Analysis and Measurement of Multi-tone Intermodulation Distortion of Microwave Frequency Converters

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Abstract — **Using a time-variant 3rd order model, it is shown that the distortion arising in microwave frequency converters can be identified by signal-uncorrelated distortion components, but also by another signal-correlated part. Because this correlated contribution cannot be observed with conventional intermodulation set-ups, a new measurement bench is proposed and tested.**

I. Introduction

Nonlinear distortion has recently deserved augmented attention because of its increased importance in modern telecommunications systems. Unfortunately, the analysis complexity of frequency converters, as compared to non frequency-translating devices, has directed those studies to nonlinear amplifiers [1], [2], while little has been done to mixers or modulators.

Although the standard nonlinear distortion analysis procedure for microwave mixers has already been established for more than 20 years [3], [4], it has not significantly evolved from the original two-tone case. However, the emerging modern wireless systems created applications requiring results of distortion arising in mixers subject to multi-tone or even continuous spectra excitations. First steps to satisfy those needs resulted in nonlinear simulations of special circuits, and under specific radio-frequency, RF, driving signals [5], which do not allow generalizations to other circuits or excitations.

Thus, a general analysis of the various forms of distortion arising in nonlinear mixers and modulators, like adjacent-channel distortion power ratio, ACPR, or co-channel distortion power ratio, CCPR, is still an open problem. Without that knowledge, it is almost impossible to derive, or even estimate, those complex signal distortion figures of merit, and thus predict the impact of the device in the overall system performance.

The main goal of this work is to present a first answer to those questions. For that, the standard mixer distortion analysis procedure of Maas [4] was followed, allowing the derivation of general closed-form expressions for in-band distortion of frequency converters subject to band-limited random inputs. Furthermore, and according to what was already observed for mildly nonlinear amplifiers [1], [2], it could be found that in-band distortion in mixers can be divided into two major groups. First one includes components uncorrelated with the output fundamentals. They can be coincident in frequency with the output intermediate frequency, IF, – thus constituting a form of co-channel distortion – while others fall exactly at its upper and lower sidebands – creating-adjacent channel distortion. The second group of components is correlated with the IF output fundamentals, and is the responsible for modeling conversion-gain compression effects. It falls exactly over the desired IF bandwidth, constituting another form of co-channel distortion.

Because conventional multi-tone intermodulation ratio, M-IMR, ACPR, or even noise power ratio, NPR, test setups are only sensitive to signal-uncorrelated nonlinear components, we also propose, and experimentally validate, a new distortion measurement bench.

II. Simplified Nonlinear Analysis of Frequency Converter's Distortion

Established weakly nonlinear analysis technique of mixer distortion [3], [4] assumes the frequency-translating device is described by a 3rd order time-varying Volterra series, in which the IF distortion components are computed in a recursive way, using the nonlinear currents method. In order to obtain closed form expressions for the distortion arising in such a system excited by a complex RF signal, without having to deal with unimportant detailed circuit analysis, our general frequency converter is herein assumed to be a memoryless transfer nonlinearity with respect to its input (RF-Port) and output (IF-Port). This allows a replacement of the Volterra analysis of considerable complexity into a much less involved time-varying power series model:

$$v_{IF}(t) = c_1(t) v_{RF}(t) + c_2(t) v_{RF}(t)^2 + c_3(t) v_{RF}(t)^3 \qquad (1)$$

The RF input signal, $v_{RF}(t)$, is a band limited white gaussian noise, generally accepted as a typical illustration of a large range of real wireless signals. It extends over a bandwidth of Bw, centered at ω_{RF}, where it has a constant power spectral density function, PSD, of N_0. Its integrated input power is thus $Pin= N_0.Bw$.

Because our model is only valid for mildly nonlinear operation regimes, the amplitude of the RF, Pin, is assumed much smaller than that of the local oscillator, LO, sinusoid. This permits the assumption of non-interacting LO and RF signals, which enables a strongly nonlinear solution of the circuit to the LO, followed by a mildly nonlinear analysis to the RF, superimposed on the previously determined independent LO solution. The drawback associated to this method is that the quasi-linearity assumption adopted for treating the RF obviates any attempt to predict distortion levels when the device is driven close to, or beyond, the 1dB compression point.

