Linearity Optimization of a Distributed Base Station Amplifier using an Automated High-Speed Measurement Protocol

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***Abstract* — This paper describes a high-speed measurement protocol using a high-order implementation of the complex power series representation (CPSR). The CPSR is used to relate AM-AM and AM-PM conversion (large signal S_{21}) to the 3rd-order intermodulation distortion (IM3) of an amplifier up to gain compression. This method requires the use of a network analyzer only and eliminates the need for a spectrum analyzer speeding-up the measurements tremendously. The method has proven to be very powerful in a real-time optimization of bias parameter of a novel linear distributed amplifier for base stations.**

I. Introduction

Linearity is one of the most important issues in the design of base-station power amplifiers for wireless communication networks. To satisfy the high demands of complex amplitude modulated signals, like wide-band CDMA (W-CDMA), several linearization techniques have been developed (e.g. adaptive predistortion and feed-forward linearization) [1], [2]. However, the complexity of these solutions and the number of variables to be optimized has increased as well. For this reason, measurement and optimization of amplifier intermodulation distortion (IMD) versus signal power becomes more and more time-consuming due to the limited measurement speed of a conventional two-tone setup.

This paper describes a method to optimize power amplifier linearity in terms of IMD by minimizing AM-AM and AM-PM conversion (large signal S_{21}). IMD is related to the S_{21} as function of power by the complex power series representation (CPSR) [1]. Consequently, full amplifier characterization and tuning for matching, gain *and linearity*, is combined in a single instrument (network analyzer) test setup.

II. Using the CPSR for IM3 Estimation

The CPSR is a method to analyze non-linearities in an amplifier which is assumed to be memoryless. This assumption is valid if the pass-band of the amplifier is relatively narrow and exhibits a relatively constant frequency response over the pass-band. Furthermore, to reduce memory effects, bias modulation effects should be dealt with properly [2], [3]. If these conditions are met in practical wireless telecommunication amplifiers, third order intermodulation distortion (IM3) is completely characterized by the AM-AM and AM-PM conversion.

Our method for determining IM3 can be outlined as follows: First, we determine AM-AM and AM-PM conversion by measuring S_{21} versus input power using a vector network analyzer (VNA). Secondly, we obtain the required CPSR coefficients using a least square method to fit the data. And lastly, we compute the IM3 as function of power up to the gain compression region using the CPSR coefficients. Fig. 1 shows a black-box representation of the power amplifier used in the following analysis.

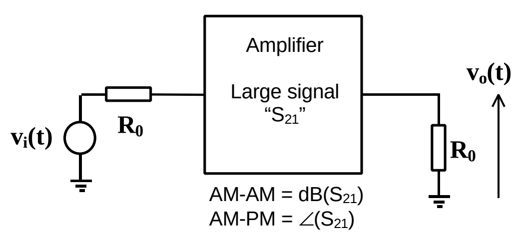

Fig. 1. Black-box representation of a non-linear amplifier characterized by its AM-AM and AM-PM conversion.

A. Characterizing AM-AM and AM-PM Conversion

Eq. (1) formulates the general expression for the complex power series in terms of voltage [1].

$$v_o(t) = a_1 v_i(t-t) + \mathrm{Re}\left[\sum_{n=3}^{N} \overrightarrow{a_n}\left\{v_i^n(t-t) + jH\left[v_i^n(t-t)\right]\right\}\right] \quad (1)$$

where,

$v_i(t)$, $v_o(t)$ are the input and output voltage,
a_1 is the linear gain,
$\vec{a}_n = R_n e^{jf_n}$ is a complex constant,
$H[v_i^n(t)]$ is the Hilbert transform of $v_i(t)$,
t is the delay-time of the amplifier,
and n is a positive odd integer.

In order to obtain an accurate description of the single-tone amplifier response over a wide range of input powers, the model needs some flexibility in calculating and fitting higher order terms describing the system. Therefore, it is convenient to introduce a generalized description of the system. Eq. (2) gives the generalized expression of the amplifier output voltage $v_o(t)$ at the fundamental frequency if we substitute a single sinusoid $v_i(t) = A\cos(w_0 t)$ in Eq. (1) and t is set to zero.

