CALIBRATED MEASUREMENTS OF NONLINEARITIES IN NARROWBAND AMPLIFIERS APPLIED TO INTERMODULATION AND CROSS MODULATION COMPENSATION

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Abstract — **A vectorial nonlinear-network analyzer (VNNA) is absolutely calibrated in amplitude and phase for narrowband measurements of nonlinearities. Calibration is performed with a reference generator, characterized with a calibrated sampling oscilloscope. The reference signal, consisting of an upconverted baseband signal, is optimized for statistically efficient measurements. Using the calibrated VNNA, narrowband amplifiers can be measured, enabling compensation of intermodulation and cross-modulation.**

I. INTRODUCTION

In this paper the vectorial nonlinear-network analyzer concept [1] is extended for measuring narrowband devices. This is very important for testing and designing narrowband communication systems. Several mathematical tools exist for describing and analysing these components [2]. However, the measurements of their parameters is a difficult problem. We used a VNNA, already available for broadband measurements at a fixed equidistant grid. This approach can be extended for narrowband measurements. An important issue is the calibration of the system, which should now be carried out on a dense grid around the band of interest only. Therefore we propose a variation of the reference generator absolute calibration method described earlier [1]. The advantages of this new method are a more flexible reference signal, which can easier be adapted for statistically efficient measurements. Indeed, it was shown that the signal-to-noise ratio during the calibration measurements depends on the crest factor (peak value to effective value ratio) of the reference signal. A new method will be described for making narrowband signals with very low crest factors. The uncertainty on the absolute calibration factor will decrease due to this new calibration reference generator.

Once our VNNA is calibrated this way, both amplitude and phase of the Volterra kernels of a narrowband amplifier can be determined accurately. Thanks to the absolute calibration, measurements are repeatable on different VNNA's. The measured kernels can be used to determine the third order intermodulation and cross modulation distortion. Also, the accurate knowledge of the Volterra kernels of the main amplifier, both in amplitude and phase, are essential for a successful distortion compensation.

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II. MEASUREMENT SETUP

The vectorial nonlinear-network setup is described in Fig.1.

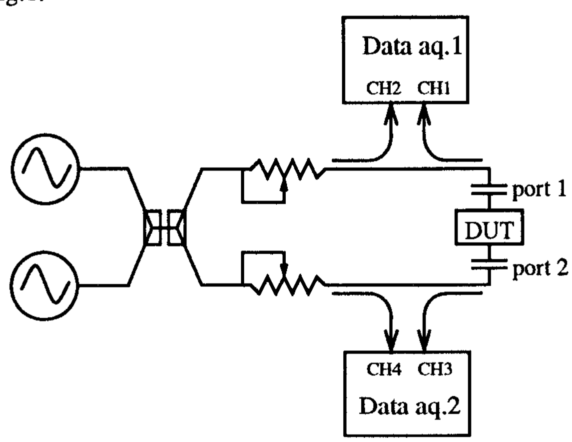

Fig. 1. Measurement setup

The system consists of a signal generator (two synthesizers HP83640A with a power combiner), a test set (directional couplers to detect incident and scattered waves) and 4 data acquisition channels. For the last part, several alternatives are available [1], e.g. a broadband sampling oscilloscope or the broadband downconvertors in a microwave transition analyzer (MTA). This system is able to coherently measure all harmonics of the incident and scattered waves at the test ports. However, since systematic errors are present, a calibration is indispensable. This requires not only a relative calibration (e.g. SOLT) but also an absolute calibration in amplitude and phase [1]. This is done using a reference generator, i.e. a multitone generator with an accurately known output impedance, where the absolute amplitudes and relative phases of all frequency components are stable and accurately specified. This reference generator is characterized by an accurate broadband signal analyzer. As signal analyzer a broadband sampling oscilloscope (HP-54121T) is used, which means that the calibration is traceable to a "nose-to-nose" procedure [5].

The requirements for the reference generator are different for narrowband and broadband measurements. For the narrowband measurements, only frequencies in the band of interest should be present. Also, we will show that a much more flexible signal design with a better statistic efficiency is possible now. Indeed, lower crest factor signals ($Cr = L_\infty/u_{eff}$, where L_∞ and u_{eff} are respectively the L-infinity norm or peak value and the effective value of the signal) can be made now, which results in higher signal-to-noise ratio during measurement of these signals [3], [4]:

$$\frac{S}{N} = \gamma - 20\log Cr \qquad (1)$$

where γ is a constant.

III. Reference Signal for Narrowband Calibrations

We will now propose a method for generating narrowband, high-frequency, multisine signals with low crest factor.

A. Double sided upconversion

Consider an arbitrary bandlimited baseband signal $B(\omega)$.