As in the linear case, the analysis begins by determining a time-varying quiescent point imposed by the LO periodic response. Expanding each of the LO time-varying Taylor series coefficients of (1) in a Fourier series, we get:

$$v_{IF}(t) = \left(\sum_{k=-K}^{K} C_{1,k}\, e^{jk\omega_{LO}t} \right) v_{RF}(t) + \left(\sum_{k=-K}^{K} C_{2,k}\, e^{jk\omega_{LO}t} \right) v_{RF}(t)^2 + \left(\sum_{k=-K}^{K} C_{3,k}\, e^{jk\omega_{LO}t} \right) v_{RF}(t)^3 \qquad (2)$$

which, in the frequency-domain, corresponds to:

$$S_{IF}(\omega) = \left(\sum_{k=-K}^{K} \left|C_{1,k}\right|^2 \delta(\omega - k\omega_{LO}) \right) * S_{RF}(\omega) + \left(\sum_{k=-K}^{K} \left|C_{2,k}\right|^2 \delta(\omega - k\omega_{LO}) \right) * S_{RF}(\omega) * S_{RF}(\omega) + \left(\sum_{k=-K}^{K} \left|C_{3,k}\right|^2 \delta(\omega - k\omega_{LO}) \right) * S_{RF}(\omega) * S_{RF}(\omega) * S_{RF}(\omega) \qquad (3)$$

in which the $C_{i,k}$ are the entries of the equivalent i'th order conversion matrix [4], $S_{RF}(\omega)$ and $S_{IF}(\omega)$ are the RF-input and IF-output PSD, $\delta(\omega - k\omega_{LO})$ is a Dirac delta function centered at the k'th LO harmonic and * stands for spectral convolution.

Expression (3) is composed of a large set of noise bands, each one centered in a certain k'th LO harmonic, $k\omega_{LO}$. Supposing (without loss of generality) the case of a down converter, in which the RF signal spectrum is located above ω_{LO}, the desired in-band IF components are centered at $\omega_{IF}=\omega_{RF}-\omega_{LO}$ (k=-1) and occupy a bandwidth of $3Bw$. Their PSD is then given by expression (4).

This expression includes lower and upper sideband components of parabolic shape expressed by (4.a) and (4.c), respectively, which constitute adjacent-channel distortion; but also co-channel distortion components as given by (4.b).

A close inspection into (4.b) will show that co-channel components can really be divided into two distortion components, besides the expected linear power represented by $|C_{1,-1}|^2(N_0/2)$. The first one is due to the term $|3C_{3,-1}Bw|^2(N_0^3/2)$ and stands for 3rd order co-channel distortion that is correlated with the output linear power. Note, for example, that its shape is not parabolic, as are the other 3rd order components, but constant, like the linear IF. The second term is similar to the sideband intermodulation (although of opposite concavity) and represents co-channel signal-uncorrelated distortion.

This scenario, analogous to what was previously detected in time-invariant systems [2], also implies that while ACPR or M-IMR tests could be sufficient for adjacent-channel distortion characterization, NPR testing misleads co-channel distortion evaluation. In fact, since an NPR test works by eliminating a slice of the RF input, and consequently of the linear IF output, it also destroys the correspondent signal-correlated distortion.

$S_{IF}(\omega) =$

$$= 18\left|C_{3,-1}\right|^2 \left(\frac{N_0}{2}\right)^3 \left[\frac{\omega^2}{2} - \left(\omega_{IF} - \frac{3}{2}Bw\right)\omega + \frac{1}{2}\left(\omega_{IF} - \frac{3}{2}Bw\right)^2 \right] \quad \text{if} \quad \omega_{IF} - \frac{3}{2}Bw < \omega < \omega_{IF} - \frac{Bw}{2} \qquad (4.a)$$