$$v_{o,w_0}(t) = \left[a_1 A + \sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}} A^n R_n \cos f_n \right] \cos w_0 t + \\ + \left[\sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}} A^n R_n \sin f_n \right] \sin w_0 t \qquad (2)$$

in which $\binom{n}{r} = \frac{n!}{r!(n-r)!}$ is the binomial coefficient

Eq. (2) shows that we can write the fundamental component as a series expression. Terms that are generated at higher harmonics are left out, because at this stage, we are only interested in the fundamental term for the AM-AM and AM-PM conversion. For further analysis Eq. (2) is rewritten in a more convenient form as expressed in Eq. (3).

$$v_{o,w_0}(t) = a_1 A R(A) \cos\{w_0 t + \Psi(A)\} \qquad (3)$$

where

$$R(A) = \left\{ \left[1 + \sum_{n=3}^{N} CA^{n-1} R_n \cos f_n \right]^2 + \left[\sum_{n=3}^{N} CA^{n-1} R_n \sin f_n \right]^2 \right\}^{\frac{1}{2}} \qquad (4)$$

$$\Psi(A) = \tan^{-1} \left\{ \frac{\sum_{n=3}^{N} CA^{n-1} R_n \sin f_n}{1 + \sum_{n=3}^{N} CA^{n-1} R_n \cos f_n} \right\} \qquad (5)$$

and

$$C = \frac{1}{a_1 2^{n-1}} \binom{n}{\frac{n+1}{2}}$$

In fact, $R(A)$ represent the AM-AM conversion normalized to a voltage gain of 1 and $\Psi(A)$ represents the AM-PM conversion if time delay is set to 0 degrees.

B. Obtaining the complex coefficients

To solve for the complex coefficients needed for our CPSR we have to fit our normalized measured large signal data (S_{21}/a_1) as function of input power. By rearranging (4) and (5) we can write the following system of equations

$$\frac{\mathrm{Im}(S_{21}(A))}{a_1} = R(A)\sin\Psi(A) = \sum_{n=3}^{N} CA^{n-1} R_n \sin f_n \\ \frac{\mathrm{Re}(S_{21}(A))}{a_1} = R(A)\cos\Psi(A) = \sum_{n=3}^{N} CA^{n-1} R_n \cos f_n + 1 \qquad (6)$$

The complex coefficients $\vec{a}_n = R_n e^{jf_n}$ can be determined by solving the system of equations by means of a least square method. A good fit was obtained up to the compression region for $N = 23$. This is in strong contrast to [1], which only deals with a third-order approximation to handle weak non-linearities. In previous work orthogonal Forsythe polynomials have been used to fit the data [4]. This method also gives good agreement between measured and fitted data. However, it proves to be rather troublesome to relate the Forsythe coefficients directly to the complex coefficients. Furthermore, the Forsythe method results in much longer calculation times with increasing order. This is, due to the recursive nature of the Forsythe polynomials in contrast to a least-square fit.

C. Calculating IM3

IM3 is defined as the ratio of the distortion output power at the frequency $2\omega_1 - \omega_2$ to the signal output power at the fundamental frequency ω_1. In order to obtain the components that land at these frequencies, we substitute for the input voltage $v_i(t)$ in Eq. (1) a two-tone signal $v_i(t) = A[\cos(w_1 t) + \cos(w_2 t)]$. Eq. (7) gives the generalized expression for IM3 as function of input voltage amplitude using the previously obtained complex coefficients $\vec{a}_n = R_n e^{jf_n}$.

$$IM_3 = 20 * \log \frac{|v_{o,2w_1-w_2}|}{|v_{o,w_1}|} \text{ [dBc]} \qquad (7)$$

where

$$|v_{o,2w_1-w_2}| = \left\{ \left[\sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}} \binom{n}{\frac{n+3}{2}} A^n R_n \cos f_n \right]^2 + \\ + \left[\sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}} \binom{n}{\frac{n+3}{2}} A^n R_n \sin f_n \right]^2 \right\}^{\frac{1}{2}}$$

$$\left|v_{o,w_1}\right| = \left\{ \left[a_1 A + \sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}}^2 A^n R_n \cos f_n \right]^2 + \left[\sum_{n=3}^{N} \frac{1}{2^{n-1}} \binom{n}{\frac{n+1}{2}}^2 A^n R_n \sin f_n \right]^2 \right\}^{\frac{1}{2}}$$

III. MULTIVARIABLE OPTIMIZATION FOR LOW IM3

The previously discussed mathematical model was implemented in a measurement program to optimize bias parameters in a novel distributed linear amplifier [5]. The basic concept of this amplifier is the parallel connection of several LDMOS devices with different gate widths and gate bias conditions. Fig. 2 shows the schematic circuit of this amplifier.