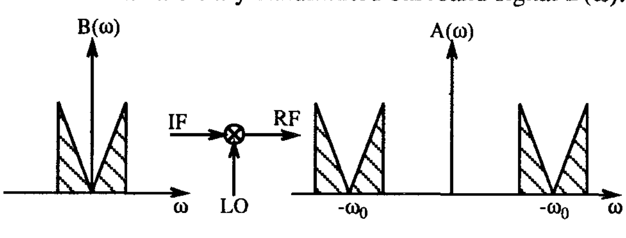

Fig. 2. Spectrum of double sided upconverted baseband signal

This signal is upconverted to a band around ω_0 using an ideal double sided mixer to $A(\omega)$:

$$A(\omega) = B(\omega - \omega_0)e^{j\phi} + B(-\omega + \omega_0)e^{-j\phi} \qquad (2)$$

This corresponds to a time signal:

$$\begin{aligned} a(t, \phi) &= \mathfrak{I}^{-1}(A(\omega)) \\ &= 2b(t)\cos(\omega_0 t + \phi) \end{aligned} \qquad (3)$$

Let us now consider the envelope of this time signal:

$$E_a(t) = \max_\phi(a(t, \phi)) = 2|b(t)| \qquad (4)$$

where b(t) is the time domain baseband signal. It can easily be verified that the RMS values are related by

$$RMS(E_a(t)) = \sqrt{2}RMS(a(t)) \qquad (5)$$

Since the crest factor is defined as the ratio of peak value to effective value

$$Cr(x) = \frac{x_{peak}}{x_{rms}} = \frac{\max_t(|x(t)|)}{x_{rms}} \qquad (6)$$

and since the maximum value of the signal approaches that of the envelope for large ω_0/ω_{max}, we have:

$$Cr(a(t)) \le \sqrt{2}Cr(E_a(t)) = \sqrt{2}Cr(b(t)) \qquad (7)$$

In Fig.1, an example of an optimized signal is shown. We considered a baseband signal with 12 frequency components of equal amplitude, spaced 1 MHz apart. This signal is intended for upconversion with a local oscillator of 6013 MHz to a band of [6001 MHz, 6025 MHz]. After optimization, the crest factor of the envelope is 1.39, which means that the complete signal (modulated carrier) will have a crest factor of 1.97.

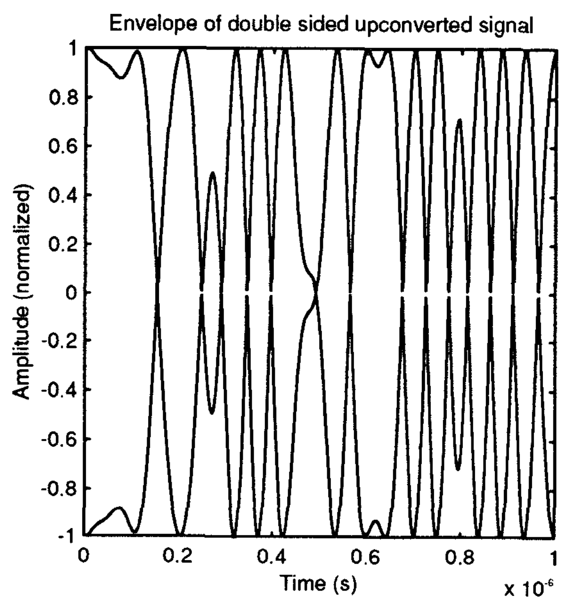

Fig. 3. Envelope $E_a(t)$ - double sided case

B. Single sided upconversion

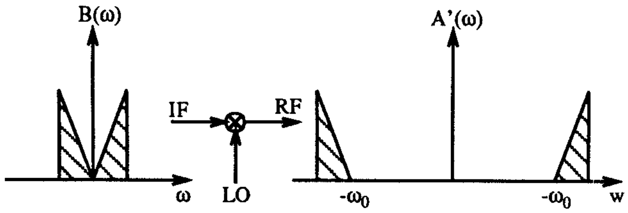

Fig. 4. Spectrum of single sided upconverted baseband signal

Let us now consider the same bandlimited baseband signal B(ω)=G(ω)+G(-ω), this time upconverted to a band around ω_0 using an ideal single sided mixer:

$$A'(\omega) = G(\omega-\omega_0)e^{j\phi} + G(-\omega+\omega_0)e^{-j\phi} \quad (8)$$

where G(ω) is the one sided baseband signal (complex). This corresponds to a time signal:

$$\begin{aligned} a'(t,\phi) &= \Im^{-1}(A'(\omega)) \\ &= \mathrm{Re}\left(g(t)e^{j(\omega_0 t+\phi)}\right) \end{aligned} \quad (9)$$

