A HIGH PERFORMANCE SWITCHING-MODE AC POWER SOURCE FOR EXCITATION OF AN ELECTRODYNAMIC SHAKER

L. Della Flora, H. A. Gründling
Grupo de Eletrônica de Potência e Controle, UFSM. Santa Maria - RS - Brazil
Emails: ldella@mail.ufsm.br, ghilton@ctlab.ufsm.br

***Abstract*— This paper presents the development of a switching-mode AC power amplifier for armature excitation of an electrodynamic shaker. The proposed solution is based on a voltage controlled single-phase PWM inverter with digital control. A model reference adaptive controller that assumes reduced knowledge of the plant and guarantees robustness to parameters variations and rejection of bounded disturbances is employed. Fixed and swept frequency experimental results show that the proposed solution is capable to achieve good voltage reference tracking over a wide frequency range (20 Hz to 2 kHz). The system configuration, the dynamic modeling and the voltage controller of the AC power amplifier are described.**

***Keywords* - Shakers, Pulse-Width Modulated Inverters, Robust Adaptive Control, Discrete-Time Control.**

I. INTRODUCTION

The electrodynamic shaker is an essential piece of test equipment for vibration-proof testing of electronic and mechanical products. The major task of vibration control is to let the table carrying the device under test be accurately vibrated according to the command. Application areas include aerospace, automotive, shipping industries, building, electronics, machine-tool and package. The most common uses are related to production control, frequency response, dynamic performance and environment tests [1].

The equipment usually employed for the execution of mechanical vibration tests includes not only the shaker, but also a power amplifier and an acceleration controlling and monitoring system. Traditionally, electrodynamic shakers are fed by linear audio power supplies with controlled voltage and frequency. This scheme usually results in high vibration performance due to its low distortion power, but also presents large size and weight and low conversion efficiency. The use of a switching-mode inverter replacing the linear one can avoid these disadvantages, but more sophisticated control is needed to reduce the effects of harmonics on vibration performance.

The development of high performance switching-mode power amplifiers for electrodynamic shakers has constituted an important research interest area for shaker manufacturers over the years [2]. In general, designing a power amplifier for armature excitation of vibration machines requires low distortion and good voltage or current control performance over a wide dynamic and frequency range (typically, 10 Hz to 2 kHz). In this case, the effects of the back electromotive force generated across the shaker armature as well as the parameter variations have a substantial influence on the response of the electrodynamic vibration machine.

In [3]–[5], the design and implementation of a switching-mode power amplifier for shaker armature excitation has been presented. The proposed solution is based on a current controlled PWM voltage source inverter with high switching frequency and low output power (50 kHz, 400 W). Good current control characteristics were demonstrated experimentally. However, current mode operation usually gives the amplifier high output impedance, which is more advantageous when acceleration closed loop control is not required and the shaker is operated in its mid frequency range [2].

A voltage controlled PWM inverter rated at 5 kVA and 80 kHz has been presented in [6]. The control scheme is based on an analog implemented PI controller. Experimental results were provided between 2 Hz and 5 kHz, but with no direct comparison between the output voltage and its command.

Another solution consisting of a 3 kVA, 12 kHz voltage controlled PWM inverter has been proposed in [7]. The control strategy is based on a digital implemented robust model reference adaptive algorithm. Good results were demonstrated experimentally between 10 Hz and 1 kHz. However, since most vibration test specifications require frequencies up to 2 kHz, some modifications in the inverter design and voltage controller need to be considered.

Then, the main contribution of this paper is to present an improvement of the AC power source frequency range of [7] by redesigning the inverter and introducing modifications in the robust adaptive voltage control law. The system configuration, the dynamic modeling and the voltage controller of the AC power amplifier are described in detail.

II. SYSTEM CONFIGURATION AND DYNAMIC MODELING OF THE PROPOSED AC POWER SOURCE

A. System Configuration

The schematic diagram of the switching-mode AC power source is shown in Fig. 1(a). It consists of a three-phase full-bridge uncontrolled rectifier, a filter capacitor, a single-phase voltage-controlled PWM inverter and a LC filter. The inverter generates a three-level pulse-width modulated voltage v_{pwm} based on the voltage control signal u. The output filter is used to reduce the harmonic content of the shaker input voltage v_o in order to eliminate vibrations outside the main frequency range. A digital control platform acquire the output variables, compute the control law and generate the PWM signals.

