Intermodulation Distortion of Third-Order Nonlinear Systems With Memory Under Multisine Excitations

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Abstract—This paper presents a methodology to compute the distortion output of a class of third-order nonlinear dynamic systems from only standard two-tone test results. Closed-form expressions are presented to compute the distortion output and metrics as adjacent channel power ratio and co-channel power ratio/noise power ratio for an arbitrary multisine with \( N \) tones. The impact of memory effects in a multisine excitation is also addressed, improving the design of RF components by a careful understanding of memory effects mechanism in real modulated signals. An experimental validation is presented to prove the proposed theory.

Index Terms—Measurements, memory effects, nonlinear systems, waveform analysis.

I. INTRODUCTION

MEMORY effects have a strong impact on the design of RF system components for the new wireless scenarios. This type of phenomenon can deeply impact any form of linearization mechanism, or even obviate its implementation in wideband systems, due to the difficulty in controlling the correspondent wideband characteristics.

For these reasons, it is imperative that a deep study is undertaken on the memory effect’s mechanisms noticed in the telecommunication systems when they are driven by real modulated excitations.

Memory effects can be divided into short and long term, with “short” and “long” referring to the time constants involved in the impulse response tail of the nonlinear dynamic system. The long-term memory time constants impact the signal’s envelope, while the short time constants affect the RF carrier. Since in a communication system the information is carried by the envelope, the understanding of the long-term memory effect mechanisms is a fundamental topic for predicting the system’s performance degradation.

Most of the research on memory effects of nonlinear dynamic systems was based on swept frequency two-tone tests [1]–[7], characterizing these dynamic responses through the variation of amplitude and phase of the observed intermodulation distortion (IMD) products.

II. RELATIONSHIP BETWEEN TWO-TONE IMD AND MULTISINE NONLINEAR DISTORTION IN A THIRD-ORDER DYNAMIC NONLINEARITY

In [11], we discuss the relationship between the two-tone excitation IMD and multisine nonlinear distortion in a third-order nonlinearity presenting memory. Here, we summarize the most...
important results from [11] to contextualize the following sections.

In order to model this phenomenon, we assumed that the nonlinear in-band response of a third-order nonlinear system presenting memory to a narrowband signal can be decomposed as the sum of a cubic polynomial direct path response with an up-converted baseband component [13]. With such a model in mind, the baseband component is demodulated from the RF signal in a second-order nonlinearity and then pressed with memory in a low-pass filter that mimics the baseband response of the nonlinear system (Fig. 1) [2], [12], [13].

According to this model, the in-band IMD transfer function for a two-tone signal is given by [2]

\[
H_3(\omega_2, \omega_2, -\omega_1) = H_{L3}(2\omega_2 - \omega_1) + [2H_{BB}(\omega_2, -\omega_1) + H_{L2}(\omega_2, \omega_2)]
\]  

(1)

where \(H_{L3}(2\omega_2 - \omega_1)\) is the third-order nonlinear transfer function arising directly from the third-order static conversion, and \(H_{BB}(\omega_2, -\omega_1)\) and \(H_{L2}(\omega_2, \omega_2)\) are the second-order nonlinear transfer functions responsible for the baseband and second harmonic signal components that will then be remixed to fall onto the system’s first zone output.

If we now consider an uncorrelated multisine excitation, i.e., when the tones do not share the same phase reference, the output distortion from a nonlinear dynamic system will be the vector addition of the several components whose amplitude and phase depend on the tone spacing.

Table I presents these components obtained for a five-tone signal (see also Fig. 2).

If the multisine excitation could be considered narrowband, i.e., if the system’s bandwidth is greater than the signal’s bandwidth, then \(H_{L3}(2\omega_2 - \omega_1)\) would be approximately constant and equal to \(K_3\). The mixing product arising from \(H_2(\omega_x, \omega_y)\), where \(\omega_x + \omega_y\) is at the second harmonic, could also be considered constant (\(K_2\)) since the relative bandwidth change with the tone spacing is very small.

Thus, for instance, we can see that the spectral regrowth tone identified as \(\omega_c\) in the output signal depends on

\[
H_3(\omega_5, \omega_5, -\omega_2) = K_3 + [2H_{BB}(\omega_5, -\omega_2) + K_2] 
\]  

(2)

and

\[
H_3(\omega_5, \omega_4, -\omega_1) = K_3 + [H_{BB}(\omega_5, -\omega_1) + H_{BB}(\omega_4, -\omega_1) + K_2].
\]  

(3)

If we manage to characterize each of those terms individually, we could get all the long-term memory effects that we need for a multisine excitation.