The envelope of this time signal is:

$$E_a(t) = \max_\phi(a'(t,\phi)) = 2|g(t)| \quad (10)$$

The RMS values of the signal a'(t) and its envelope are related by:

$$\mathrm{RMS}(E_{a'}(t)) = \sqrt{2}\mathrm{RMS}(a'(t)) \quad (11)$$

Consequently:

$$\mathrm{Cr}(a'(t)) \le \sqrt{2}\mathrm{Cr}(E_{a'}(t)) = \sqrt{2}\mathrm{Cr}(g(t)) \quad (12)$$

This is the same result as for the two-sided case, except that g(t) now is a complex time function. Since the crest factor of complex time signals can be reduced to lower values (typically 1.1) than that of real signals (typically 1.4) [6], the resulting narrowband multisine will have a lower crest factor too.

In Fig.5, again an example of an optimized signal is shown. This time we considered a baseband signal with 25 frequency components of equal amplitude, spaced 1 MHz apart. The frequency band of the upconverted signal is again [6001 MHz, 6025 MHz]. The crest factor of the envelope is now 1.11 and for the complete signal 1.58. A zero phase signal with the same amplitude spectrum but with the phases of all components set to zero would have a crest factor of 7.07. According to (1), this means that our single sided upconverted optimized multisine can be measured with a 13 dB better signal to noise ratio than a zero phase signal.

IV. Experimental Results

The single sided upconverted multisine described above was realized using an HP 8770A arbitrary waveform synthesizer (frequency band used [26;50] MHz) and two mixer stages (LO1=800 MHz and LO2=5175 MHz) with band-pass filters for selecting the upper sideband. This system will be replaced in the future by single sideband mixers. All sources are locked to a 10 MHz reference. The resulting multisine is measured with an HP54120T sampling oscilloscope, triggered by the period of the arbitrary waveform generator. The timebase of the sampling oscilloscope is calibrated as described in [9]. The spectrum is estimated from the nonuniform sampled time points using a least square technique. The results of these measurements are

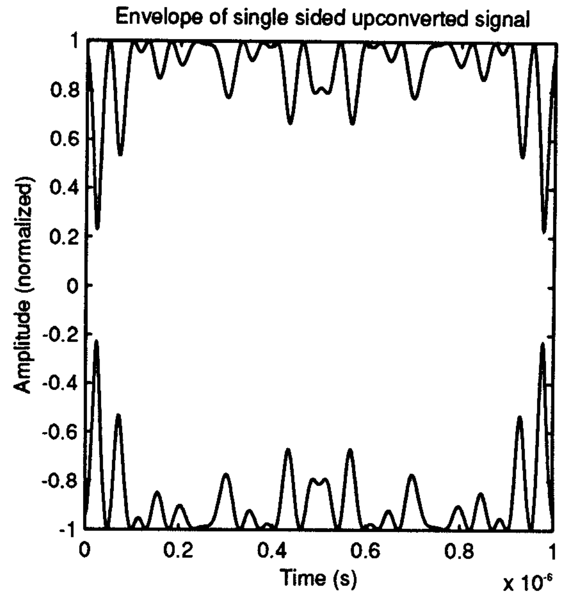

Fig. 5. Envelope $E_a(t)$ - single sided case

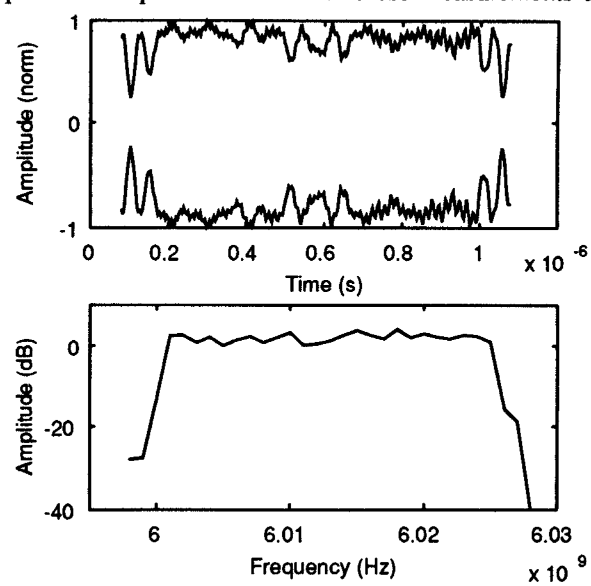

Fig. 6. Measured envelope and amplitude spectrum - single sided case

shown in Fig. 6. The measured crest factor is 1.85, while the amplitude spectrum is flat within 2 dB in the band of interest. The crest factor is slightly larger than expected due to this 2dB variation and the presence of residual LO and suppressed sidebands (all at least 15 dB down).