The electrodynamic shaker (model St 5000/300 manufactured by TIRA) is a 1 kVA, 110 V vibration machine capable to operate from 20 Hz to 5 kHz. The maximum sine force peak

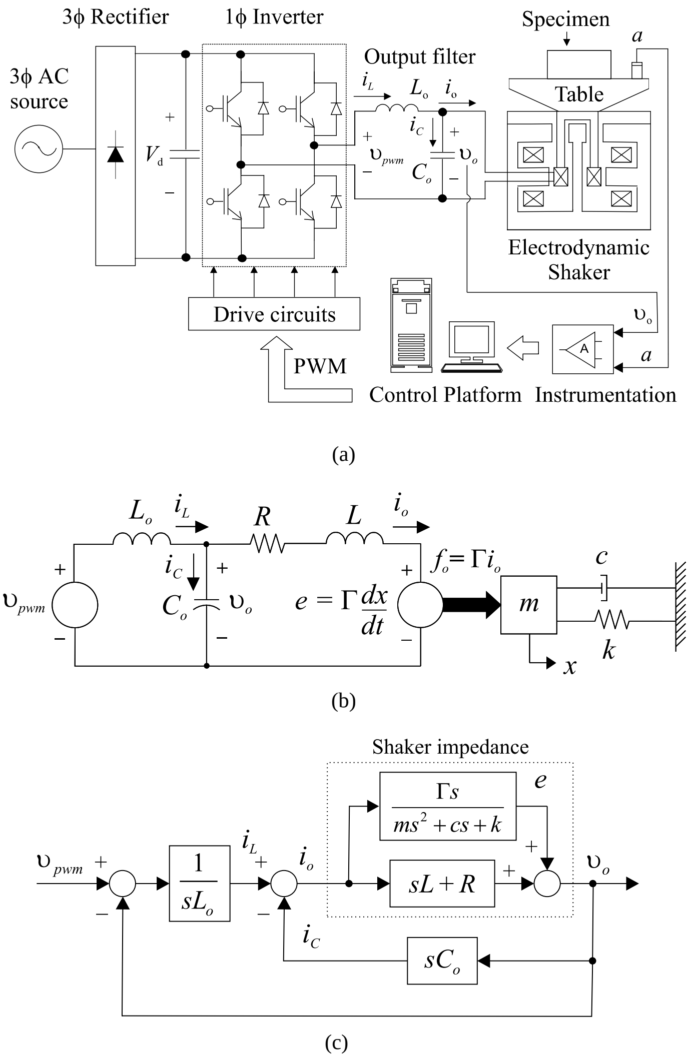

Fig. 1. The Switching-mode AC power source: (a) System configuration. (b) Equivalent electromechanical circuit. (c) Transfer function block diagram.

is 3 kN, the mass of the armature, table and fixture assembly is equal to 22 kg and the maximum specimen weight is 46 kg.

The proposed AC power supply is designed to operate as a voltage source for the electrodynamic shaker so that sinusoidal accelerations can be produced. Based on the vibration machine's constraints and on typical vibration tests requirements, the frequency range is set equal to 20 Hz to 2 kHz and the maximum voltage and current amplitudes are defined as 110 V and 20 A, respectively. A 2 kVA three-phase insulated transformer (Fig.2) supplies the input rectifier and a soft starting circuit composed by resistors, contactors, contact-bottom and timer relay is employed to protect the rectifier from the inrush current. The single phase full-bridge PWM inverter is composed by two IGBT modules manufactured by Semikron (model SK45GB063) and rated at 600 V, 30 A and 30 kHz. Due to limitations of the PWM signal generation, the switching frequency is set equal to 24 kHz. The IGBT modules are driven by two SKHI20opA Semikron drivers featuring short circuit protection and interlock circuit to prevent the two IGBTs to switch at the same time (the dead time is configured as 0.25 μs). To reduce the harmonic content of the output voltage υ_o, a LC filter is employed with a cutoff frequency of 3.2 kHz designed by observing the need of establishing a compromise between the desired maximum harmonic distortion, the high-frequency amplification and the phase change produced by the LC filter insertion. A balance between the inductor current and the capacitor voltage ripple is obtained by means of a 250 μH, 10 μF design.

