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A Swept-Sine Pulse Compression Procedure for an Effective Measurement of Intermodulation Distortion

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Abstract—The ability to produce accurate models of nonlinear systems has grown enormously thanks to the increase in the available computing power, thus making possible to use such models in a wide range of applications. Adopting these techniques enables to design much more efficient systems, but also increases the need for accurate and straightforward methods to monitor distortion levels, even out of the testing laboratories. In this context, efficient estimators of total harmonic distortion, harmonic distortion of order n , and intermodulation distortion are needed, which are among the most widely adopted indicators of non-linearity. In this paper, we focus on the intermodulation distortion, which is a robust feature indicator within the whole systems operating bandwidth, whose measurement is based on the effects produced by the nonlinear system when a composite signal feeds its input. We propose an efficient procedure to measure the intermodulation distortion allowing the simultaneous evaluation of harmonic distortion of order n and of total harmonic distortion. The proposed technique relies and extends the synchronized exponential swept-sine technique for the characterization of non-linear systems by introducing a suitable double exponential chirp input signal. The method estimates all the distortion level indicators by processing the response to a single, short-duration signal, showing a moderate computational cost and low sensitivity to ambient noise. These advantages make the application of the proposed method extremely rapid and convenient, even in the measurement of operating field devices. Tests have been carried out both in simulated and real-world scenarios, confirming the proposed approach validity.

Keywords—Pulse compression, intermodulation distortion measurement, non-linear device, Hammerstein system

I. INTRODUCTION

The availability of efficient procedures for the characterization of a system from the point of view of its non-linear behavior has become an aspect of ever growing importance in the recent years. In fact, the increased computing power available even on-board portable devices has enabled a large-scale use of techniques modeling real world systems and their non-linear characteristics. These techniques allow the degree of non-linearity in reproduction devices to be controlled, so as to adequately compensate it, for example in audio reproduction

systems [1]; or to take into account the non-linearity during the design process. The latter case happens in the design of switch-mode power supplies, where the use of ferrite core power inductors operating in partial saturation allows both volume and weight of the power supply itself to be significantly reduced [2], [3]. In the field of non destructive testing and evaluation, the use of non-linear techniques is becoming widespread, allowing the inspection capability of particularly critical structures to be improved [4], [5]. The common use of these techniques has considerably widened the number of cases in which it is extremely useful to have reliable estimators of the system non-linearity degree available. The presence of these advanced modeling techniques requires thus the development, in parallel to them, of efficient methods to evaluate the quality of the overall result. This is achieved by measuring the parameters that give a quantitative measure of the degree of non-linearity present in the system under test. Such estimators can be very effective if they exploit the modeling and identification techniques of non-linear systems. Although theoretically possible, a model describing the non-linear behavior of a real device, which starts from its physical description, cannot be adopted in most cases of practical relevance. As a consequence, phenomenological models are used that characterize the non-linear link between input and output [6], [7], [8]. A number of approaches to non-linear modeling are available in the technical literature: a functional representation of non-linear systems was first studied by Volterra, who proposed an explicit input/output relation when both input and output are bounded [9], [10]. The Volterra series consists, however, of an infinite number of terms and therefore in practice its truncation is necessary. The definition of the convergence criteria of the series associated with a given system is a major problem, and there is no general criterion available to define the convergence range of the series. Moreover, even if the order of the Volterra series is truncated, the number of coefficients needed to define the model quickly becomes very large as the degree of non-linearity increases: a system that has a memory of N samples requires N^p coefficients for the definition of the kernel of order p . Consequently, only systems with a moderate degree of non-linearity can be represented in practice by adopting the Volterra series model. Several techniques have been developed starting from the one proposed by Volterra. In particular, the so-called best linear approximation (BLA) framework [11], [12], assumes that the system under test can be well-described by a Volterra series, but it does not reconstruct the Volterra series itself. Instead, the frequency response function of the BLA is estimated together with noise

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and non-linear distortion variances. Research studies have been focused on simplified Volterra series expansion models [13]. Among them, the Hammerstein and Wiener models rely on splitting the linear dynamic part from the static, non-linear one. The Hammerstein model consists of a cascade of memoryless non-linearity and a linear dynamic system [14]. In addition, the generalized version of the Hammerstein model of order N_H has been defined: it is composed of N_H parallel branches, each of them consisting of a static non-linearity followed by a linear filter [15], see Figure 1. Although less general than Volterra series expansion, the Hammerstein model provides an accurate modeling of even strong non-linearities, using a reduced number of parameters [6], [8]. Several efforts have been made in the literature to develop an algorithm for identifying such models, mainly exploiting a suitable input swept-sine signal [15], [16], [17], [18].

The Wiener model implements a different simplified version of Volterra model. In this case, the static non-linearity is placed after the linear filters [19], [20], [21], [22], [23].

It is not always necessary to have a complete model of the non-linear system: it may be useful to have numerical estimators available that give a synthetic and meaningful measure of the non-linearity degree of the device, whose values can be measured in a relatively simple and repeatable way. These estimators are also useful to verify the quality and stability of operation of non-linear system models. These evaluation parameters must be associated with input signals of defined characteristics. In addition, they must also be representative of the signals with which the system operates in reality. Sinusoidal signals, combinations of sinusoidal signals, random noise with a speech-shaped spectrum are among the signals used as inputs to identify quantities such as the Harmonic Distortion (HD), the Total Random Distortion (TRD), the InterModulation Distortion (IMD) which, for different aspects, provide indications on the degree and type of non-linearity introduced by the considered system. Indeed, procedures for an efficient evaluation of these synthetic estimators of devices' non-linearity are essential to verify the quality of the models and to control their non-linear behavior. Among the synthetic estimators, the Total Harmonic Distortion (THD), the Harmonic Distortion of order k (HD_k) and the IMD are the most widely used for non-linearity estimation [18], [23].

Different measurement techniques have been defined for both HD and IMD , based on the use of gliding tones (linear chirp signals), or on step-by-step sinusoidal tones, choosing the speed of variation (in the former case) and the frequencies and duration of the tones to obtain the necessary measurement accuracy, and measurement times compatible with practical needs. For IMD , techniques have been defined that consider a fixed tone and a variable tone (modulation distortion of the second and third order) as input signals, or techniques that consider both the tones variable (difference frequency distortion). In this work, we consider the latter technique, which takes into account both tones to be variable and defined at symmetrical frequencies with respect to the nominal one, and we refer to this measure as IMD .

The different synthetic parameters estimated this way, highlight different aspects that are useful to characterize the non-

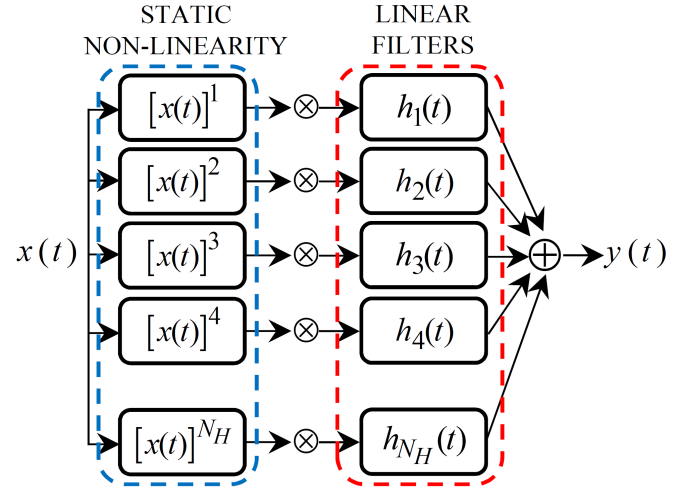


Fig. 1. Generalized polynomial Hammerstein model scheme. The model is composed of N_H parallel branches, each of them consisting of a static non-linearity followed by a linear filter.

linear behavior of the device, but this implementation is not always compatible with practical constraints. For this reason, starting from these assumptions and from previous results on THD and HD_k estimation [18], [23], a novel approach for IMD prediction has been proposed in [24]. The proposed technique relies and extends the synchronized exponential swept-sine technique for the characterization of non-linear systems by introducing the use of a suitable input signal, which consists of the sum of two equal exponential swept-sine signals shifted by a proper time interval. This way, an estimate of the distortion related to harmonics combination is obtained, providing an effective procedure for IMD estimation, as reported in [24]. Taking into account these preliminary results, in the present paper we propose a detailed analysis of the characteristics and limitations of the proposed estimation technique and identify the procedures to overcome them. We show that the measurement accuracy can be adjusted according to the needs by changing the parameters that characterize the excitation signals. Finally, we show how one can get an estimate of HD_k , THD and IMD from a single measurement through appropriate post processing techniques applied on the system response. Several experiments using simulated and real-world devices are reported, analyzing pros and cons of the proposed solutions.

