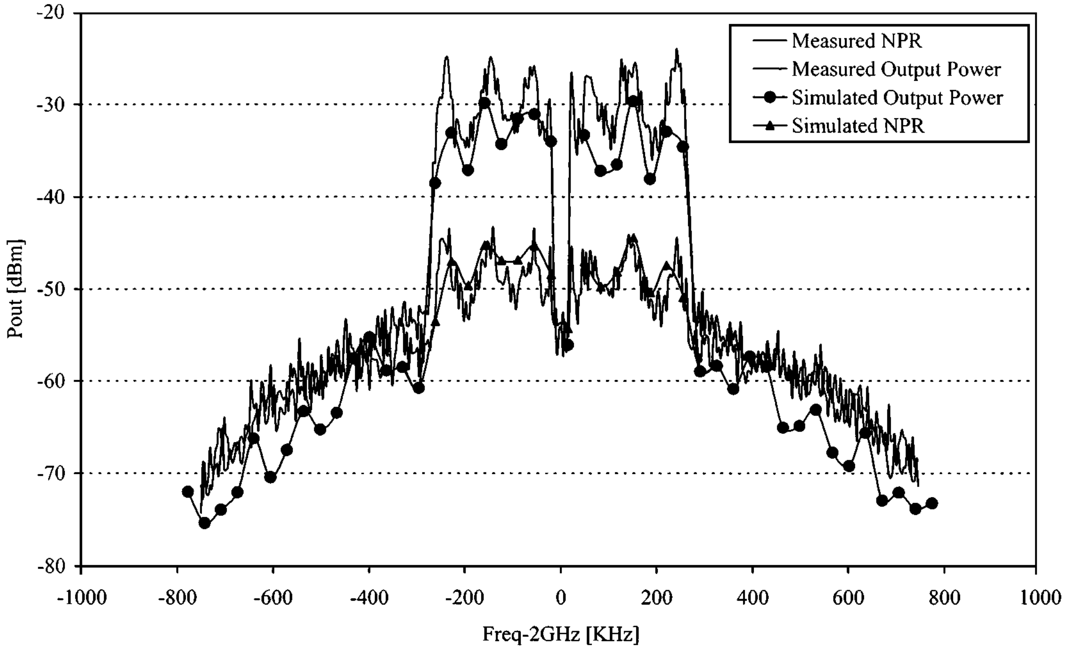

Fig. 4. Large-signal NPR test results as measured with the proposed setup and simulated with our HB simulator.

spectrum was sampled with 16 uniformly distributed tones, whose individual amplitudes were calculated by integrating the input power spectrum density function in the corresponding bandwidth slots. The tone phases were selected as random and, thus, every simulated data results from averaging 15 different simulation outputs obtained with 15 random phase sets.

The method used for simulating corrected CCPR closely follows the operation routine explained for the experimental setup [3]. The only difference is that the output linear components can now be canceled directly by simply subtracting them from an amplified version of the excitation spectrum. The necessary exact amplifier gain is automatically present at the HB Jacobian matrix calculated in the quiescent point.

The close agreement obtained in Figs. 2 and 3 between measured and simulated data validates this HB simulator, converting it in an alternative and powerful way of CCPR evaluation. In fact, due to the exactness of the calculated linear gain and the high numerical range provided by the artificial frequency mapping used in the HB core [7], the simulator allows almost unlimited SNR and, thus, CCPR measurements.

IV. Large-Signal NPR and CCPR Experimental Results

To extend these discussions to large-signal operating regimes, where the assumptions of [1] and [2] fail, the NPR and CCPR experiments of Figs. 2 and 3 were repeated for a driving level that puts the DUT near its 1-dB compression point. The corresponding measured and simulated results are shown in Figs. 4 and 5, respectively.

When compared to the small-signal results previously shown in Figs. 2 and 3, there are two evident differences.

The first one is the expected visibility of the alternate channel power, now appearing as sidebands to the already existing adjacent channel distortion. This new form of spectral regrowth, typical of systems of order higher than three, is a clear indication of strong nonlinear effects and, thus, of an amplifier operation regime close to saturation.

The second and most important one is the enlarged difference between in-band distortion power observed with the NPR and CCPR tests. That difference increased from nearly 5 to 8 dB in the bandwidth center, and from approximately 8 to 13 dB in the bandwidth extremes.

To understand this intriguing observation, one additional NPR and CCPR experiment was performed under higher driving levels, as is shown in Figs. 6 and 7.