A multifunction interface board is used to communicate the microprocessor and drive module. It includes two 12 bits A/D converters at 24 kHz sampling frequency, three 16 bits counters and three PWM signal generators. The software and hardware synchronization is carried out by the use of the three 16 bits counters. A program developed in Borland C programming language in the microprocessor acquires the external variables and computes the control law according to the specified voltage amplitude and frequency. The control signal is converted into a PWM signal and then applied to the ACPS module through the multifunction interface board.

The AC power source and microcomputer's interface is provided by an internal board that converts the TTL level PWM control signals from the microcomputer to the input drive circuits of the inverter module. The output voltage is measured by means of a signal conditioner that converts the - 110 V to 110 V output voltage range to TTL levels needed for digital conversion in the microcomputer.

Fig. 2. The designed AC power source

B. Dynamic Model

Assuming that the mechanical dynamic behavior of the electrodynamic vibration machine is approximately represented by the single degree-of-freedom mass-spring system of Fig. 1(b) and that the air-gap flux density built up by the field winding is constant, the electrical and mechanical equations that governs the shaker system can be written as

$$\upsilon_{pwm} = L_o \frac{di_L}{dt} + \upsilon_o \tag{1}$$

$$\upsilon_o = L\frac{di_o}{dt} + Ri_o + e \tag{2}$$

$$C_o\frac{d\upsilon_o}{dt} = i_c = i_L - i_o \tag{3}$$

$$f_o = \Gamma i_o = m\frac{d^2x}{dt^2} + c\frac{dx}{dt} + kx \tag{4}$$

where

$\Gamma \triangleq Bl$ is a force-generating constant;
B is the magnetic flux density;
l is the effective length of the armature conductors;
e is the back electromotive force (EMF);
f_o is the linear force developed by the armature winding;
υ_{pwm} is the PWM inverter output voltage;
i_L is the inductor current of the output filter;
L_o is the inductance of the output filter;
C_o is the capacitance of the output filter;
υ_o is the voltage across C_o;
L is the armature winding inductance;
R is the armature winding resistance;
x is the displacement of the armature;
$\upsilon = dx/dt$ is the vibration velocity of the armature;
$a = d^2x/dt^2$ is the vibration acceleration of the armature;
i_o is the armature winding current;
m is the mass of the whole armature and table assembly;
c is the damping coefficient;
k is the spring constant.

One of the most important components in the equivalent circuit is the motional back EMF, which gives the shaker the active nature. Since $e = \Gamma dx/dt$, the shape of the terminal voltage characteristic will initially tend to follow that of the velocity characteristic. However, for a constant amplitude sinusoidal acceleration *a*, the motional EMF dominance in the equivalent circuit fades as the frequency increases, owing to the continual decrease in velocity ($\upsilon = a/\omega$, where ω is the vibration angular frequency). Then, the armature's winding reactance and resistance drops become significant and finally dominant. At very low frequencies (0 to l0 Hz), suspension forces exceed *ma* inertial forces, the inductance drops are negligible and the equivalent circuit approximates to *R* in series with the EMF generator *e*. At the suspension mass-spring resonance (typically 20 Hz), motion can be sustained with virtually no current. Beyond this point, the EMF generator has a value inversely proportional to the vibration frequency. Above 1 kHz or so, the equivalent circuit can then be approximated to one comprising *R* and *L* in series. These parameters are not constant due to skin effects [2]. In addition, the shaker's υ_o and i_o characteristic changes from an capacitive load at low frequencies to an inductive nature over the mid frequency range.

III. AC Power Source Voltage Controller

The main requirement of the AC power amplifier voltage controller is to regulate accurately the amplitude and frequency of the shaker supply voltage υ_o according to the specified command. To achieve this goal, the harmonic vibration effects due to nonsinusoidal driven power should be reduced as far as possible. So, in addition to have good power electronics, sophisticated control technique is also indispensable for obtaining the desired control characteristics.

The choice of a nominal load representing the actual one makes the design of the AC power source voltage controller simpler. In this case, the dynamic of a voltage-fed shaker can be even more disregarded by means of a robust adaptive control scheme intended to compensate systems subjected to parameter variations, unmodeled dynamics and bounded disturbances.