In order to clearly identify each of those components, a two-tone test is performed and the result is computed according to (1).

Since the most important terms are the ones that vary with tone spacing, we start by first identifying the constant part of the expression. That is done from the asymptotic behavior of \(Y(2\omega_2 - \omega_1)\) at very low-frequency separations, i.e., in the limit when \(\omega_2, -\omega_1\) tends to zero Hz. Thus, the two-tone output distortion becomes

\[
Y(2\omega_2 - \omega_1) = [K + 2F_2(\omega_2, -\omega_1)]3X(\omega_2)X(\omega_2)X(-\omega_1) 
\]  

(4)

where \(K\) is

\[
K = K_3 + K_2 + 2H_{BB}(\omega_x, -\omega_x) 
\]  

(5)

and \(F_2(\omega_2, -\omega_1)\) is the remaining term that varies with tone spacing representing the memory contribution.

This way, by changing the tone spacing, the different components can be obtained. If the tone spacing is made sufficiently small, \(K\) can also be extracted by continuity in zero separation frequency if a smooth frequency response is assumed. However, since the terms \(F_2(\omega_x, -\omega_y)\) in the multisine case are accounted in vector additions (2) and (3), we must have them characterized in both amplitude and phase.

Thus, for each frequency component, we need to solve the following equation:

\[
H_3(\omega_x, \omega_x, -\omega_y) = K + 2F_2(\omega_x, -\omega_y). 
\]  

(6)

The computation of the system nonlinear transfer functions is achieved by using higher order statistics (HOS), considering

\footnote{1}This formula is presented in a compact way contrary to what is in (5) [11]. In the current form, the complex value \(K_2\) represents the constant part of the second term of (5) [11].
a two tone as the input test signal [14]. This way, we can obtain the transfer function both in amplitude and phase.

These equations should be calculated for each tone spacing, having \( N - 1 \) different tone spacing’s involved, and thus at least a linear system of \( N \) equations should be solved, corresponding to \( N - 1 \) tone separations plus the constant \( K \).

This system of equations is built by measuring the HOS for each two-tone signal at each different tone spacing. In a test with an arbitrary number of tones, the system to be solved can be represented in matrix form as

\[
\begin{bmatrix}
1 & 0 & \cdots & 0 \\
1 & 2 & \cdots & 0 \\
\cdots & \cdots & \cdots & \cdots \\
1 & 0 & \cdots & 2
\end{bmatrix}
\begin{bmatrix}
F_2(\Delta \omega) \\
F_2((N-1)\Delta \omega)
\end{bmatrix}
= \begin{bmatrix}
H_3^{n\Delta \omega}(\cdot)/3 \\
H_3^{n\Delta \omega}(\cdot)/3 \\
H_3^{(N-1)\Delta \omega}(\cdot)/3
\end{bmatrix}
\]

(7)

where \( H_3^{n\Delta \omega}(\cdot) \) is the third-order statistic for an \( n \Delta \omega \) tone spacing, \( F_2[n\Delta \omega] \) is the second-order transfer function for an \( n \Delta \omega \) tone spacing, and \((N - 1)\) is the number of tone spacings considered.

In Section III, we show how to use this matrix for the calculation of an overall nonlinear distortion response to an \( N \)-tone multisine excitation.

### III. IMD Computation—Generalization to an Arbitrary Number of Tones

In Sections I and II, we discussed the inherent relation between multisine and two-tone responses in a nonlinear third-order system presenting memory. Now we will discuss how to quantify this relation deriving analytical results amenable to predict the multisine characteristic outcome.

In order to do so, we need to account for every nonlinear mixing product that falls on each output spectral regrowth component (Fig. 3) \( \omega_k = \omega_x + \omega_y - \omega_z \).

This task is somehow complex since it will demand an extensive account of every combination of three frequencies that will land on a specific position of the spectral regrowth. Nevertheless, this study is somehow simplified considering an equally spaced multisine since, in that case, any \( \omega_{k+1} - \omega_k = \Delta \omega \).

The main objective will thus be to obtain, in an automatic way, every mixing product for an arbitrary number of tones \( N \) for a specific spectral regrowth mixing position. This tool will allow the computation of the multisine nonlinear distortion very efficiently by exclusively calculating the number of baseband components to be analyzed.

In order to understand how to apply the proposed formulation, consider an equally spaced \( N \)-tone input signal (Fig. 3) described by \( \omega_k = \omega_0 + (n-1)\Delta \omega, 1 < n < N \). The response of a third-order nonlinear system to this input excitation is obtained combining all three tones of the input signal \( \omega_k = \omega_x + \omega_y - \omega_z \) with \( x, y, \) and \( z \) representing the input spectral positions.