The paper is organized as follows. The proposed approach is presented in Section II that illustrates the procedure for IMD computation by exploiting a suitable input signal and a generalized Hammerstein model. Then, the validation of the proposed method is discussed in Section III showing experimental measurements both for simulated and real-world devices. Finally, concluding remarks and future works are reported in Section IV.

II. PROPOSED APPROACH

A. Chirp-based identification of the Hammerstein model

The structure of the Hammerstein model of a non-linear system is represented in Figure 1. A powerful method for its identification has been proposed, which makes use of an exponential swept-sine signal to excite the non-linear system [17], [18]. In (1), a general expression for an exponential swept-sine signal is given:

$$x(t) = \sin(\Phi(t)) \times [u(t) - u(t - T)] = C_{exp}(t, f_{MIN}, f_{MAX}, \Phi_0) \quad (1)$$

where $\Phi(t)$ is the instantaneous phase, T is the time duration of the swept-sine, f_{MIN} and f_{MAX} are the lower and the upper limits respectively of the swept frequency range, *i.e.*, the bandwidth B is equal to $B = f_{MAX} - f_{MIN}$, Φ_0 is the initial phase, and $u(t)$ is the Heaviside step function. The exponential swept-sine instantaneous frequency f_{inst} is defined as $f_{inst}(t) = \frac{1}{2\pi} \frac{d\Phi(t)}{dt}$ and $\Phi(t)$ is defined such that its time derivative follows the exponential law $f_{inst}(t) = k_0 e^{\frac{t}{L}}$. The constant L is defined as $L = \frac{T}{\ln(f_{MAX}/f_{MIN})}$ and describes the time variation rapidity of f_{inst} . It follows that if $f_{inst}(t)|_{t=0} = f_{MIN}$ is wanted, then $k_0 = f_{MIN}$. Thus, the instantaneous frequency expression $f_{inst}(t) = f_{MIN} e^{\frac{t}{L}}$ is obtained after having set both L and f_{MIN} . The function $x(t) = \sin(2\pi f_{MIN} L [e^{\frac{t}{L}} - 1] + \Phi_0)$ implements the exponential chirp C_{exp} . The corresponding instantaneous frequency function $f_{inst}(t)$ is strictly monotonic, thus its inverse can be used to derive the time instant at which $f_{inst}(t)$ equals the generic frequency value f^* :

$$t^* \ni f_{inst}(t^*) = f^* \rightarrow t^* = L \ln\left(\frac{f^*}{f_{MIN}}\right). \quad (2)$$

Consider now the filter matched with the excitation, whose impulse response $\Psi(t, f_{MIN}, f_{MAX}, \Phi_0)$ is such that its convolution with the exponential swept-sine signal C_{exp} outputs an approximated Dirac's delta function $\delta(t)$, as for (3):

$$C_{exp}(t, f_{MIN}, f_{MAX}, \Phi_0) * \Psi(t, f_{MIN}, f_{MAX}, \Phi_0) \simeq \delta(t). \quad (3)$$

It has been shown in [15], [17], [18] that, if the non-linear system response to the exponential swept-sine signal is filtered through the matched filter, and the swept-sine signal fulfills the phase alignment conditions shown in [25], [26], then this filter works in an equivalent way on both the signal and its harmonics. In [25], authors show that the phase alignment condition is fulfilled if the exponential swept-sine signal $x_{exp}(t) = \sin[2\pi f_{MIN} L (e^{\frac{t}{L}} - 1) + \Phi_0]$ can be written in the form $x_{exp}(t) = \sin[2\pi f_{MIN} L e^{\frac{t}{L}}]$. In this case the phase alignment condition is fulfilled independently of the extreme frequencies f_{MIN} and f_{MAX} of the chirp signal and of its duration T . The initial phase can be constrained by appropriately choosing the value of the product $f_{MIN} L$. Within the validity limits of the Hammerstein model, the fulfillment of these conditions implies that the output of the matched filter is in the form of a sequence of impulsive functions $g_k(t)$ related to the different harmonics. This is because there is a close relation between the harmonics of the frequency f^* reached by the signal at time t^* (the harmonics whose frequencies are kf^*) and the same

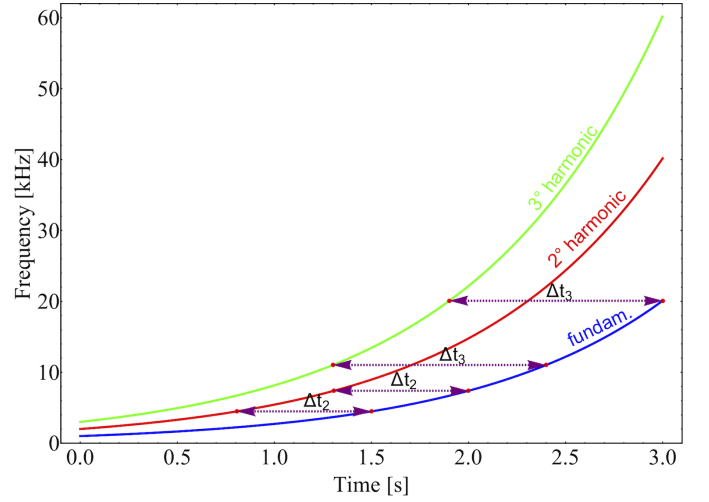


Fig. 2. Exponential chirp signal in a non-linear device: time behavior of the fundamental frequency and of the first two harmonics. The time distance between the frequency of a harmonic and the same frequency seen as fundamental frequency is independent from the considered frequency.

frequencies kf^* present in the undistorted exponential swept-sine signal at a time distance ΔT_k from t^* . The time shift ΔT_k is equal in value to $\Delta T_k = L \ln(k)$ and it is related solely to the order of the harmonic. Consequently, the k -th harmonic of the signal at time t equals the fundamental of the same signal at time $t + \Delta T_k$:

$$C_{exp}(t, kf_{MIN}, kf_{MAX}, k\Phi_0) = C_{exp}(t + \Delta T_k, f_{MIN}, f_{MAX}, \Phi_0). \quad (4)$$

This property is depicted in Figure 2 and it is the basis of the identification procedure of the Hammerstein model described in [15], [17], [18]: each one of the $g_k(t)$ functions is associated with a specific harmonic generated by the non-linear system, and the time positions of each $g_k(t)$ in the response of the matched filter show the order of the corresponding harmonic. In other words, the $g_k(t)$ function associated with the k -th harmonic is shifted earlier in time by the amount ΔT_k with respect to the $g_0(t)$ function, associated with the fundamental.

The $g_k(t)$ functions can be thus obtained by simply windowing the matched filter response in a time interval associated with the k -th harmonic. Finally, the $h_k(t)$ functions identifying the Hammerstein model can be retrieved from the $g_k(t)$ through a linear transformation.

B. Chirp based estimate of the harmonic distortion

It is possible to estimate the frequency behavior of THD and HD_k through $g_k(t)$ [18], [23]. In fact the response of the Hammerstein model can be written in the time domain as:

$$y(t) = \sum_{k=1}^{N_H} x(t)^k * h_k(t). \quad (5)$$

The input $x(t)$ is a sine swept signal, and its powers can be reformulated in terms of its harmonics by using the Chebyshev polynomials of the first kind $T_n[\cdot]$. We can write:

$$\begin{aligned} y(t) &= \sum_{k=1}^{N_H} [C_{exp}(t, f_{MIN}, f_{MAX}, \Phi_0)]^k * h_k(t) = \\ &= \sum_{k=1}^{N_H} \sum_{i=1}^{N_H} ([A_c]^{-1})_{i,k} C_{exp}(t, i f_{MIN}, i f_{MAX}, i \Phi_0) * h_k(t) = \quad (6) \\ &= \sum_{k=1}^{N_H} [C_{exp}(t, k f_{MIN}, k f_{MAX}, k \Phi_0)] * g_k(t) \end{aligned}$$

where the $g_k(t)$ are related to the kernels $h_k(t)$ of the Hammerstein model through the linear transformation

$$g_k(t) = \sum_{i=1}^{N_H} ([A_c]^{-1})_{i,k} h_i(t) \quad (7)$$

and $[A_c]$ is the matrix of coefficients of Chebyshev polynomials of the first type [18] and the indices, in the adopted notation, indicate the trasposition of the matrix. Equation (6) can be written in the frequency domain as:

$$Y(\omega) = \mathcal{F}\{y(t)\} = \sum_{k=1}^{N_H} G_K(\omega) \mathcal{F}\{C_{exp}(t, k f_{MIN}, k f_{MAX}, k \Phi_0)\} \quad (8)$$

where $\mathcal{F}\{\cdot\}$ is the Fourier transform of the k -th harmonic of the input signal and $G_K(\omega)$ are the Fourier Transforms of the impulsive functions $g_k(t)$. As (8) shows, if the input is a sinusoidal signal, *i.e.*, a special case of an exponential swept-sine signal, the output signal consists of the superposition of its harmonics; the amplitude of the k -th harmonic is given by $|G_k(\omega)|$ for each angular frequency ω_0 . Thus, the $HD_k(\omega)$ and the $THD(\omega)$ can be estimated from the retrieved $G_k(\omega)$:

$$HD_k(\omega) = \frac{|G_k(k\omega)|}{\sqrt{\sum_{n=1}^{N_H} |G_n(n\omega)|^2}} \quad (9)$$

$$THD(\omega) = \frac{\sqrt{\sum_{n=2}^{N_H} |G_n(n\omega)|^2}}{\sqrt{\sum_{n=1}^{N_H} |G_n(n\omega)|^2}}. \quad (10)$$

C. Chirp-based estimate of the intermodulation distortion

A method for estimating the intermodulation distortion is here proposed in the same framework. We start by describing the standard technique for *IMD* estimation as detailed in IEC 60268 [27]. This is based on the observation that if two sinusoidal tones at angular frequencies ω_A and ω_B , with $\omega_A > \omega_B$, are input to a non-linear system, the output signal will contain, at each time t , harmonic components at integer multiples of ω_A and ω_B of the two tones. Moreover, the output signal will contain also harmonic components at angular frequencies that are linear combinations (sum or difference) of the individual harmonic tones $\omega = \pm n_A \omega_A \pm n_B \omega_B$ (with n_A and n_B integer values). As the system is band limited, the largest part of the harmonics produced by sum fall outside the useful band. Consequently, the harmonics produced by difference are

considered, particularly those for which $n_A = n_B = k$, for $\{k = 1, 2\}$, being the most relevant as they show the highest energy content. According to IEC 60268, a quantitative evaluation of the intermodulation produced by the non-linear characteristics of the system can be measured by considering the frequency behavior of the component $\omega_A - \omega_B$, the difference frequency distortion of the second order (*i.e.* when $k = 1$), when two signals of equal amplitude are input to the device. A narrow bandpass filter is adjusted for the angular frequency $\omega_A - \omega_B$ to measure the Root Mean Square (RMS) value of the component at that frequency. The value $\omega_A - \omega_B = 2\pi \times 80$ [rad/s] is suggested for the standard procedure. Denoting as P_{f_i} the sound pressure at frequency f_i , the quantity:

$$IMD = \frac{P_{f_A - f_B}}{P_{f_A} + P_{f_B}} \times 100 \quad (11)$$

is evaluated as the second order difference frequency *IMD* and is assigned to the frequency value of $[\omega_A + \omega_B]/2$. The same measurement is repeated for different frequencies and amplitudes of the input tones to produce a result that is a function of applied voltages and frequency. In the more general case, for non-audio/speech applications, (11) can be formulated equivalently by replacing the pressure P_{f_i} with the amplitude of the harmonic component of the signal measured at the output of the non-linear system at the specified frequency.

In the present paper, a pulse compression-based alternative procedure to estimate the complete frequency behavior of the second order *IMD* by means of a single measurement is proposed. In fact, the *IMD* can be measured by feeding the input of the non-linear system with a signal $x_D(t, \Delta T, f_{MIN}, f_{MAX}, T)$ consisting of the sum of two exponential swept-sine signals $x_{exp}^A(t)$ and $x_{exp}^B(t)$, sharing the same rate of exponential increase in frequency L , but shifted in time by the quantity ΔT (double chirp signal):

$$\begin{aligned} x_D(t, T, \Delta T, f_{MIN}, f_{MAX}, L, \Phi_0) &= \quad (12) \\ &= x_{exp}^A(t) + x_{exp}^B(t) = \\ &= x_{exp}^A(t) + x_{exp}^A(t + \Delta T) = \\ &= \left\{ \sin(2\pi f_{MIN} L [e^{\frac{t}{L}} - 1] + \Phi_0) + \right. \\ &\quad \left. + \sin(2\pi f_{MIN} L [e^{\frac{t+\Delta T}{L}} - 1] + \Phi_0) \right\} \cdot \\ &\quad \cdot [u(t) - u(t - (T - \Delta T))]. \end{aligned}$$

The ratio between the instantaneous frequencies of the two exponential swept-sine signals summed at the input is constant, thus not dependent from a given time instant t :

$$\frac{f_{inst}^B(t)}{f_{inst}^A(t)} = \frac{f_{MIN} e^{\frac{(t+\Delta T)}{L}}}{f_{MIN} e^{\frac{t}{L}}} = e^{\frac{\Delta T}{L}}. \quad (13)$$

The non-linearity of the system will generate harmonic components in the output signal. Thus, there will be harmonics of the instantaneous frequency of each of the two exponential swept-sine signals and linear combinations (sum and difference) of these harmonics. The instantaneous frequency of the components generated by the non-linearity will vary over time following an exponential law, with the same growth rate L of the two chirps summed at the input. We verify this

aspect for the intermodulation difference component of the second order, a signal that we indicate hereinafter with C_{exp}^{DIFF} , whose instantaneous frequency is the difference of the two fundamentals. It is simple to show that the argument of the signal has a component at angular frequency $\omega_1(t) - \omega_2(t)$ which can be expressed in the form $2\pi f_{MIN} L e^{\frac{t}{L}} (e^{\frac{\Delta T}{L}} - 1)$, i.e., it is proportional to $e^{\frac{t}{L}}$. It follows that the time trend of its instantaneous frequency $f_{inst}^A(t) - f_{inst}^B(t)$ is given by:

$$f_{inst}^A(t) - f_{inst}^B(t) = f_{MIN} (e^{(t+\Delta T)/L} - e^{t/L}) = f_{MIN} e^{t/L} (e^{\Delta T/L} - 1). \quad (14)$$

As shown in (14), C_{exp}^{DIFF} is a chirp signal whose instantaneous frequency, produced by the difference frequency distortion component of the second order, evolves in time following an exponential law whose growth rate is L , i.e. the same of each one of the two exponential swept-sine signals summed at the input. In particular, the instantaneous frequency equals the one of the first exponential swept-sine multiplied by the factor $(e^{\frac{\Delta T}{L}} - 1)$. This factor can be expressed in terms of a delay τ , imposing that $(e^{\frac{\Delta T}{L}} - 1) = e^{\frac{\tau}{L}}$, which gives:

$$\tau = L \times \ln(e^{\frac{\Delta T}{L}} - 1), \quad (15)$$

where τ is the time difference between the current instant and the one at which a frequency equal to the difference $f_{inst}^A(t) - f_{inst}^B(t)$ is found in the undistorted exponential swept-sine signal, and ΔT is the time shift between the two exponential swept-sine signals at the input. The intermodulation component $f_{inst}^A(t) - f_{inst}^B(t)$ grows exponentially with the same rate L of the two exponential chirps summed at the input. Thus an impulsive function will be produced at the output of the matched filter if the phase alignment condition is met as well. As already pointed out, the pulse compression method can be in fact successfully applied when the instantaneous phase of the exponential swept-sine signal and the one of its harmonics are aligned. It is simple to show that the same condition is satisfied for the intermodulation signal.