Assuming, therefore, that the back electromotive *e* indicated in Fig. 1(c) corresponds to an unknown bounded disturbance and that the shaker's armature inductance is not modeled, the AC power source nominal load can be represented by the 1 kVA, 110 V resistive load R shown in Fig. 3. Then, the ACPS nominal plant composed by the L_oC_o output filter-resistive load association can be stated as

$$G_v(s) = \frac{\upsilon_o(s)}{\upsilon_{pwm}(s)} = G_{vo}(s)\left[1 + \mu\Delta_m(s)\right] + \mu\Delta_a(s) \tag{5}$$

where

$$G_{vo}(s) = k_p\frac{Z_o(s)}{R_o(s)} = \frac{4\times 10^8}{s^2 + 8.33\times 10^3 s + 4\times 10^8} \tag{6}$$

is the transfer function of the modeled part of the plant, $\mu\Delta_m(s)$ and $\mu\Delta_a(s)$ are multiplicative and additive plant perturbations and $Z_o(s)$ and $R_o(s)$ are monic polynomials of degree m_p and n_p, respectively.

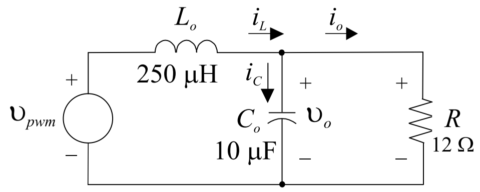

Fig. 3. The nominal load of the AC power amplifier.

By doing so, the following assumptions are satisfied:
A1 - $Z_o(s)$ is a monic Hurwitz polynomial of degree m_p ($\leq n_p{-}1$);
A2 - $R_o(s)$ is a monic polynomial of degree n_p;
A3 - the sign of k_p and the values of m_p and n_p are known.
A4 - $\Delta_a(s)$ is a strictly proper stable transfer function;
A5 - $\Delta_m(s)$ is a stable transfer function;
A6 - a lower bound $p_o > 0$ for which the poles of $\Delta_a(s-p)$ and $\Delta_m(s-p)$ are stable is known.

The adaptive control objective can then be stated as follows: given the reference model

$$\frac{\upsilon_m(s)}{r(s)} = W_m(s) = k_m\frac{1}{D_m(s)} \tag{7}$$

where $D_m(s)$ is a monic Hurwitz polynomial of degree $n^* = n_p - m_p$ and r is an uniformly bounded reference, design an adaptive controller so that for some $\mu^* > 0$ and any $\mu \in [0, \mu^*)$ the resulting closed-loop plant is stable and the plant output υ_o tracks the reference model output υ_m as closely as possible for all plant perturbations $\Delta_a(s)$ and $\Delta_m(s)$ satisfying **A4-A6**.

To achieve the adaptive control objective, the input *u* and the output υ_o are used to generate $(n_p - 1)$ dimensional auxiliary vectors as

$$\begin{aligned} \boldsymbol{\omega}_1 &= (s\mathbf{I} - \mathbf{F})^{-1}\mathbf{q}u \\ \boldsymbol{\omega}_2 &= (s\mathbf{I} - \mathbf{F})^{-1}\mathbf{q}\upsilon_o \end{aligned} \tag{8}$$

where $\mathbf{F}$ is a $(n_p - 1) \times (n_p - 1)$ stable matrix and **(F,q)** is a controllable pair. Then, the input control law can be computed from

$$u = \boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\omega} + c_o r \tag{9}$$

where $\boldsymbol{\theta}^{\mathrm{T}} = [\ \boldsymbol{\theta}_1^{\mathrm{T}} \quad \boldsymbol{\theta}_2^{\mathrm{T}} \quad \theta_3\ ]$ is the $(2n_p - 1)$ control parameter vector, $\boldsymbol{\omega}^{\mathrm{T}} = [\ \boldsymbol{\omega}_1^{\mathrm{T}} \quad \boldsymbol{\omega}_2^{\mathrm{T}} \quad \upsilon_o\ ]$ and c_o is a scalar feedforward parameter.

The augmented error ε_1 can be expressed as

$$\varepsilon_1 = \upsilon_o - \upsilon_m + \boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\zeta} - W_m(s)\boldsymbol{\theta}^{\mathrm{T}}\boldsymbol{\omega} \tag{10}$$

where $\boldsymbol{\zeta} = W_m(s)\mathbf{I}\boldsymbol{\omega}$.