The output \( \omega_k \) will be given by

\[
\text{IMD}(\omega_k) = H_3(\omega_x, \omega_y, -\omega_z)X(\omega_x)X(\omega_y)X(\omega_z)
\]

(8)

where

\[
H_3(\omega_x, \omega_y, -\omega_z) = [K + F(\omega_x - \omega_z) + F(\omega_y - \omega_z)]
\]

(9)

as was previously seen and as in [11].

Despite the large number of combinations that fall in \( \omega_k \) for a multisine excitation, a precise account of those combinations is needed before proceeding with the overall power calculation of the mixing product.

In [7], the number of different combinations that fall on each position was calculated for a memoryless nonlinearity. The different arrangements were then considered for the ones falling within the input signal components and, in that case, there were some contributions that were correlated with the input, and some others uncorrelated. Beyond these, there were some others that fall at the adjacent channel, where two types of mixing products were also considered: the ones presenting equal \( \omega_x = \omega_y \) and the ones where \( \omega_x \neq \omega_y \).

However, in that case, due to memoryless nonlinearity, no importance was given to the fact that each \( \omega_k \) is affected by a different \( H_3(\omega_x, \omega_y, -\omega_z) \). This coefficient is dependent on a binomial function that depends on two baseband values \( (\omega_x - \omega_z) \) and \( (\omega_y - \omega_z) \), responsible for the nonlinear system’s memory effects. In fact, we are considering that \( H_3(\omega_x, \omega_y, -\omega_z) \) can be expressed as

\[
\Im(\alpha \Delta \omega, \beta \Delta \omega) \equiv H_3(\omega_x, \omega_y, -\omega_z)
\]

(10)

where

\[
\alpha \Delta \omega = \omega_x - \omega_z \\
\beta \Delta \omega = \omega_y - \omega_z
\]

Thus, within this model, accounting for the system’s nonlinear dynamic effects means to calculate each different \( \alpha \) and \( \beta \) values that falls on each \( \omega_k \) position for an arbitrary number of tones \( N \).

In order to obtain closed formulas to calculate these different arrangements, we will divide the nonlinear mixing products in three different classes, i.e.: 1) the ones falling onto the adjacent-
channel; 2) the co-channel products that are correlated to the signal; and 3) the co-channel uncorrelated products.

The measurement procedure starts with a set of swept frequency two-tone tests covering \( N - 1 \) tone spacings, and recording the amplitude and phase values of each IMD component [14], [15]. Using (7), the function of the tone spacing \( F(n\Delta\omega) \) is then computed. With these functions, it is then possible to calculate each distortion component that falls at a specific frequency position according to (9). Finally, the various components are added together, according to the characteristics of each mixing product. The power of the correlated distortion products must be accounted for as the average power of the addition of phasor quantities, while the power of the uncorrelated ones can be computed by the direct addition of the powers of the individual components.

### A. Adjacent-Channel Mixing Products

For a given adjacent spectral position \( \omega_k \), we must account for all different values of \( \alpha \) and \( \beta \). In that respect, and in order to simplify the calculation, we have developed a matrix scheme, shown in Fig. 4, where for each spectral regrowth, \( \omega_k \) (in the illustrated case for a five-tone multisine excitation), the binomial combinations were calculated—\( \alpha \) and \( \beta \) pairs. The combinations marked with an “\( x \)” correspond to different \( \alpha \) and \( \beta \) values, while the ones marked with an “\( o \)” correspond to equal \( \alpha \) and \( \beta \). This separation is important since the amplitude of mixing products with equal \( \alpha \) and \( \beta \) must be multiplied by the multinomial coefficient 3/8, while the ones with different \( \alpha \) and \( \beta \) must be scaled by 3/4.

By expanding these matrices for an increasing number of tones, from 2 to \( N \), and then adding up the corresponding mixing products, we were able to derive a formula that automatically generates the output power of the component at \( \omega_k \) and \( P(\omega_k) \) for each pair of \( \alpha \) and \( \beta \) values. The obtained formula is

\[
P(\omega_k) = \sum_{\alpha=0}^{N-1} \sum_{\beta=0}^{N-1} \left\{ \frac{3}{8} \left[ (n-(k-a)) \Delta\omega \right. \right. \left. \Delta\omega \right\}^2 A^6 \tag{11}
\]

where \( A \) is calculated for all values of \( \Delta\omega \) spanning from 0 to \((N-1)\Delta\omega \). This will then be stored into a

**Fig. 4.** Different matrices of spectral regrowth mixing products that fall on the frequency position \( \omega_k \) for a five-tone excitation.

where \( \text{mult} \) corresponds to the multinomial coefficient [6] that values 3/8 if \( \omega_x = \omega_y \) and 3/4 if \( \omega_x \neq \omega_y \), and \( k \) is the spectral regrowth tone accordingly to Fig. 3. The amplitude of each multisine component is represented by \( A \) and is a constant.