In fact, the instantaneous phase of the difference frequency swept-sine signal can be written as:

$$\begin{aligned} \phi_{A-B}(t) &= \phi_A(t) - \phi_B(t) = \\ &= (2\pi f_{MIN} L e^{\frac{(t+\Delta T)}{L}}) - (2\pi f_{MIN} L e^{\frac{t}{L}}) = \\ &= 2\pi [f_{MIN} (e^{\frac{\Delta T}{L}} - 1)] L e^{\frac{t}{L}}. \end{aligned} \quad (16)$$

The difference frequency swept-sine signal has thus an instantaneous phase $\phi_{A-B}(t)$ which can be written in the form that guarantees the fulfillment of the phase alignment condition, and whose minimum frequency is $f_{MIN}(e^{\frac{\Delta T}{L}} - 1)$. Consequently, the conditions for phase alignment for each individual exponential chirp also guarantees phase alignment for the frequency-difference chirp signal. If the above conditions hold, when the signal $y(t)$ at the output of the non-linear system is processed through the matched filter, a signal $o(t)$ consisting of a sequence of short duration waveforms is obtained. The peaks deriving from the compression of the fundamental and

from the harmonics of each one of the two exponential swept-sine signals can be identified in $o(t)$. Moreover, the peak associated to the difference of the two exponential swept-sine signals is found at a time delay τ from the reference time; the corresponding waveform is denoted as $g_{A-B}(t)$. Figure 3 shows the trend of the $o(t)$ signal in the simple case of a quadratic non-linearity. The two plots at the top of Figure 3 show the $o(t)$ signal if the exponential chirp $x_{exp}^A(t)$ (black curve) and the $x_{exp}^B(t)$ (red curve) are placed separately at the input to the non-linear system: in this case the fundamental component is not present in the signal raised to the second power, the continuous component falls out of the chirp band. Consequently in the $o(t)$ signal there is only the peak associated to the second harmonic, at a time distance ΔT_2 (or $\Delta T_2 + \Delta T$) from the origin of time. In the case in which the two exponential chirp signals, temporally distant ΔT , are summed and put at the input of the quadratic non-linearity, the signal $o(t)$ at the output of the matched filter (represented in green) will contain, besides the two peaks relative to the components of second harmonic (already present for the single chirp), also a component associated to the instantaneous sum frequency and a component associated to the instantaneous difference frequency; this latter peak will be located with a delay τ with respect to the origin of time. Starting from the above waveforms obtained after compression with the matched filter, the *IMD* can be estimated as:

$$IMD(f)_{-chirp-} = \frac{|G_{A-B}(f)|}{|G_1^A(f) + G_1^B(f)|} \times 100 \quad (17)$$

where $G_{A-B}(f)$, $G_1^A(f)$ and $G_1^B(f)$ are the Fourier Transforms of the kernels $g_{A-B}(t)$, $g_1^A(t)$ and $g_1^B(t)$ respectively, where $g_1^A(t)$ and $g_1^B(t)$ are the impulsive functions associated to the fundamental component of the two swept-sine signals summed at the input. This formulation allows the estimation of the entire frequency trend of the intermodulation distortion to be obtained starting from the response to a single signal by which the non-linear system is energized. To properly select the $g_k(t)$, windows were adopted whose width is comparable to ΔT . We have adopted windows based on Reactance Transformation [28] with a very low value of the control parameter, in order to guarantee both a consistent reduction of the side lobes and a minimum alteration of the spectrum of the signals obtained.

Note that the *IMD* estimation based on the proposed method makes use of the Fourier transforms of three $g_k(t)$ among the functions obtained at the output of the matched filter. Therefore, the proposed method does not depend on the choice of the order N_H of the Hammerstein model to which the estimation is implicitly linked.

In fact, the proposed *IMD* estimation procedure is implicitly based on the hypothesis that the non-linear system can be represented by a parallel Hammerstein model. The validity of this hypothesis has been verified indirectly by comparing the estimates of the *IMD* obtained with the proposed method and the one provided by the standard procedure.

III. EXPERIMENTAL RESULTS

Several experiments were carried out to test the proposed approach to measure the *IMD*, *THD* and *HD_k*, analyzing

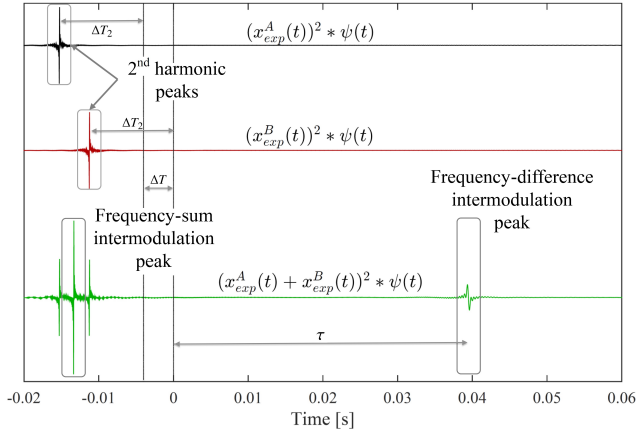


Fig. 3. Trend of the $o(t)$ signal in the simple case of a quadratic non-linearity: when the exponential chirp $x_{exp}^A(t)$ is input (black curve); when the exponential chirp $x_{exp}^B(t)$ is input (red curve); when the sum of the two exponential chirp signals $(x_{exp}^A(t) + x_{exp}^B(t))$ is input (green curve). The peaks associated to the sum and difference intermodulation components are evidenced in the green curve.

several scenarios. In particular, both a simulated system and real-world audio devices were tested to analyse the results obtained by means of the proposed method, compared to a well-known measurement system, *i.e.* the National Instruments sound and vibration toolkit [29]. The latter provides the possibility of measuring *IMD*, *THD* and *HD_k* according to the international standard [27]. A National Instruments PXI chassis (PXIe-1073) equipped with the PXI-4461 dynamic signal acquisition module [30], and connected to a desktop PC were used for the signals generation and acquisition. A LabVIEW script has been developed to manage the whole procedure. All the measurements were performed by setting a sampling frequency of 196 kHz, and repeated at various frequencies in order to obtain an estimation of the distortion parameters in the frequency range from 1.5 kHz to 15 kHz. The IEC standard-compliant measurement procedure for the estimate of harmonic distortion and intermodulation distortion was carried out separately for each frequency. For the harmonic distortion measurements, the frequency range of interest was divided into frequency intervals of 100 Hz width; for each of the 136 frequencies identified, a sinusoidal tone of 10 ms duration was sent, and the response signal of the non-linear system was analysed, detecting the amplitudes of the fundamental component and the harmonics up to the seventh one. From this data the values of *THD*, *HD₂* and *HD₃* were calculated for each of the 136 frequencies. About the measurement of the intermodulation distortion, for each of the 136 frequencies, a signal was sent consisting of the superimposition of two sinusoidal tones of frequencies adjacent to the measurement frequency, separated by 80 Hz from each other. In the case of the chirp measurement, the signal used for the experiments consisting of the superimposition of two exponential up-chirps, was defined taking into account the frequency band in which the estimate of distortion was to be made: the frequency

range of each of the two overlapping chirps was defined as a function of the band to be covered, extending it beyond the measurement frequencies to avoid the oscillations that occur at the ends of the band in the chirp signal spectrum. These aspects have been here deeply investigated and a way of selecting the parameters of the input signal has been defined that allows an accurate control of the quality of the estimate throughout the band of interest. In the work [24], in fact, limitations in the accuracy of the estimation had emerged at the extremes of the frequency band. This effect was analyzed in detail, verifying that it can be associated with two different aspects. A first aspect is related to the frequency band covered by the difference chirp, generated by the intermodulation of two chirps whose frequency ratio is $\frac{f_{inst}^B(t)}{f_{inst}^A(t)} = e^{\Delta T/L}$. If each of the incoming chirps covers a frequency range from f_{MIN} to f_{MAX} , the difference chirp C_{exp}^{DIFF} will extend from the minimum frequency $(e^{\Delta T/L} - 1)f_{MIN}$ to $(e^{\Delta T/L} - 1)f_{MAX}$. Consequently, it is necessary that the matched filter covers a frequency band that includes this extended range to get a correct estimate of the *IMD* in the entire frequency band. A second aspect is related to the value of the $T * B$ product associated to each one of the two chirps summed at the input: for the *IMD* estimation to be accurate, both frequencies on which the estimation is based must fall into a range where the chirp spectrum is appreciably flat. Meeting this condition requires (i) that the $T * B$ product is high enough; (ii) that the entity of the $f_{inst}^B(t) - f_{inst}^A(t)$ frequency difference is considered, in particular at the upper end of the covered band. The *IMD* is calculated between two harmonic components that differ, instant by instant, by a quantity that varies between $f_{MIN}(e^{\frac{\Delta T}{L}} - 1)$ and $f_{MAX}(e^{\frac{\Delta T}{L}} - 1)$ as t increases. The estimated *IMD* is associated at each instant with the intermediate frequency between the two. An excessive increase in the difference between the two frequencies on which the *IMD* is calculated can take one of the two components out of the useful band of the matched filter, and thus alter the *IMD* estimate for the corresponding frequency. Taking into account both these conditions, it was possible to define input signals that allow to predictably control the accuracy of the *IMD* estimation, as shown by the synthetic and real examples below. The time distance ΔT between the two chirps was defined in such a way that the difference between the instantaneous frequencies of the two overlapping chirps would be kept at contained values. The results shown in the following figures were obtained by sending a single input signal to the system, whose total duration was of 2.3 s; the two chirp components were superimposed at a time distance equal to 7.45 ms. The signal thus defined was sent only once to the system, to obtain the estimate of the frequency trend of the distortion. A number of experiments, not reported here, were performed, and similar results were obtained, with different input signals, for other frequency bands. The description of the measurement process provides a clear indication of the considerable difference in complexity and duration of the measurement operations carried out according to the two procedures, the one standardized by IEC and the one herein proposed.