By comparing these results with the ones previously reported, it was concluded that those differences remain approximately constant up to the onset of a strong nonlinear regime, and then suffer a rapid and monotonic increase when the device gets into deep saturation. Also, it could be noted that this behavior followed the DUT's gain compression characteristic. The knowledge that NPR tests eliminate distortion components correlated with the linear signal and, thus, are not capable of detecting gain compression phenomena, is now crucial to explain this continuously untying CCPR and NPR behavior. The energy balance of any amplifier dependent on a limited available power supply determines that both the output signal and uncorrelated distortion components tend to saturate to constant values. Also, since the overall distortion was assumed to be the deviation of the actual amplifier output to the ideal one (that would be given by a constant gain linear performance), it is clear that the correlated distortion must permanently increase to compensate the actual output power compression.

In this sense, one must conclude that the theoretical derived difficulties associated with NPR tests in evaluating small-signal in-band distortion are progressively enlarged when the DUT is pushed into saturation.

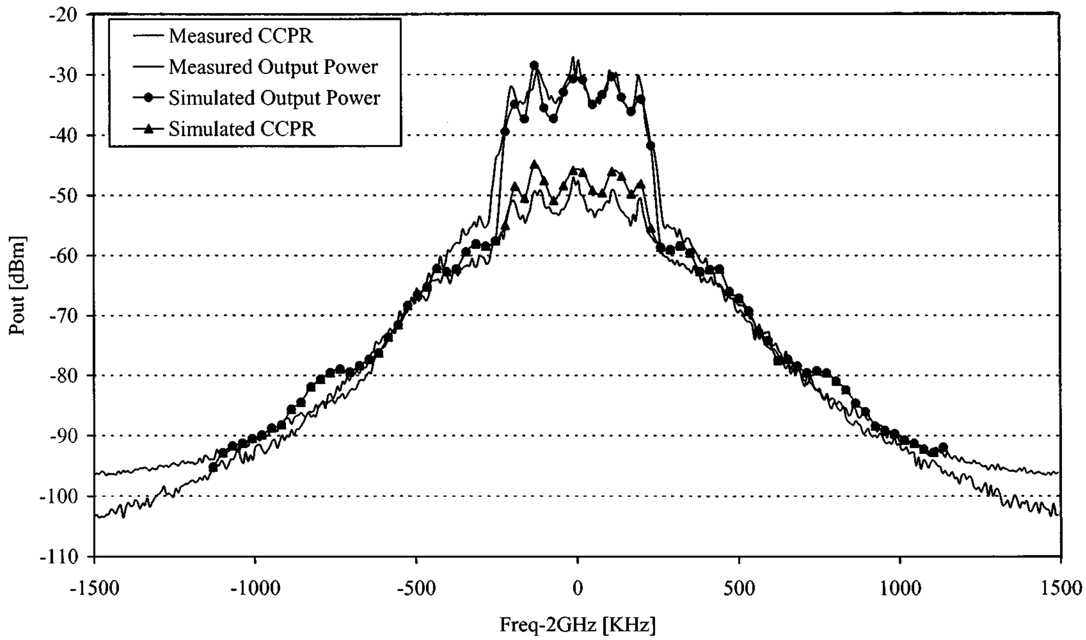

Fig. 5. Large-signal CCPR test results as measured with the proposed setup and simulated with our HB simulator.