V. INTERMODULATION AND CROSS-MODULATION COMPENSATION

A. Definitions

According to [8], both the third order intermodulation IM3 and the cross modulation CM_F of a narrowband amplifier are function of the Volterra kernels H_i:

$$IM_3 = \frac{3}{4}\frac{|H_3(j\omega_1, -j\omega_1, j\omega_2)|}{|H_1(j\omega_2)|}V^2 \qquad (13)$$

$$CM_F = 3\frac{|H_3(j\omega_1, -j\omega_1, j\omega_2)|}{|H_1(j\omega_2)|}V^2 \qquad (14)$$

$$\phi = \angle H_3(j\omega_1, -j\omega_1, j\omega_2) - \angle H_1(j\omega_1) \qquad (15)$$

where V and ω_i are the input component amplitude and frequencies respectively. $CM_F \sin\phi$ is the amplitude cross modulation and $CM_F \cos\phi$ is the phase cross modulation.

B. Feedback amplifier

If a unilateral feedback $F(j\omega)$ is applied to an amplifier $H(j\omega)$, IM_3 and CM_F are reduced by a factor [8]

$$\frac{1}{1 + F(j\omega)H(j\omega)} \qquad (16)$$

Also, ϕ' after feedback is given by

$$\phi' = \phi - \angle(1 + F(j\omega_2)H_1(j\omega_2)) \qquad (17)$$

This indicates that the amplitude and phase cross modulation can be controlled through the phase of $F(j\omega_2)$. Note however that for an exact compensation, the knowledge of the amplitude and phase of a third order Volterra kernel is indispensable (through ϕ). This can only be measured in practice with a calibrated VNNA as we proposed.

C. Feedforward amplifier

From the Volterra kernels of an overall feedforward amplifier follows that intermodulation distortion can be made perfectly zero. However, in a practical realization, a certain gain imbalance a and phase imbalance δ remain. As a consequence, intermodulation distortion and crossmodulation don't vanish. It can be shown that [8]:

$$\phi' = \phi + \operatorname{atan}\frac{a\delta}{1 - a} \qquad (18)$$

where ϕ' is with and ϕ without feedforward. This illustrates again that the measurements of the amplitude and phase of the Volterra kernels of the main amplifier are important in the design process of a feedforward amplifier.

VI. CONCLUSION

The use of a vectorial nonlinear network analyzer in narrowband applications has been demonstrated. The need for an accurate absolute calibration requires a new technique based on a narrowband reference generator with a low crest factor to allow a calibration with a high statistic efficiency. A method for realizing such a signal is described based on upconverting a baseband signal and a new crest factor optimization technique. The calibrated VNNA is then used to measure the Volterra kernels of a narrowband amplifier. These results can be used to compensate for the intermodulation and crossmodulation in a feedback or feedforward amplifier configuration.

REFERENCES

[1] T.Van den Broeck and J.Verspecht, "Calibrated Vectorial Nonlinear-Network Analyzers," *IEEE MTT-S Int. Microwave Symp. Digest*, San Diego (USA), 1993, pp.1069-1072.

[2] F.Verbeyst and M. Vanden Bossche, "Viomap, the S-Parameter Equivalent for Weakly Nonlinear RF and Microwave Devices" *IEEE MTT-S Int. Microwave Symp. Digest*, San Diego (USA), 1993, pp.1369-1372.

[3] J. Schoukens and R. Pintelon, *Identification of Linear Systems*, Oxford: Pergamon Press, 1991.

[4] T.Van den Broeck, R. Pintelon and A.Barel, "Design of a Microwave Multisine Source Using Allpass Functions Estimated in the Richards Domain," *IEEE Trans. Instrum. Meas.*, vol. 43, pp.753-757, October 1994.

[5] J. Verspecht and K. Rush, "Individual Characterization of Broadband Sampling Oscilloscopes with a Nose-to-Nose Calibration Procedure," *IEEE Trans. Instrum. Meas.*, vol. 43, pp.347-354, April 1994.

[6] J.Schoukens, Y.Rolain and P.Guillaume, "Design of Narrow Band, High Resolution Multisines," ELEC internal report, JS3/93.

[7] R.G.Meyer, M.J.Shensa and R.Eschenbach, "Cross Modulation and Intermodulation in Amplifiers at High Frequencies," *IEEE Journal of Solid-State Circuits*, vol. 7, pp.16-23, February 1972.

[8] A.Javed, P.A.Goud and B.A. Syrett, "Analysis of Microwave Feedforward Amplifier Using Volterra Series Representation," *IEEE Transactions on Communications*, pp.355-360, March 1977.

[9] J. Verspecht, "Accurate Spectral Estimation Based on Measurements with a Distorted-Timebase Digitizer," *IEEE Trans. Instrum. Meas.*, vol. 43, pp.210-215, April 1994.