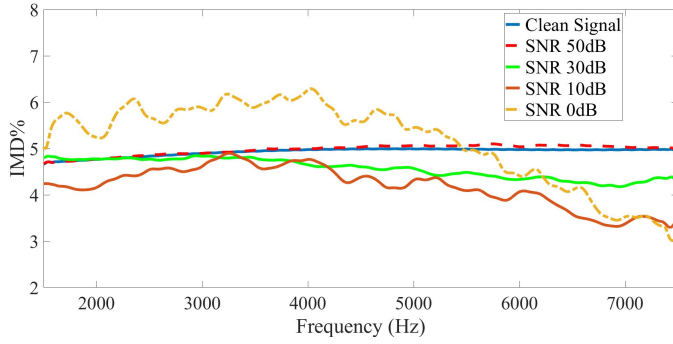


Fig. 4. IMD estimation for different Signal To Noise Ratio (SNR) values, in case the case of the non-linear polynomial function defined by (18).

A. Experiments with simulated systems

For the simulated experiments, the distortion parameters produced by a non-linear polynomial function and by the synthetic model of a vacuum tube have been measured. The first synthetic experiment was performed considering the following function:

$$f(x) = A \cdot x^2 + x^3, \quad (18)$$

where the parameter A allows to vary the amount of quadratic non-linearity. Since the non-linearity is defined in analytical form, it is possible to calculate the theoretical value of IMD , THD and HD_k according to the value assigned to the constant A : in this case its value was set to $A = 0.1125$, with an input signal of 2 V peak-to-peak amplitude (V_{pp}), leading to the theoretical values $IMD = 5\%$, $THD = 32.33\%$, $HD_2 = 7.10\%$, $HD_3 = 31.54\%$.

The robustness of the pulse compression techniques with respect to environmental noise is well known. We wanted to verify if this feature is also proper to the method of estimating the intermodulation distortion we propose, which is based on a pulse compression technique. This characteristic has been verified through the following experiment: we have added a zero-mean white Gaussian noise to the signal at the output of the non-linear system. The variance of the additive noise has been modified to adjust the resulting signal-to-noise ratio (SNR). It should be noted that the SNR ratio is evaluated on the signal input to the matched filter. At the output of this filter, the signal will be affected by a noise that will overlap the sequence of functions $g_k(t)$: since the amplitude of the different impulsive functions varies significantly between the different harmonics, the signal associated with the intermodulation component, typically much smaller than the amplitude of the fundamental and the first harmonics, will be greatly disadvantaged, buried in the noise even for not very low SNR values. Despite this, the $IMD(f)$ estimate, although worsening as the SNR value decreases, maintains trends close to the ideal one even for very low SNR values, as clearly shown in Figure 4.

A further simulated experiment was carried out considering a non-linear behavior that represents the effect of a vacuum tube [31]. Figure 5 shows the block diagram of the non-linearity.

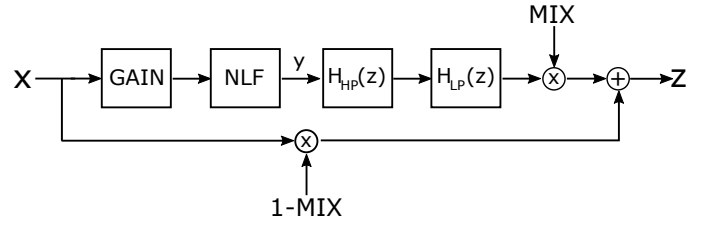


Fig. 5. Vacuum tube synthetic model block diagram.

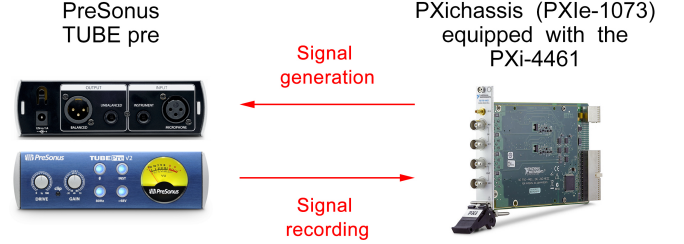


Fig. 6. Hardware connection for the real experiment considering the PreSonus TubePre preamplifier.

The first step is related to the application of the following non-linear function (NLF):

$$f(x) = \frac{x - \beta}{1 - e^{-\alpha(x-\beta)}} + \frac{\beta}{1 - e^{\alpha\beta}}, \quad \alpha \neq 0, x \neq \beta, \quad (19)$$

where x is the input and α, β are the parameters used to modify the distortion values. Then, both a low-pass and a high-pass filtering of the output signal y are performed to remove the DC component and to shape the higher frequencies.

The following filters are used:

$$H_{HP}(z) = \frac{1 - 2z^{-1} + z^{-2}}{1 - 2r_h z^{-1} + r_h^2 z^{-2}}, \quad H_{LP}(z) = \frac{1 - r_l}{1 - r_l z^{-1}}, \quad (20)$$

where H_{HP} and H_{LP} are the high-pass and low-pass filters, respectively, and r_l, r_h are their control parameters. A final control is used to mix the distorted and the original signal managing the amount of resulting distortion through the use of the MIX variable (see Figure 5).

For the experiment, an input signal of 2 V_{pp} has been used and the following values have been considered:

- $\beta = -0.2$;
- $r_l = 0.1, r_h = 0.99$;
- $\alpha = 10$;
- $GAIN = 1$;
- $MIX = 1$,

where $GAIN$ and MIX refer to the variables of Figure 5 and the selected MIX value determines the presence of the distorted signal only.

B. Experiments with real devices

For the real experiments, the distortion parameters produced by two real devices based on vacuum-tube technology, were measured.

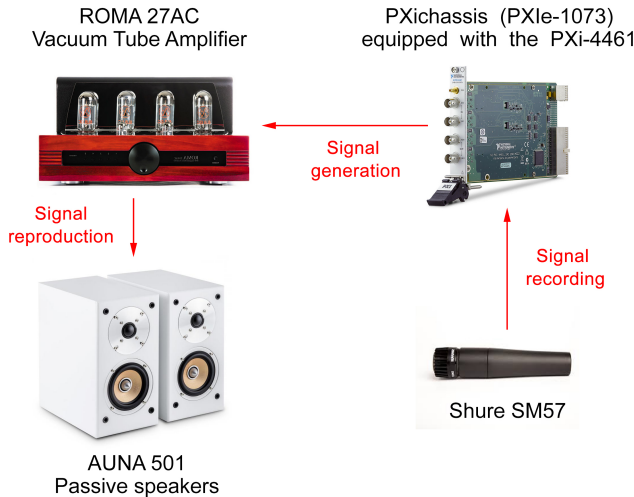


Fig. 7. Hardware connection for the real experiment considering the Synthesis Roma 27AC amplifier.

The former device was a preamplifier, *i.e.*, the PreSonus TubePre, a microphone and instruments preamplifier equipped with a 12AX7 triode valve. The device has been connected directly to the PXI-4461 acquisition board, as reported in Figure 6. For this experiment, the two main controls of the preamplifier have been used, *i.e.* the drive that acts on the valve saturation and the gain control which acts on the overall gain. They have been set to 3 dB and to 0 dB respectively, to obtain a low level of non-linearity, as reported in the experiments. Measurements have been made using a $2 V_{pp}$ input signal.

The latter device was a Hi-Fi amplifier, *i.e.* the Synthesis Roma 27AC, a 25 Watt AB Class tube amplifier. The hardware connection is reported in Figure 7. The amplifier was connected to the PXI-4461 acquisition board for the signal generation and to a pair of loudspeakers that are used to load the amplifier and for the signal reproduction. In particular, an Auna 501 passive 2-way speaker has been used featuring an RMS rated power of 100 W, an impedance of 8 Ω , and a frequency range of 92 Hz - 20 kHz. The volume of the amplifier has been set to the highest value and an input signal of $1 V_{pp}$ has been used. After half an hour of warm up, the input signal was applied and the output signal has been measured by means of a dynamic microphone, *i.e.* a Shure SM57, placed in front of one of the speakers.