The parameter adaptation algorithm corresponds to a modified least-squares algorithm presented in [8], i.e.,

$$\dot{\boldsymbol{\theta}} = -\sigma\mathbf{P}\boldsymbol{\theta} - \frac{\mathbf{P}\boldsymbol{\zeta}\varepsilon_1}{m_\upsilon^2} \tag{11}$$

$$\dot{\mathbf{P}} = -\frac{\mathbf{P}\boldsymbol{\zeta}\boldsymbol{\zeta}^{\mathrm{T}}\mathbf{P}}{m_\upsilon^2} + \left(\lambda\mathbf{P} - \frac{\mathbf{P}^2}{R_\upsilon^2}\right)\bar{\mu}^2 \tag{12}$$

with $\mathbf{P} = \mathbf{P}^{\mathrm{T}}$ such that

$$0 < \mathbf{P}(0) \leq \lambda R_\upsilon^2\mathbf{I}, \quad \mu^2 \leq k_\mu\bar{\mu}^2 \tag{13}$$

$$\dot{m_\upsilon} = -\delta_0 m_\upsilon + \delta_1\left(|u| + |\upsilon_o| + 1\right), \quad m_\upsilon(0) \geq \delta_1/\delta_0 \tag{14}$$

where λ, $\bar{\mu}$, R_υ^2 δ_0 and δ_1 are positive constants and δ_0 satisfies

$$\delta_0 + \delta_2 \leq \min\left[p_0, q_0\right] \tag{15}$$

where $q_0 > 0$ is such that the poles of $W_m(s - q_0)$ and the eigenvalues of $\mathbf{F} + q_0\mathbf{I}$ are stable and δ_2 is a positive constant. $p_0 > 0$ is defined in **A2** and σ is given by

$$\sigma = \begin{cases} 0 & \text{if} \quad \|\boldsymbol{\theta}\| < M_0 \\ \sigma_0\left(\dfrac{\|\boldsymbol{\theta}\|}{M_0} - 1\right) & \text{if} \quad M_0 \leq \|\boldsymbol{\theta}\| \leq 2M_0 \\ \sigma_0 & \text{if} \quad \|\boldsymbol{\theta}\| > 2M_0 \end{cases} \tag{16}$$

with

$$\sigma_0 \geq 2\bar{\mu}^2/R_\upsilon^2 \tag{17}$$

where M_0 is an upper bound $\|\boldsymbol{\theta}^*\|$ (see Assumption **A5**).

As stated before, the main control problem consists in determining an appropriate control parameter vector so that the resulting closed-loop plant is stable and the plant output υ_o tracks the model reference output υ_m as closely as possible, i.e., the desired closed-loop performance is clearly expressed in terms of the model reference's choice.

Unfortunately, there is no rule of thumb for designing $W_m(s)$ but a common approach relates the model reference to the dynamic of the open loop system (see [9] and [10]). More specifically, selecting $W_m(s)$ with smaller rise and settling time than the modeled part of the plant (Eq. (18)) usually results in good closed-loop performance and also avoids overloading the actuator during transient periods. Then,

$$W_m(s) = \frac{9.87 \times 10^8}{s^2 + 3.96 \times 10^4 s + 9.87 \times 10^8} \tag{18}$$

Fig. 4 indicates the step responses of $G_{\upsilon o}(s)$ and $W_m(s)$. Comparing these results, one can conclude that the performance of the closed-loop system is now mainly dependent on how the nominal model of the AC power source load represents the actual electrodynamic shaker and, more importantly, on how the parameter adaptation algorithm can deal with such differences.

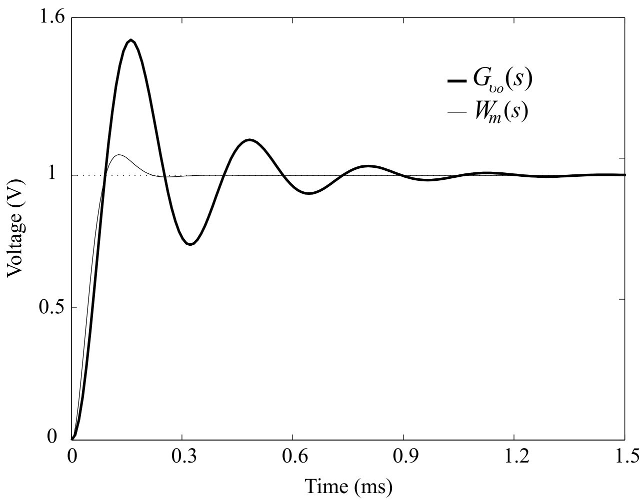

Fig. 4. Step response of $G_{\upsilon o}(s)$ and $W_m(s)$.