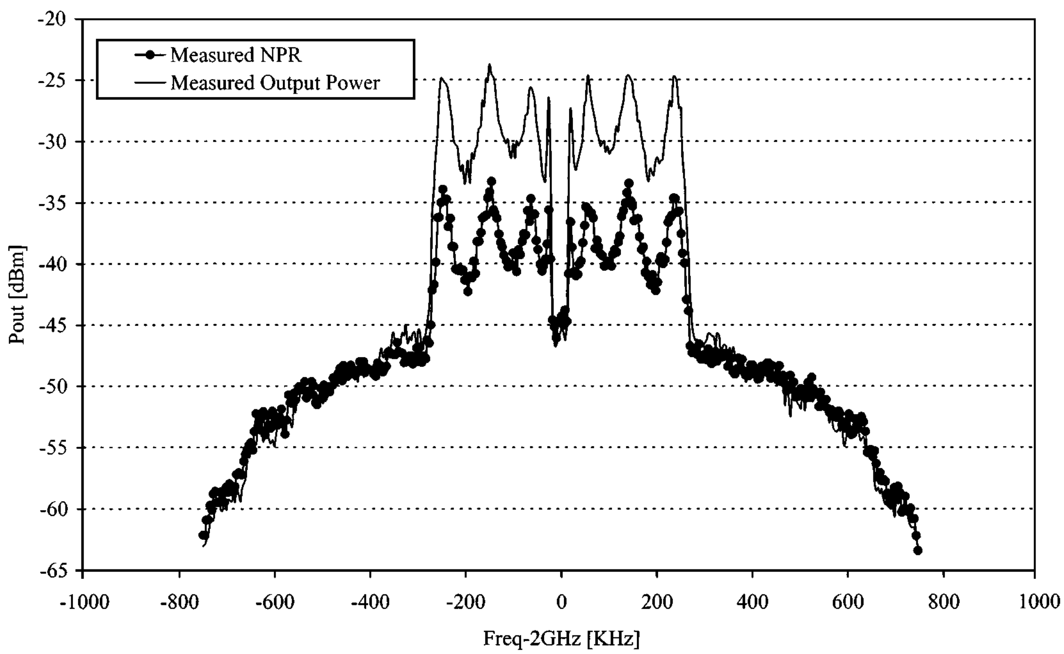

Fig. 6. Experimental NPR test results when the DUT is in deep compression.

V. Moderate Bandwidth Co-Channel Distortion Evaluation

This section is devoted to extend the theoretical predicted results of co-channel distortion, derived for memoryless systems, to frequency-dependent devices.

Since the normal use of narrow-band microwave amplifiers assumes the input signal bandwidth is comfortably enclosed in the system's bandwidth, it can be expected that $\mathrm{H}_1(\omega_1) \approx \mathrm{H}_1(\omega_2)$ for any ω_1 and ω_2 pertaining to the signal spectrum. Accordingly, the constancy of $\mathrm{H}_2(\omega_1, \omega_2)$ and $\mathrm{H}_3(\omega_1, \omega_2, \omega_3)$ can also be guaranteed for every set of ω_1, ω_2, and ω_3, whereas $\mathrm{H}_2(\omega_1, -\omega_2)$ should be expected to change. In fact, although in a narrow-band amplifier the difference frequency occupies an insignificant relative bandwidth compared to the fundamentals, the second or third harmonics, it represents a large value near dc. Therefore, it may be expected that frequency-dependent effects in this type of amplifiers are most probably caused by eventual variations of their baseband terminating impedances, in the whole range of possible $\Delta\omega = |\omega_1 - \omega_2|$.

In the amplifier under study, it was found that the base-band terminating impedance is nearly a short circuit if $\Delta\omega$ is kept below some 30 MHz, become clearly inductive until 165 MHz, and capacitive over this value. Thus, for studying the IMD

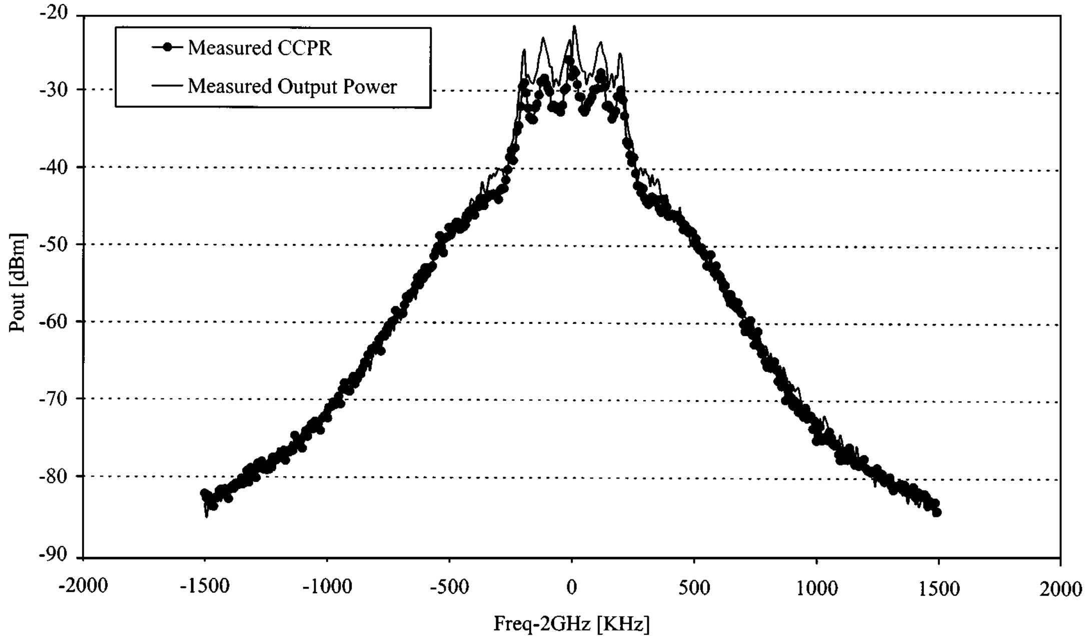

Fig. 7. Experimental CCPR test results when the DUT is in deep compression.