C. Discussion

Figure 8 shows the results obtained with the polynomial function introduced in Section III-A. The proposed approach exhibits excellent results in comparison to both the theoretical values and to the method of IEC 60268, for all the measured distortion parameters, showing only some minor difference in the estimation, ranging from 0.1% for HD_2 to 0.02% for HD_3 , on average. Figure 9 shows the results obtained with the synthetic tube model of Section III-A: a good accordance

TABLE I. MEAN PERCENT DIFFERENCE BETWEEN THE PROPOSED APPROACH AND IEC 60268 MEASUREMENTS, FOR EACH DISTORTION PARAMETER CONSIDERED.

Experiment	Δ_{IMD}	Δ_{THD}	Δ_{HD_2}	Δ_{HD_3}
Polynomial model	0.1%	0.03%	0.1%	0.02%
Vacuum Tube model	0.9%	0.09%	0.08%	0.02%
Preamplifier	0.1%	0.05%	0.6%	0.01%
Hi-Fi Amplifier	0.05%	0.4%	0.3%	0.3%

with the IEC measured parameters is obtained, showing only some minor difference in the estimation, ranging from 0.02% for HD_3 to 0.9% for IMD , on average.

Figure 10 shows the obtained results for the first experiment on real devices, *i.e.*, the PreSonus TubePre, considering a low level of non-linearity. Taking into consideration the HD_k and THD measurement, the proposed approach shows good results in comparison to the IEC 60268 methodology. This fact is also confirmed by the overall distortion measurement, that presents a very small difference at lower frequencies, ranging from 0.1% for IMD to 0.01% for HD_3 , on average. Figure 11 shows the results obtained for the Hi-Fi amplifier plus the loudspeakers that present a high level of distortion. In this case, from the point of view of harmonic distortion measurement, there is a difference in the second harmonic, that seems to be under determined. However, the THD measurement is less affected and the result accurately follows the one obtained according to the IEC 60268 methodology, also in terms of IMD values. In particular, the difference on the measurement varies from 2% for HD_2 , to 0.1% for IMD , on average. Moreover the underestimation is limited to the frequency range from 1 to 8 kHz, where the system seems to have a strong non-linear behavior. Future work will be oriented to a deeper investigation of this aspect performing further experiments with different loudspeakers and amplifiers.

Table I shows the experimental values of the differences in estimation between the proposed method and the one according to the IEC 60268 specification: the proposed approach allows an accurate estimation of IMD , HD_k and THD , showing only a limited error in some measurements. In a real scenario the proposed approach can be a valid alternative to support the IEC procedure for preliminary, fast and reliable distortion parameters estimation. In the case of very complex frequency trends and high distortion levels, if an estimate with a high degree of accuracy is required, the standard IEC procedure in an acoustically controlled environment should be used.

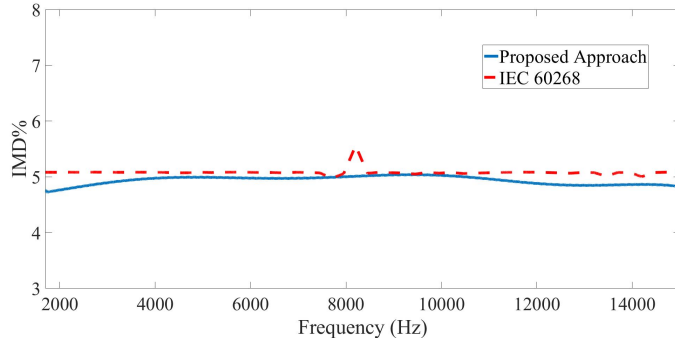
IV. CONCLUSION

In this paper, an innovative approach based on pulse compression capable of estimating the difference frequency intermodulation distortion has been proposed. The procedure was theoretically derived and experimentally verified on both synthetic and real-world non-linear systems. The procedure proved to be reliable and to allow an accurate estimate, even when extremely low distortion values are involved. The proposed technique allows to obtain the intermodulation distortion with

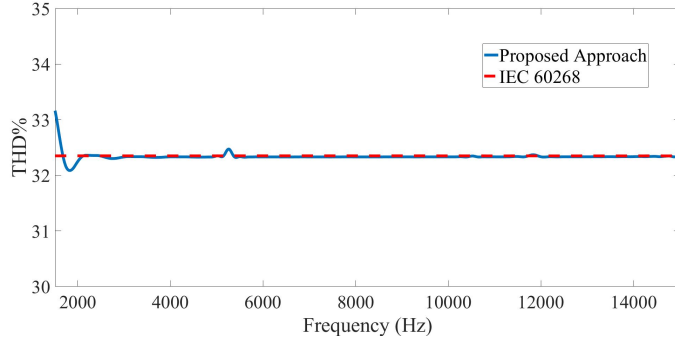
a single pulse compression operation and, at the same time, the estimate of the total harmonic distortion and of the harmonic distortions of different orders. The practical implementation of the procedure requires to process the response of the nonlinear system to a single signal with a duration of the order of a few seconds. The short duration of the signal, the light computational cost, comparable to that of an *FFT*, and the low sensitivity of the pulse compression techniques to the environmental noise, make the procedure extremely simple to apply even in uncontrolled operating situations. Future works will be oriented to the introduction of suitable time windows to be applied to the input signal and/or to the matched filter, in order to further improve the *IMD* estimate.