$$= \left|C_{1,-1} + 3C_{3,-1}N_0 Bw\right|^2 \frac{N_0}{2} + 18\left|C_{3,-1}\right|^2 \left(\frac{N_0}{2}\right)^3 \left[-\omega^2 + 2\omega_{IF}\omega + \frac{3}{4}Bw^2 - \omega_{IF}^{\ 2} \right] \quad \text{if} \quad \omega_{IF} - \frac{Bw}{2} < \omega < \omega_{IF} + \frac{Bw}{2} \qquad (4.b)$$

$$= 18\left|C_{3,-1}\right|^2 \left(\frac{N_0}{2}\right)^3 \left[\frac{\omega^2}{2} - \left(\omega_{IF} + \frac{3}{2}Bw\right)\omega + \frac{1}{2}\left(\omega_{IF} + \frac{3}{2}Bw\right)^2 \right] \quad \text{if} \quad \omega_{IF} + \frac{Bw}{2} < \omega < \omega_{IF} + \frac{3}{2}Bw \qquad (4.c)$$

Therefore, convenient distortion testing of frequency converters requires an alternative measurement setup that is able to simultaneously get rid of the perturbing IF linear components, while preserving RF input integrity. An example of such setup is presented and tested in next section.

III. NEW FREQUENCY CONVERTER IN-BAND DISTORTION MEASUREMENT SETUP

Fig. 1 presents a new setup amenable for measuring the whole in-band distortion of any frequency converter.

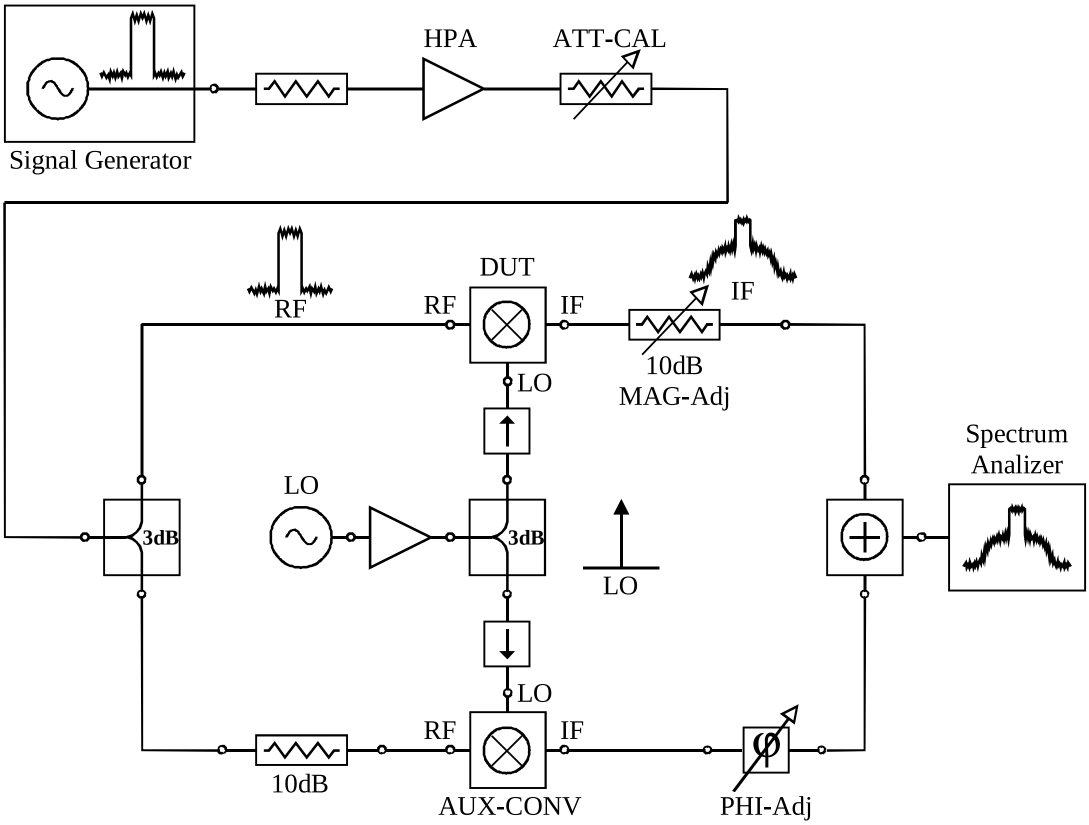

Fig. 1. Proposed measurement setup for in-band distortion evaluation of frequency converters.