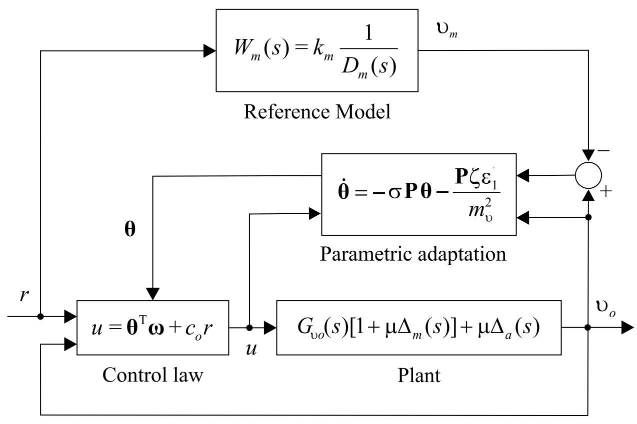

Fig. 5. The AC power source voltage controller.

The modelled part $G_{\upsilon o}(s)$ of the plant has a magnitude increase at high frequency operation due to resonances that occur between the output filter's inductor and capacitor. In practice, this amplification can be dealt as an increase of k_p and consequent necessity of reducing the control parameter c_o as long as k_m remains constant (according to the model reference control theory [11], $c_o = k_m/k_p$). Then, differently from [12], the following modification is introduced in the robust adaptive control,

$$c_o(f) = \begin{cases} 2.2 - (1.7)^{f/500} & \text{if} \quad f \leq 500 \text{ Hz} \\ 0.5 & \text{if} \quad f > 500 \text{ Hz} \end{cases} \tag{19}$$

where f is the frequency of the reference r.

By doing so, the proposed voltage control system indicated in Fig. 5 is capable to guarantee good reference tracking and stability both at low and high frequency operation. Table I summarizes the main control parameters.

TABLE I
Parameters of the adaptation algorithm

Parameter	Symbol	Value
Control parameter vector	$\boldsymbol{\theta}(0)$	$[\ -1 \quad 0.3 \quad 0.7\ ]^{\mathrm{T}}$
Covariance matrix	$\mathbf{P}(0)$	$100 \times \mathbf{I}_{3\times3}$
Regression vector	$\boldsymbol{\zeta}(0)$	$[\ 0 \quad 0 \quad 0\ ]^{\mathrm{T}}$
Dynamic matrix	$\mathbf{F}$	2000
Weighting matrix	$\mathbf{q}$	2000
Filter of u	$\boldsymbol{\omega}_1(0)$	0
Filter of v_o	$\boldsymbol{\omega}_2(0)$	0
Constant design parameter	λ	10
Limitation factor	$\overline{\mu}$	0.1
Covariance matrix limitation factor	R	10
Constant design parameter	δ_0	0.991
Constant design parameter	δ_1	1
σ function limitation factor	M_0	9
Upper bound of the σ function	σ_0	0.1

IV. Implementation and Experimental Results

The feasibility of the proposed AC power source and voltage controller has been verified experimentally. Firstly, the robust adaptive control scheme has been written in a recursive form by applying the Euler's approximation to the algorithm's differential equations and z-transformation on the zero-order hold/model reference transfer function (sampling frequency equal to 24 kHz). Then, the proposed voltage control scheme has been applied to compensate the tracking error between the shaker's input voltage v_o and the model reference v_m. Figs. 6(a) and 6(b) show experimental results at two frequencies conditions. As indicated, the proposed solution is capable to achieve good fixed frequency reference tracking performance both in 20 Hz and 2 kHz despite of the the back electromotive force disturbance, the parameter variations and unmodeled dynamics on the plant.

To verify the transient tracking response with frequency varying continuously, a constant amplitude voltage has been employed to drive the electrodynamic shaker. The resonant search test requirements recommended by the standard ANSI/ASAE EP455 [13] have been taken as a practical example of frequency sweep. According to these specifications, the device under test must be submitted to a sinusoidal frequency sweep at the rate of 1 octave/minute, which demands approximately 7 minutes to complete a whole sweep between 20 Hz and 2 kHz. Then, to facilitate the exposition of the results obtained, some particular frequencies were selected and combined in the same figure. Figs. 7(a) and 7(b) show the corresponding measured voltage waveforms from 30 Hz to 100 Hz and from 600 Hz to 2 kHz, respectively. The good dynamic tracking characteristics observed in these results indicates that the proposed solution is capable to be applied on vibration test control by adding an acceleration control scheme.