REFERENCES

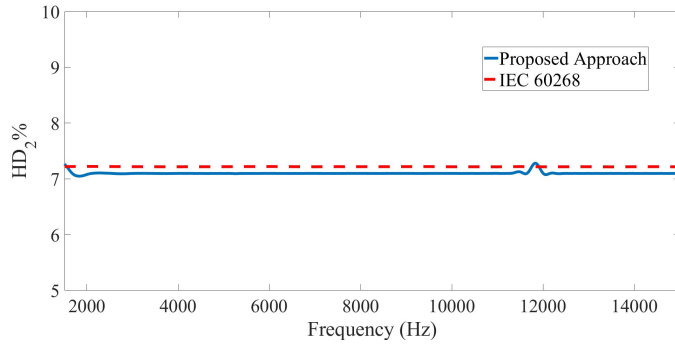
- [1] D. Bard, "Horn loudspeakers nonlinearity comparison and linearization using Volterra series," in *Audio Engineering Society Convention 124*, May 2008. [Online]. Available: <http://www.aes.org/e-lib/browse.cfm?elib=14448>
- [2] G. D. Capua and N. Femia, "A novel method to predict the real operation of ferrite inductors with moderate saturation in switching power supply applications," *IEEE Transactions on Power Electronics*, vol. 31, no. 3, pp. 2456–2464, March 2016.
- [3] P. Burrascano, G. D. Capua, N. Femia, S. Laureti, and M. Ricci, "Pulse compression for ferrite inductors modeling in moderate saturation," in *2018 15th International Conference on Synthesis, Modeling, Analysis and Simulation Methods and Applications to Circuit Design (SMACD)*, July 2018, pp. 229–232.
- [4] M. Scalerandi, A. Gliozzi, and D. Olivero, "Discrimination between cracks and recrystallization in steel using nonlinear techniques," *Journal of Nondestructive Evaluation*, vol. 33, pp. 269–278, 06 2014.
- [5] P. Burrascano, S. Laureti, and M. Ricci, "Harmonic distortion estimate for damage detection," in *NAECON 2018 - IEEE National Aerospace and Electronics Conference*, July 2018, pp. 274–279.
- [6] M. Gasparini, L. Romoli, S. Cecchi, and F. Piazza, "Identification of Hammerstein model using cubic splines and FIR filtering," in *Proc. of 8th Int. Symposium on Image and Signal Processing and Analysis*, Trieste, Italy, Sep. 2013.
- [7] L. Romoli, M. Gasparini, S. Cecchi, A. Primavera, and F. Piazza, "Adaptive identification of nonlinear models using orthogonal nonlinear functions," in *Proc. 48th Audio Engineering Society Conference*, Monaco, Germany, Sep. 2012.
- [8] M. Gasparini, A. Primavera, L. Romoli, S. Cecchi, and F. Piazza, "System identification based on Hammerstein models using cubic splines," in *Proc. of 134th Audio Engineering Society Convention*, Rome, Italy, May 2013.
- [9] V. Volterra, *Theory of functionals and of integral and integro-differential equations*. Blackie & Son Limited, 1930.
- [10] M. Schetzen, *The Volterra and Wiener theories of nonlinear systems*. Wiley, 1980.
- [11] J. Schoukens, R. Pintelon, and T. Dobrowiecki, "Linear modeling in the presence of nonlinear distortions," *IEEE Transactions on instrumentation and measurement*, vol. 51, no. 4, pp. 786–792, 2002.
- [12] J. Schoukens, R. Pintelon, T. Dobrowiecki, and Y. Rolain, "Identification of linear systems with nonlinear distortions," *Automatica*, vol. 41, no. 3, pp. 491–504, 2005.
- [13] S. Boyd and L. O. Chua, "Fading memory and the problem of approximating nonlinear operators with Volterra series," *IEEE Transactions in Circuits and Systems*, vol. CAS-32, no. 11, pp. 1150–1161, Nov. 1985.
- [14] K. Narendra and P. Gallman, "An iterative method for the identification of nonlinear systems using a Hammerstein model," *IEEE Transactions on Automatic Control*, vol. 11, no. 3, pp. 546–550, Jul 1966.
- [15] A. Novak, L. Simon, F. Kadlec, and P. Lotton, "Nonlinear system identification using exponential swept-sine signal," *IEEE Trans. Instrum. Meas.*, vol. 59, no. 8, pp. 2220–2229, Aug. 2010.
- [16] A. Novak, B. Maillou, P. Lotton, and L. Simon, "Nonparametric identification of nonlinear systems in series," *IEEE Trans. Instrum. Meas.*, vol. 63, no. 8, pp. 2220–2229, Aug. 2014.
- [17] A. Farina, A. Bellini, and E. Armelloni, "Non-linear convolution: a new approach for the auralization of distorting systems," in *Proc. 110th Audio Engineering Society Convention*, Amsterdam, NL, May 2001.
- [18] M. Rébillat, R. Hennequin, E. Corteel, and B. Katz, "Identification of cascade of Hammerstein models for the description of nonlinearities in vibrating devices," *Journal of Sound and Vibration*, vol. 330, pp. 1018–1038, Feb. 2011.
- [19] A. E. Nordsjö and L. H. Zetterberg, "Identification of certain time-varying nonlinear Wiener and Hammerstein systems," *IEEE Transactions on Signal Processing*, vol. 49, no. 3, pp. 577–592, Mar 2001.
- [20] V. J. Mathews and G. L. Sicuranza, *Polynomial Signal Processing*. New York: Wiley, 2000.
- [21] M. Scarpiniti, D. Comminiello, R. Parisi, and A. Uncini, "Comparison of Hammerstein and Wiener systems for nonlinear acoustic echo cancelers in reverberant environments," in *17th International Conference on Digital Signal Processing (DSP)*, July 2011, pp. 1–6.
- [22] A. Primavera, M. Gasparini, S. Cecchi, W. Hariya, S. Murai, K. Oishi, and F. Piazza, "A novel measurement procedure for Wiener/Hammerstein classification of nonlinear audio systems," in *Audio Engineering Society Convention 144*, May 2018. [Online]. Available: <http://www.aes.org/e-lib/browse.cfm?elib=19476>
- [23] J. Vanderkooy and S. P. I. Thomson, "Harmonic distortion measurement for nonlinear system identification," in *Proc. 140th Audio Engineering Society Convention*, Paris, France, Jun. 2016.
- [24] P. Burrascano, S. Laureti, M. Ricci, A. Terenzi, and S. Cecchi, "Pulse Compression Procedure for an Effective Measurement of Intermodulation Distortion," in *Proc. 10th Int'l Symposium on Image and Signal Processing and Analysis*, Ljubljana, Slovenia, Sep. 2017.
- [25] A. Novak, P. Lotton, and L. Simon, "Synchronized swept-sine: Theory, application, and implementation," *Journal of the Audio Engineering Society*, vol. 63, no. 10, pp. 786–798, 2015.
- [26] A. Farina, "Simultaneous measurement of impulse response and distortion with a swept-sine technique," in *Audio Engineering Society Convention 108*. Audio Engineering Society, 2000.
- [27] "Sound system equipment part 5: Loudspeakers," International Standard, IEC 60268-5, 2003.
- [28] P. Burrascano, S. Laureti, L. Senni, and M. Ricci, "Pulse compression in nondestructive testing applications: Reduction of near sidelobes exploiting reactance transformation," *IEEE Transactions on Circuits and Systems I: Regular Papers*, vol. Available on line at 10.1109/TCSI.2018.2862868, p. 11, Aug. 2018.
- [29] "Sound and vibration toolkit," National Instruments, 2017. [Online]. Available: <https://www.ni.com/trysv/>
- [30] "PXIe-4461 DSA dynamic signal acquisition device," National Instruments, 2008. [Online]. Available: <http://www.ni.com/pdf/manuals/373770j.pdf>
- [31] U. Zölzer, *DAFX: digital audio effects*. John Wiley & Sons, 2011.



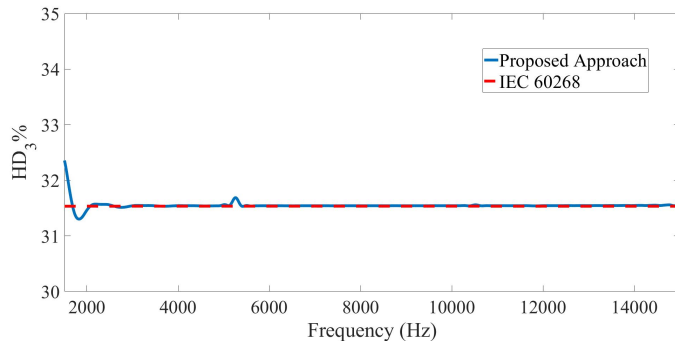
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(b)

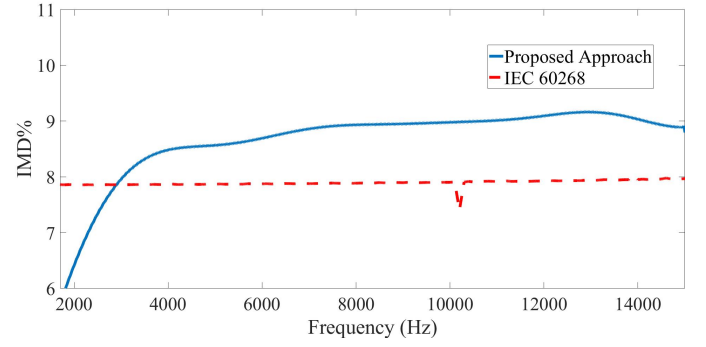


(c)

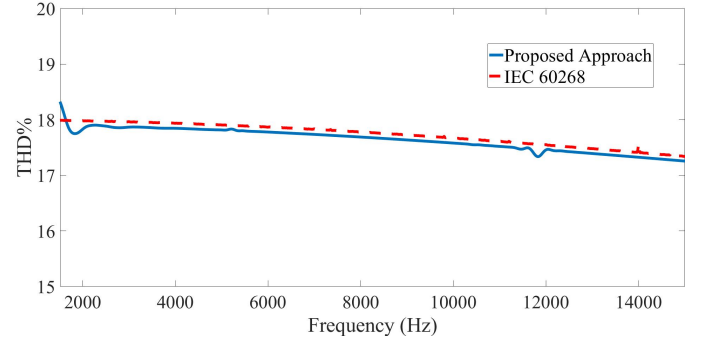


(d)

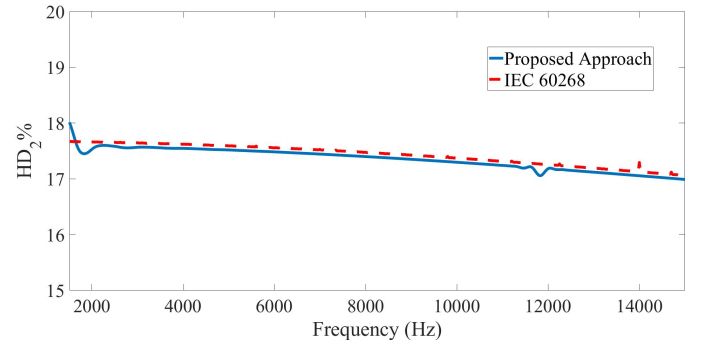
Fig. 8. Results in terms of percent in the case of non-linear function defined by 18 (a) IMD , (b) THD , (c) HD_2 , and (d) HD_3 considering the proposed approach and the IEC 60268 measurements.