### B. Co-Channel Signal-Correlated Mixing Products

As was explained earlier, the co-channel mixing products are divided into signal-correlated and uncorrelated components. The correlated ones are the products that obey the constraint \( \omega_k = \omega_x + \omega_y - \omega_z = \omega_x \) when two involved frequencies are equal.

Now \( \beta = 0 \) at all times, and \( \alpha \) spans from 1 to \( N - 1 \). Thus, the output power at each co-channel mixing product becomes

\[
P(\omega_k) = \sum_{a=1}^{N} \left\{ \frac{3}{8} \left[ (a-b) \Delta\omega, (i) \Delta\omega \right] \text{mult} \right\}^2 A^6 \tag{12}
\]

where \( b \) is the frequency position of the desired mixing product according to Fig. 3. In this case, each contribution is accounted for in a vector addition since the distortion products are all correlated to each other.

### C. Co-Channel Signal-Uncorrelated Mixing Products

Since we are assuming that there is no phase correlation between any of the input tones, the generation mechanism of signal-uncorrelated co-channel mixing products is similar to the one responsible for the ones falling at the adjacent channel.

Indeed, as any \( \omega_k = \omega_x + \omega_y - \omega_z \) input frequency combination has a phase \( \phi_1 + \phi_2 + \phi_3 \) that is different from one of any other combination (as long as \( \omega_3 \) is different from both \( \omega_1 \) and \( \omega_2 \)), the mixing products that fall at a certain \( \omega_k \) will all be uncorrelated. Therefore, the final result of the power at each \( \omega_k \) position is, on average, equivalent to the addition of each individual component’s power. This implies that, while in (12) the output power \( P(\omega_k) \) was given as the square of the voltage sum, it must now be given as the sum of the squares of voltage

\[
P(\omega_k) = A^6 \sum_{a=1}^{N-1} \sum_{b=0}^{N-1} \left\{ \frac{3}{8} \left[ (a-b) \Delta\omega, (i) \Delta\omega \right] \text{mult} \right\}^2 \tag{13}
\]

where

\[
\text{vertex} = -\left\lfloor \frac{b}{2} \right\rfloor - 1 \tag{14}
\]

and \( \lfloor x \rfloor \) is a function that rounds the elements of \( x \) to the nearest integers towards minus infinity.

Thus, if the overall output nonlinear distortion is to be calculated from a two-tone measurement, the procedure is the one presented below. First, a two-tone measurement is made and the corresponding \( F(x) \) is calculated for all values of \( \Delta\omega \) spanning from 0 to \((N-1)\Delta\omega \). This will then be stored into a
matrix. The next step involves the calculation of \( Y(\alpha \Delta \omega, \beta \Delta \omega) \) for all binomials at each \( \omega_k \) position, and, using \( \alpha \) and \( \beta \), obtaining the value of the function \( F_2(\alpha \Delta \omega) \) from the previous matrices. The solution of the nonlinear distortion to a multisine excitation then becomes straightforward.

IV. IMPACT OF MEMORY IN THE IMD OF MULTISINE SIGNALS

Here, we aim to analyze the impact of memory effects on the output of a nonlinear system presenting memory when driven by a random multisine signal.

For performing this task, the formulation presented above will be very important since by accounting for the values of \( \alpha \) and \( \beta \), we can infer information of what are the most important characteristics of the baseband that impact the overall nonlinear distortion.

Beginning with the adjacent-channel scenario, we calculate the number of times that each binomial function \( Y(\alpha \Delta \omega, \beta \Delta \omega) \) appears for each \( \omega_k \). This result is visible in Fig. 5.

An interesting observation from this analysis is that the main contribution to the adjacent-channel power is at the middle of the figure, which means that the binomial \( \alpha = N/2 \) and \( \beta = N/2 \), \( Y[(N/2)\Delta \omega, (N/2)\Delta \omega] \) is the most important term to be considered in order to minimize the adjacent channel power ratio (ACPR) figure-of-merit [6].

The same analysis was done for the co-channel distortion components, as is presented in Fig. 6. In this case, the most important components, i.e., having stronger impact in the co-channel distortion, are the baseband components of lower frequency.