V. Conclusion

The development of a high performance switching-mode AC power source for vibration test control of an electrodynamic shaker is an important and challenged issue. Low harmonic distortion and good voltage or current tracking performance are required over a wide dynamic and frequency range and can only be obtained by using a high switching rate PWM inverter associated with a sophisticated control technique.

This paper has presented the development of a switching-mode AC power amplifier for armature excitation of an electrodynamic shaker. The proposed solution is based on a voltage controlled single-phase PWM inverter with digital control. A model reference adaptive controller that assumes reduced knowledge of the plant and guarantees robustness to parameters variations and rejection of bounded disturbances is employed. Fixed and swept frequency experimental results showed that the proposed solution is capable to achieve good voltage reference tracking over a wide frequency range (20 Hz to 2 kHz). The development of an advanced acceleration control technique is currently being undertaken to accomplish the proposed vibration test control system. The results will be reported in the near future.

References

[1] C. M. Harris, *Shock and Vibration Handbook*, 4th ed. New York: McGraw-Hill, 1995.
[2] R. Fair and H. R. Bolton, "Analysis and design of electromagnetic moving coil vibration generators," in *Proceedings...*, International Electrical Machines and Drives Conference. Oxford, UK: IEEE, 1993, pp. 529–534.
[3] T. H. Chen, K. C. Huang, and C. M. Liaw, "A high-frequency switching-mode power amplifier for shaker armature excitation," *IEE Electric Power Applications*, vol. 144, pp. 415–422, Nov. 1997.
[4] T. H. Chen and C. M. Liaw, "Vibration acceleration control of an inverter-fed electrodynamic shaker," *IEEE/ASME Transactions On Mechatronics*, vol. 4, pp. 60–70, Mar. 1999.
[5] ——, "Soft-switching inverter for electrodynamic shaker," *IEE Electric Power Applications*, vol. 146, pp. 515–523, Sept. 1999.
[6] J. Han, T. Tang, and X. Wang, "A high-performance switching mode power amplifier for electrodynamic shaker," in *Proceedings...*, International Conference on Industrial Technology. Hong Kong, China: IEEE, 2005, pp. 491–495.
[7] L. Della Flora and H. A. Gründling, "Digital environment for sinusoidal vibration test control of an ac power source-fed electrodynamic shaker," *Journal of the Brazilian Power Electronics Society*, vol. 11, pp. 167–174, Nov. 2006.
[8] R. Lozano-Leal, J. Collado, and S. Mondié, "Model reference adaptive control without a priori knowledge of the high frequency gain," *IEEE Transactions on Automatic Control*, vol. 35, pp. 71–78, Jan. 1990.
[9] K. J. Astrom and B. Wittenmark, *Adaptive Control*. Addison Wesley, 1989.
[10] I. D. Landau, R. Lozano, and M. M'Saad, *Adaptive Control*. Springer, 1998.
[11] P. A. Ioannou and J. Sun, *Robust Adaptive Control*. Upper Saddle River: Prentice Hall, 1996.
[12] L. Della Flora and H. A. Gründling, "Acceleration control of an inverter-fed electrodynamic shaker," in *Proceedings...*, Power Electronics Specialists Conference. Jeju, Korea: IEEE, 2006, pp. 2799–2805.