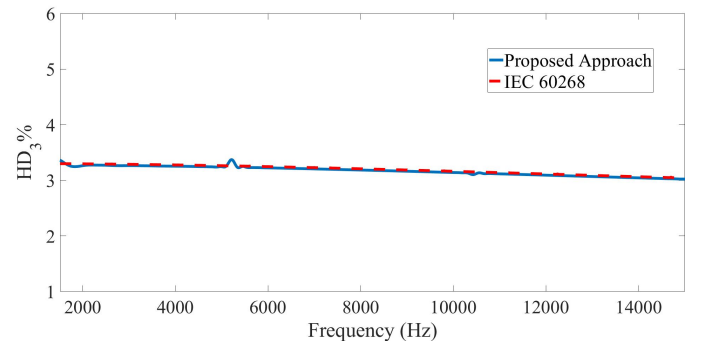
(a)



(b)



(c)



(d)

Fig. 9. Results in terms of percent for the synthetic experiment described in Figure 5 in terms of percent (a) IMD , (b) THD , (c) HD_2 , and (d) HD_3 considering the proposed approach and the IEC 60268 measurements.

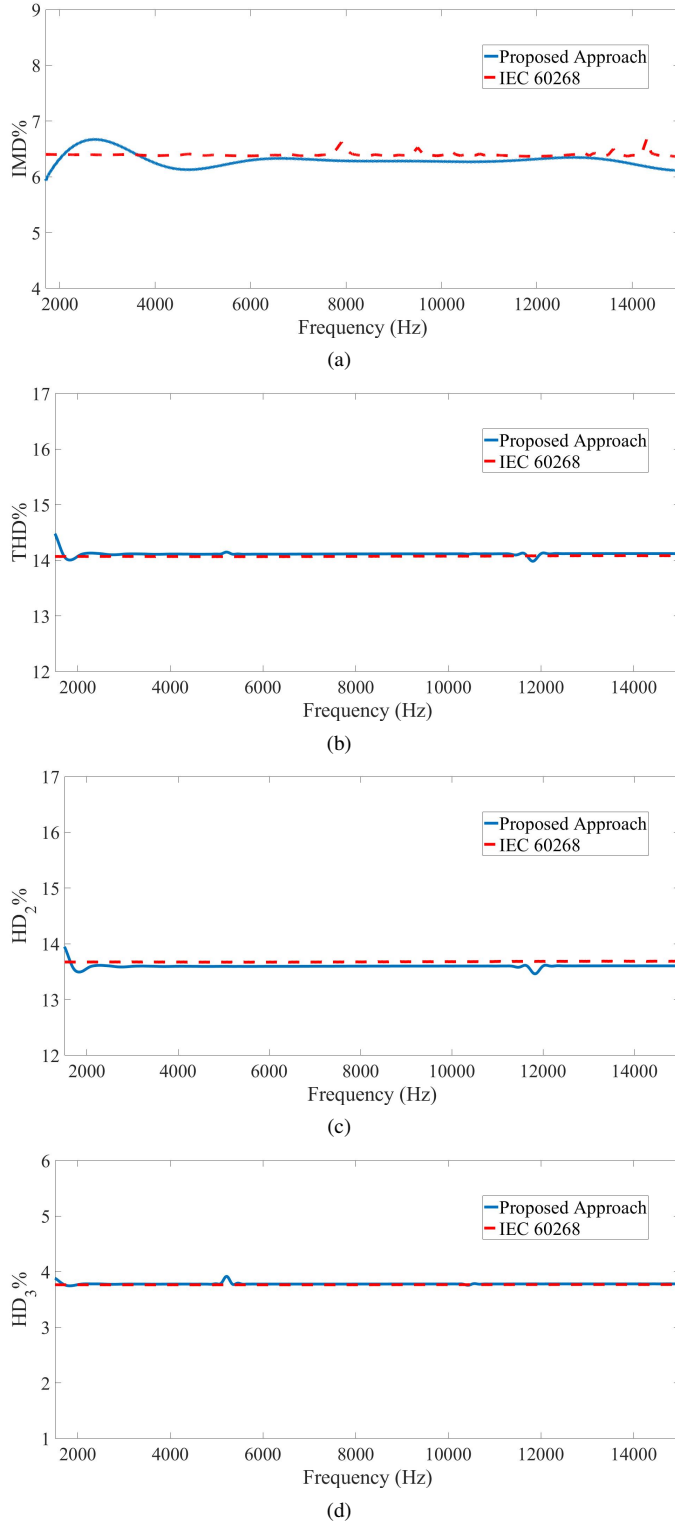


Fig. 10. Results in terms of percent for the PreSonus tube preamplifier experiment (a) IMD , (a) THD , (c) HD_2 , and (d) HD_3 considering the proposed approach and the IEC 60268 measurements.

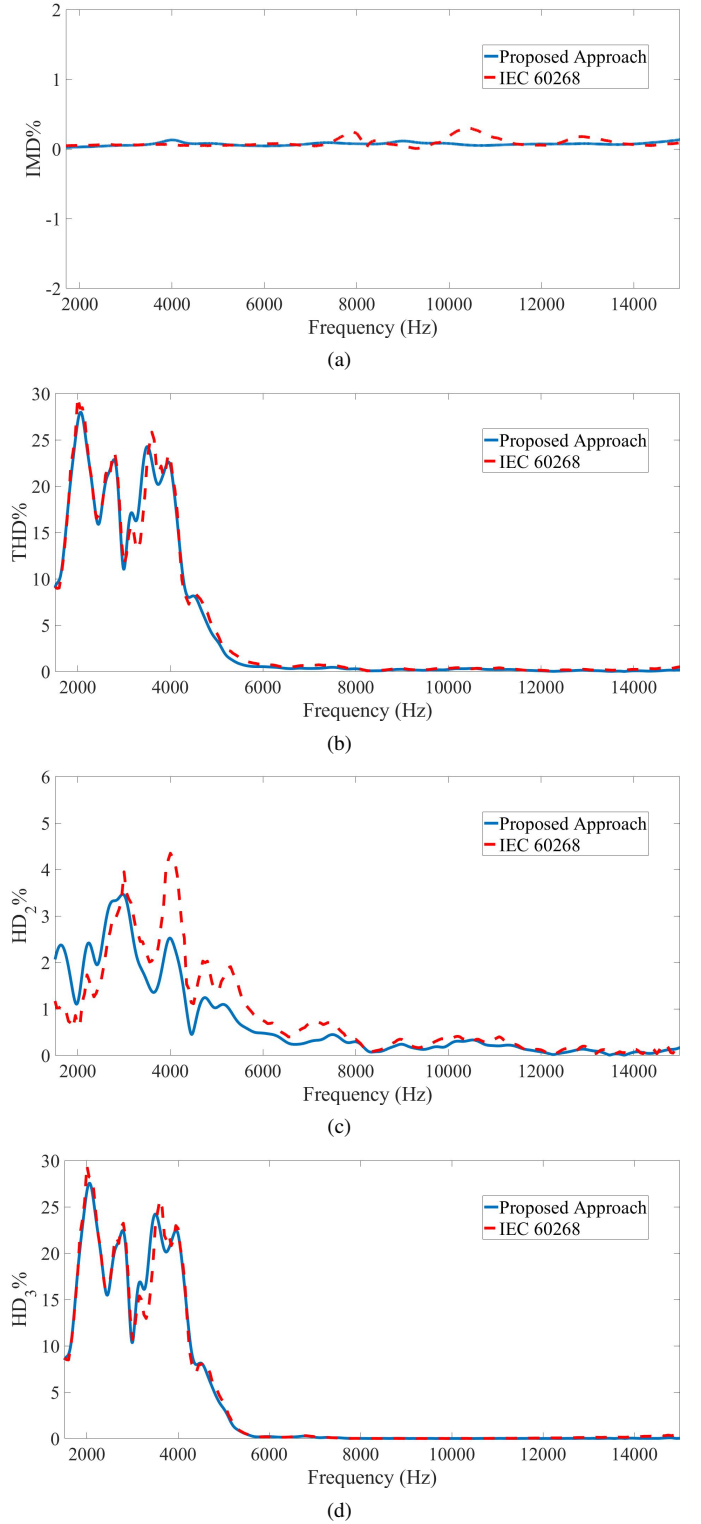


Fig. 11. Results in terms of percent for the Roma Hi-Fi amplifier with Auna louspeakers experiment, (a) IMD , (b) THD , (c) HD_2 , and (d) HD_3 considering the proposed approach and the IEC 60268 measurements.