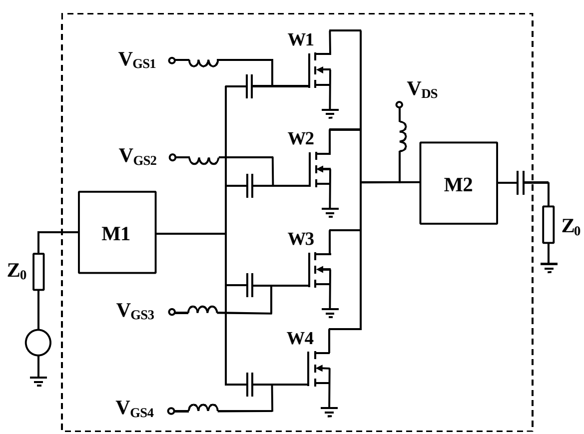

Fig. 2. Distributed amplifier design using four parallel-connected LDMOS devices with different gate widths and gate bias offset.

This amplifier has been realized and embedded in the measurement setup shown in Fig. 3. In this setup an HP 8753E network analyzer and an HP 4145B bias source is used to minimize IM3 by optimizing the four bias parameters. Fig. 4 shows a flowchart of the measurement protocol which has been implemented in HP VEE®, a tool capable to perform automatic data acquisition and data processing. First, the initial gate voltages are set equal in value and the amplifier is matched for maximum output power and linearity at the 1 dB gain compression point. For this, we used a MAURY® passive load-pull system. Then, the gate voltages are optimized until we have minimized IM3 versus input power using the new method.

The advantage of this method is that there is no need for a spectrum analyzer and separate signal generators, speeding up the measurement tremendously. For instance, a power sweep with the NWA of 51 points including the calculation of IM3 takes only 10 seconds, whereas an automated spectrum analyzer setup for the same range and amount of points takes as long as 10 minutes! It is clear that for our amplifier with four sensitive bias parameters the speed advantage works out to be very beneficial in the optimization process. The method is now in use within Philips Semiconductors DSC-N, the Netherlands in the development of base station amplifiers.

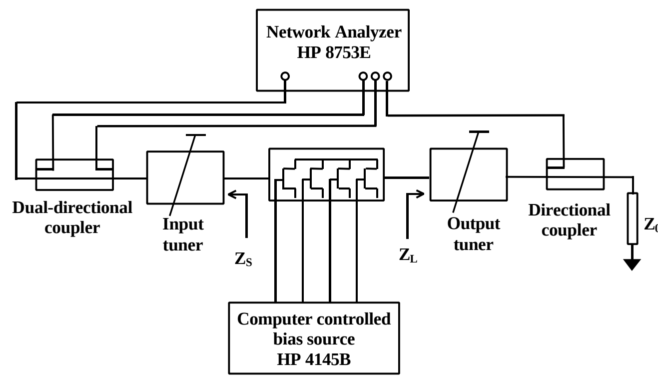

Fig. 3. Computer controlled measurement setup for optimizing gate bias voltages for minimum IM3 over a wide power range.

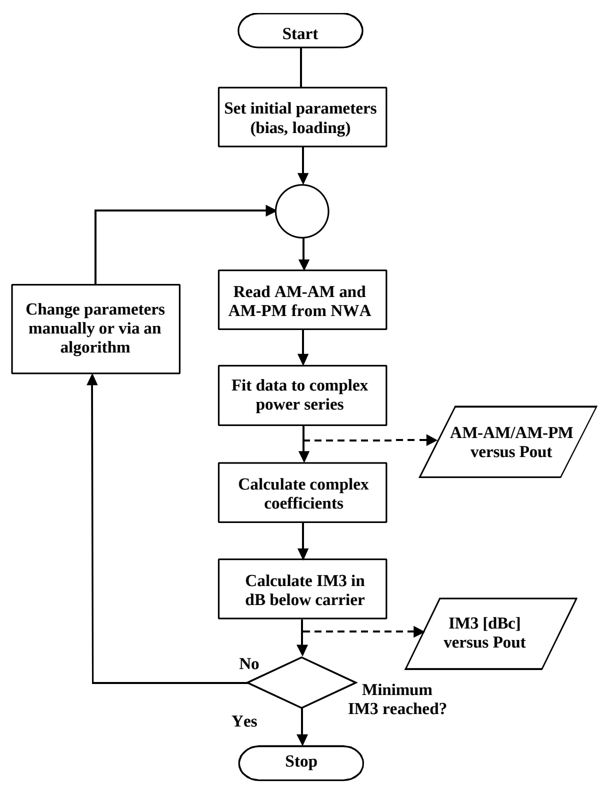

Fig. 4. Flow-chart of the measurement routine in HP VEE to minimize for IM3.