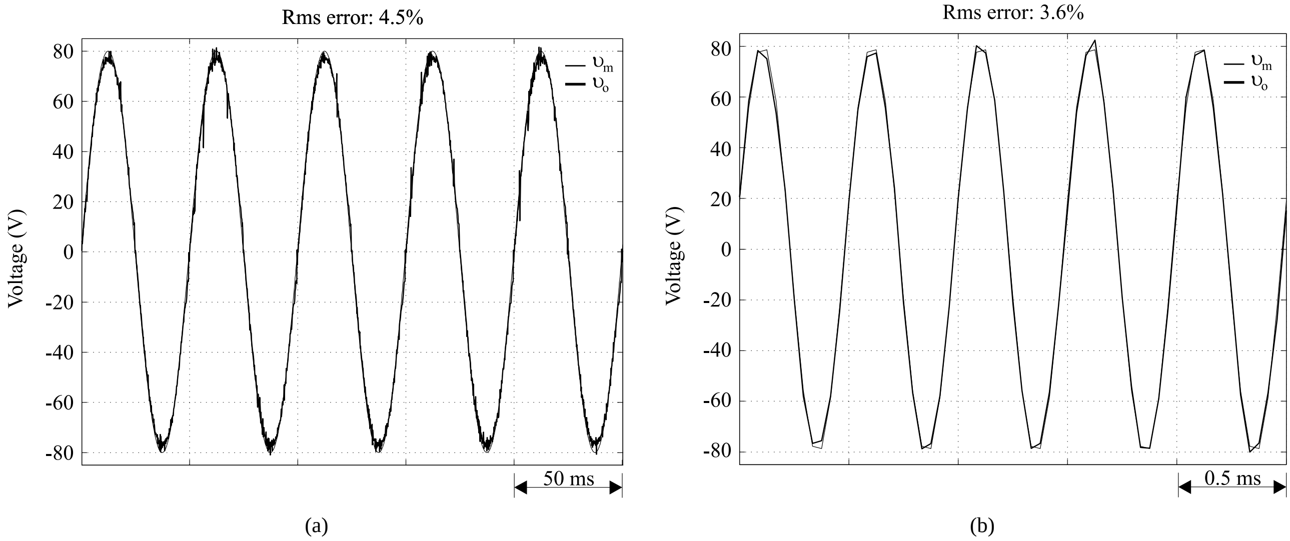

Fig. 6. Reference model output v_m and the measured voltage waveform v_o. (a) 20 Hz. (b) 2 kHz.

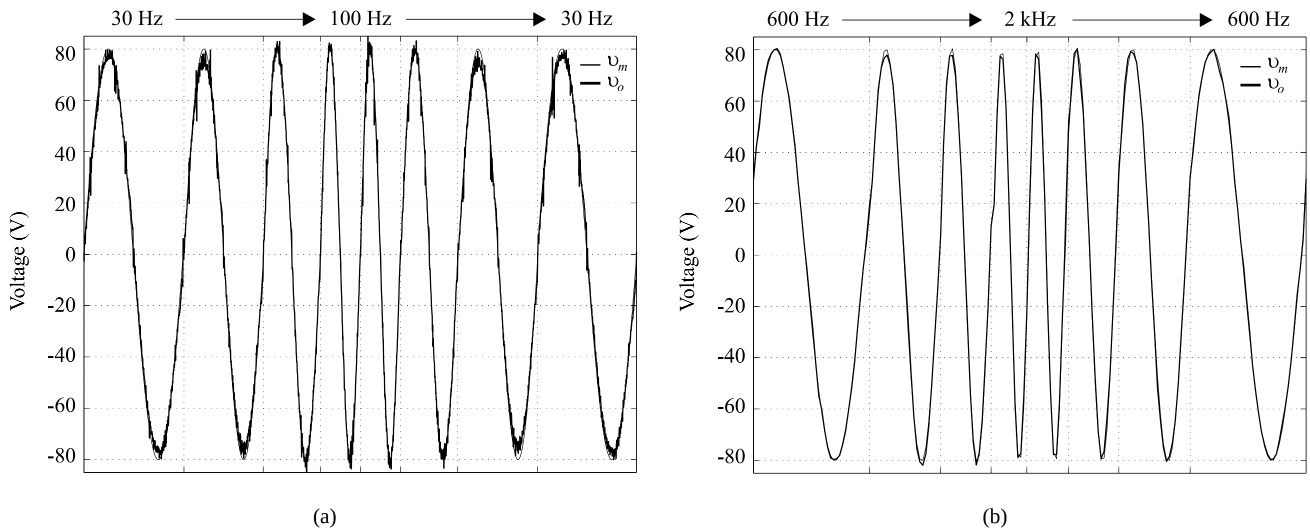

Fig. 7. Reference model output v_m and the measured voltage waveform v_o with frequency variation. (a) 30 Hz to 100 Hz. (b) 600 Hz to 2 kHz.

[13] *Environmental Considerations in Development of Mobile Agricultural Electrical/Electronic Components*, ANSI/ASAE EP455, St Joseph, July 1991.