Considering now the correlated co-channel case (Fig. 7), we can conclude that the most important components are again the baseband components of lower frequency. Another interesting aspect is that all the correlated distortion components have at least one dependence with zero tone spacing, corresponding to the dc value of the baseband filter.

In order to validate these hypotheses a computer-aided design (CAD)/computer-aided engineering (CAE) simulation was run by using different baseband filters.

The operating bandwidth of 1.7 MHz was split into three bands with equal bandwidth, as shown in Fig. 8.

The three filters considered are: 1) a low-pass filter for the baseband components; 2) a filter reinforcing the middle of the baseband; and finally 3) a filter accounting for the remaining part of the baseband.
The input signal used in the simulation is an uncorrelated multisine of 100 tones. To simulate a system with memory, a direct implementation of the model of Fig. 1 was done.

Fig. 9 presents the obtained results for the multisine output and the output obtained by our formulation. The ACPR and co-channel power ratio (CCPR) [6], were computed and the corresponding values are 41.2 and 24.7 dB. A good agreement is visible at the spectral regrowth.

Fig. 10 presents the obtained results for a passband baseband filter and with all the other configurations remaining unchanged. The obtained ACPR and CCPR results are 38.2 and 32.6 dB, respectively.

Finally, a last test was run with the higher frequency passband filter. In this case, the ACPR and CCPR results are 40.9 and 34.9 dB (Fig. 11).

Based on previous observations, we can conclude that the low-pass and higher band bandpass filters impose a similar ACPR. On the contrary, the passband filter has a higher contribution to the adjacent distortion degrading the ACPR values. This validates our previous hypothesis based on the formulation developed.

The same analysis was done for the co-channel distortion. In this case, the most important components are the ones with lower tone spacing, as can be seen in the lower CCPR values for the low-pass filter, and a decrease for the other two subsequent filters.

V. EXPERIMENTAL RESULTS

An experimental test was also run in order to validate the theory presented above. The setup comprises a low-power...
amplifier based on the ATF 55143 pseudomorphic HEMT (pHEMT) device biased in classes A and B.

A drain bias network was designed to present a varying baseband response in order to press memory effects in the output response.

A memoryless network was also designed to prove the validity of the presented algorithm in the standard memoryless situation.

The test was run at a central frequency of 900 MHz with a tone spacing of 20 kHz. The input signal is composed by 20 tones leading to a 400-kHz bandwidth signal obtained from an arbitrary waveform generator (AWG). A record of 1000 waveform segments with random phase was used as a random multisine signal to predict the output of the system to a narrowband Gaussian noise input. The measurement procedure comprised the synchronous acquisition of both the input and output signals by a high-speed sampler and the post-processing of the data in order to get system output [14].

In Fig. 14, we present the $S_{11}$ of the biasing network that was applied to the drain of the device. As can be seen, the impedance of the memoryless bias is almost constant and close to a short circuit. The bias presenting memory has varying impedance with a resonance in the middle of the band.

A first test was done in the class A operation with the bias network that presents a varying baseband characteristic.

Fig. 15 presents the computed output distortion obtained by our method and the measured results. A good agreement can be noticed between the computed values and measured output. The increasing error observed in the low-power tones was attributed to the fifth-order distortion that is always present in a real system, but was not accounted for in our third-order analysis.

Fig. 16 presents the obtained results for the memoryless case. A good agreement can be noticed, even in this ideal case, indicating the applicability of the method.
To test the practical ability to predict asymmetry (strong evidence of long-term memory effects), the amplifier was biased in class B. The output is presented in Fig. 17.

As can be seen, the predicted response and measured output are in perfect agreement. The memory effects can also be noticed in the shape of the correlated and uncorrelated distortion components falling in the co-channel band.

VI. CONCLUSIONS

This paper has presented an analytical methodology to compute the output distortion of a certain class of dynamic third-order systems to a multisine excitation using only two-tone tests. A qualitative explanation of the impact of the baseband filter in the co-channel and adjacent channel has also given, permitting a better bias network design for power amplifiers. The presented methodology has been tested against simulated and laboratory experiments, demonstrating its validity even in the case of adjacent-channel asymmetric responses. The computed results enable the computation of multitone figures-of-merit as ACPR and CCPR from only a small set of two-tone measurements, which are standard in every RF laboratory. This study is a step forward toward the understanding of the memory generation mechanisms and in the extrapolation of the usual standard RF test results to the prediction of the dynamic system’s output to a multisine signal excitation.

REFERENCES


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