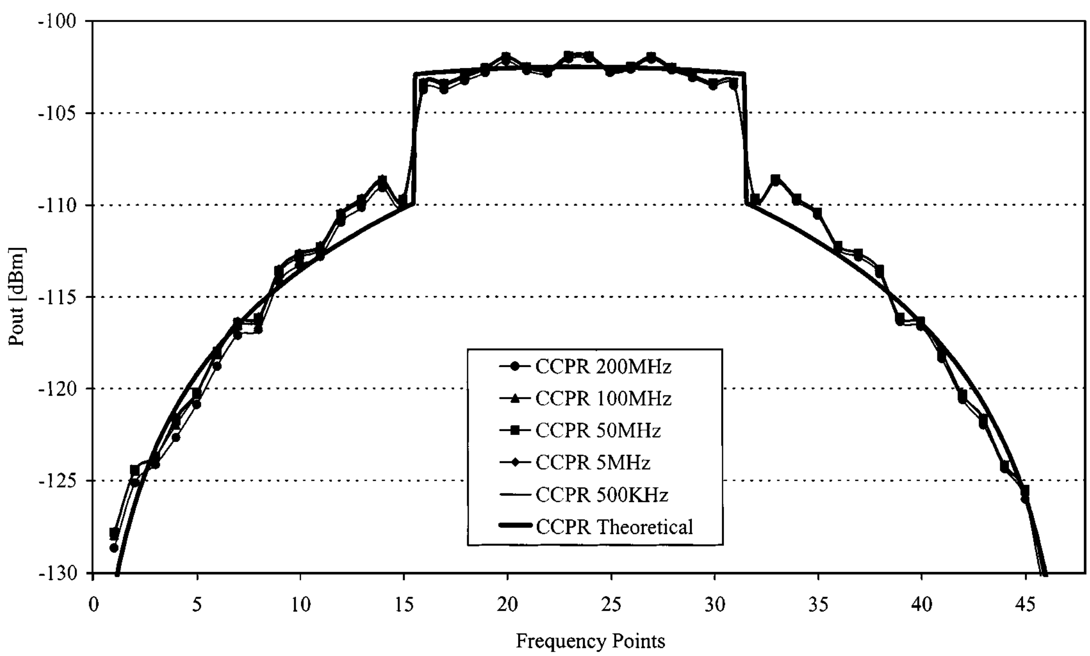

Fig. 8. Simulated CCPR results in small-signal operation for relative bandwidths of 500 kHz, 5 MHz, 50 MHz, 100 MHz, and 200 MHz.

dependence on frequency variations, we tested the amplifier for bandwidths of $\mathrm{Bw} = 500$ kHz (0.025%), 5 MHz (0.25%), 50 MHz (2.5%), 100 MHz (5%), and 200 MHz (10%). Since the implemented CCPR measurement setup cannot handle such wide bandwidths, the above referred HB simulator was used for this purpose. Simulated results of these CCPR tests are plotted in Fig. 8 along with the theoretical predictions of [1] and [2].

As can bee seen, the simulated and theoretically predicted results are almost coincident. This is a consequence of the fact that the IMD of this DUT is dominated by third-degree nonlinearities and not by mixing products generated by interactions between first- (fundamentals) and second-order mixing components [10]. The ripple observed in the simulated results is due to the modulated nature of the input spectrum. In fact, although the discretized tone amplitudes were all equal, the phases were not selected as completely random. A random phase was attributed to one-half of the tones, while this phase set was simply mirrored for the other half. This is implied by the real base-band signal, (which was then up-converted to C-band) and, thus, to the fact that its positive frequency components must be complex conjugate of the negative ones.

Since the situation used is believed to be a good illustration of the conditions usually found in communication systems, we can conclude that these results allow the extension of the theoretically predicted results valid for memoryless devices to normally operated microwave amplifiers.

VI. Conclusions

An experimental setup and an in-house developed HB simulator have been presented as two complementary means of microwave amplifier corrected co-channel distortion evaluation. Measurement and simulated results of NPR and CCPR, under small-signal, large-signal, and wide-band excitations, allowed the generalization of previously published results valid for third-degree memoryless nonlinearities to general microwave devices of practical interest.

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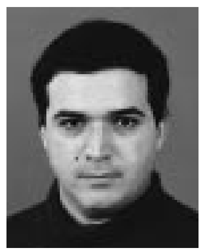

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