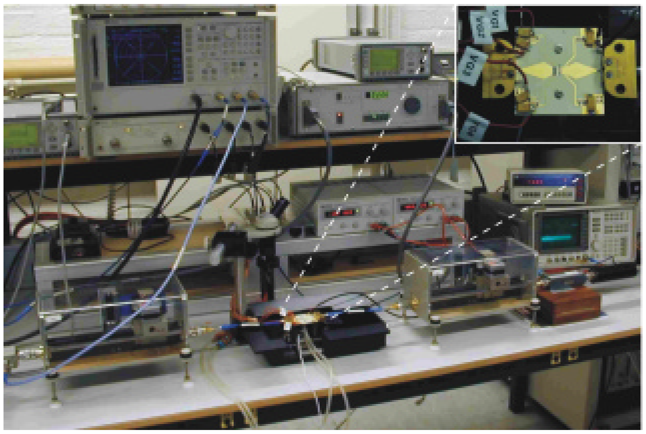

Fig. 5. Measurement setup for the optimization of the distributed amplifier, consisting of a MAURY® passive load-pull system and an HP 8753E network analyzer.

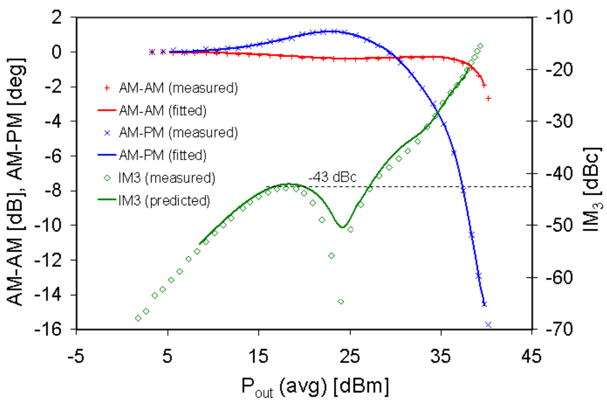

Fig. 6. Measured and predicted IM3, AM-AM, and AM-PM versus output power using the high-speed measurement protocol for equal V_{GS} of the individual LDMOS devices.

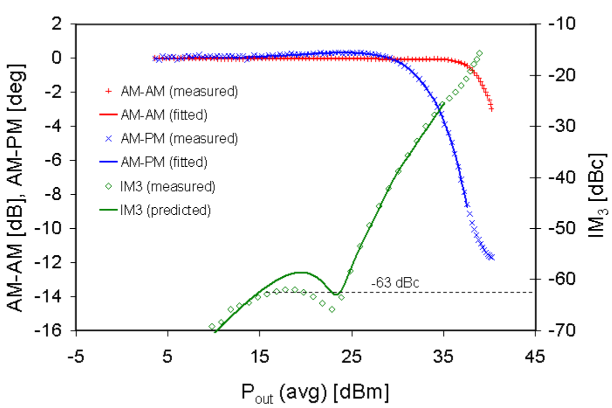

Fig. 7. Measured and predicted IM3, AM-AM, and AM-PM using the high-speed measurement protocol after optimization of V_{GS} of the individual LDMOS devices.

Fig. 5 shows a picture of the measurement setup used to optimize amplifier linearity. Fig. 6 and Fig. 7 show the result before and after optimizing IM3 versus signal power based on the AM-AM and AM-PM conversion. For verification purposes we also measured IM3 using a spectrum analyzer.

IV. Conclusion

A new measuring and optimization routine is introduced speeding up the design, tuning and testing of highly linear amplifiers involving many variable parameters. The program is based on the complex power series representation of a band-pass amplifier. By measuring the single-tone response of an amplifier in terms of AM-AM and AM-PM conversion using a NWA, IM3 can be computed and optimized. The setup proposal facilitates single instrument characterization for RF power amplifiers considering matching, gain, as well as, linearity. The improvement in speed is about 50 times compared to a conventional two-tone measurement setup.

Acknowledgement

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