By using a frequency translating bridge configuration, this setup guarantees that the whole RF input is applied to the DUT, while the linear components are appropriately cancelled at the output. Since correct operation demands that the bridge must be adjusted (tuning MAG-Adj and PHI-Adj) at a reasonably large input power back off (increasing ATT attenuation), it is assured that only IF linear components are eliminated, while not affecting desired signal-correlated distortion. As in any feedforward signal cancellation loop [6], it is required that the auxiliary arm be linear. For that, we could either use a much more linear auxiliary converter, AUX-CONV, or one that is similar to the DUT, but driven with significantly less RF power. In the implemented setup, we adopted the latter solution. For example, if a 10dB back off is considered for the AUX-CONV, in comparison to the DUT, the amount of AUX-CONV distortion level would be 20dB below the DUT one, which already guarantees a measurement error not greater than 1dB.

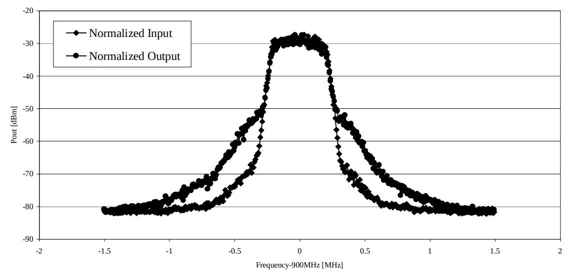

Fig. 2. RF signal input and complete IF output as applied and measured on the DUT.

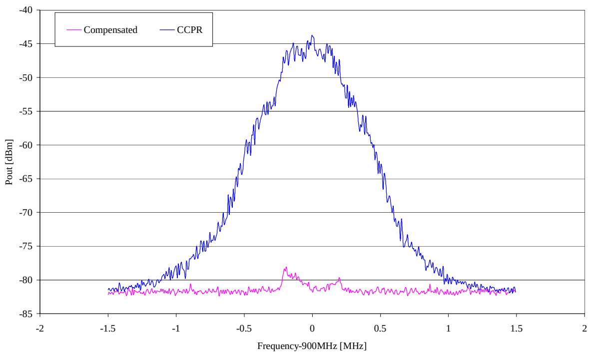

Fig. 3. Test setup output signals as measured in the bridge adjustment condition (11dB input back off), and at nominal DUT drive level.

Fig. 2 and Fig. 3 represent an example of the measurements that were performed with such a setup. Fig. 2 shows the RF input applied to the DUT, along with the complete output IF. This would be the expected result of an usual mixer intermodulation measurement. On the other hand, Fig. 3 shows the bridge output when the DUT is driven by a very low noise power level (and so, just after bridge adjustment), and at the same RF input noise level used for Fig. 2. This is the true DUT in-band distortion. Please note the same adjacent-channel sidebands as seen in Fig. 2, but also co-channel distortion components. From these, it is worth observing the level jump present exactly at Bw edges. Since expression (4) indicates that the signal-uncorrelated components should be continuous at these interfaces, one must conclude that this jump is exactly due to the signal-correlated distortion term $|3C_{3,-1}Bw|^2(N_0^3/2)$, valuing about 7dB. So, it can be concluded that this should be approximately the error committed by a conventional NPR test, if it were used to estimate the DUT's co-channel distortion.

IV. CONCLUSIONS

A band-limited random input distortion analysis was presented for mildly nonlinear frequency converters. Assuming a 3rd order time-variant memoryless model, it could be shown that the in-band distortion components include adjacent- and co-channel signal-uncorrelated distortion, but also signal-correlated co-channel distortion. Since the correct evaluation of the latter components cannot be done using traditional intermodulation distortion test setups, an alternative measurement bench was proposed, and its utility established.

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