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Self-tuned driving of piezoelectric actuators The case of ultrasonic motors

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Abstract

A number of piezoelectric devices with wide applicability in industrial and technological areas, i.e. ultrasonic motors and piezoelectric transformers, are based on piezoelectric resonators. Piezoelectric resonators are electromechanical structures for which the operation point is close to resonance. This operation point must be kept close to resonance regardless of external perturbations. This paper introduces the mechatronic design of control electronic circuits for piezoelectric resonators. The design approach has been experimentally validated on ultrasonic motors. The paper analyzes the specification of optimal driving signals, the digital electronic generation of such signals and the self-tuning characteristics of these electronic drivers based on feedback loops.

Keywords: Piezoelectric properties; PZT; Actuators; Ultrasonic motors

1. Introduction

Ultrasonic motors are particular instances of piezoelectric or electroceramic resonators. This is also the case of piezoelectric transformers in which a mechanical vibration in resonance (obtained through the application of the direct piezoelectric effect) is transformed into an amplified voltage (through the application of the converse piezoelectric effect). In the case of ultrasonic motors, a resonant piezoelectric ceramic is used to excite travelling waves on an elastic substrate, which in turn are transmitted to the motor's rotor through frictional forces.[1]

In both cases, the piezoelectric ceramic is driven close to resonance. It is well-known that the resonance parameters (frequency, modal shape, modal damping . . .) are dependent on both internal (temperature) and external (load) conditions.[2] In the particular case of ultrasonic motor these conditions are mainly temperature fluctuations due to frictional heat and externally applied torque.

Under these driving conditions it is extremely important to have power and control electronics of sufficient robustness so that external load or temperature conditions do not affect the operational point of electroceramic resonators.[3] The design of power electronics has been presented by the authors,[4] where the electrical impedance was matched between the piezoelectric resonator and the power driver. In so doing, the electromechanical and material properties of the resonator were taken into account in the design of a matched power stage.

However, the control driver for these systems must be able to take into account possible deviations between driving and resonant frequency during operation and make them be as close as possible. For this to be done, the intrinsic resistive nature of piezoelectric resonator in the vicinity of resonance and antiresonance is taken into account, so that self-tuning performance of the control drive is not dependent on electromechanical and material characteristics of the resonator.

In addition to the self-tuning characteristics, it is also of importance that control signals with low harmonic content can be generated. Frequency harmonics present in the power signals result in reduced efficiency of the electroceramic resonator. The resonator will act as an electrical filter when driven in resonance and harmonics will contribute to global heating.

This paper addresses in a first section the analysis of optimal control signals for electroceramic resonators. Optimal signals are shown to be digital and with low harmonic content so that heating is minimized. In a second section, the issue of self-tuned power drivers will be studied and a possible solution based on

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phase locked loops will be proposed. It is shown that the driving frequency is maintained close to resonance regardless of temperature or load perturbations. Eventually, we will discuss the implementation of optimal control signals and self-tuned control drivers on an untrasonic motor and will draw some conclusions.

2. Optimal power signals for electroceramic resonators

In general, control signals for electroceramic resonators can be analogue or digital. When control signals are digitally generated, their main parameters (frequency, amplitude and phase lag) can be easily controlled. However, when digital controllers are used it is in general more difficult to obtain adequate filters in the complete range of control frequency and it is the electroceramic resonator that will filter out undesired harmonics. This, as mentioned before, leads to heating and reduced efficiency. Once proper control signals are generated, they are amplified and converted into power signals to drive the resonator.[4]

A schematics representation of this power driver is shown in Fig. 1. In this figure, the inverter driving stage is in charge of generating proper digital control signals. The series and parallel resonant converters are introduced to, on the one hand, amplify the control signal to provide power signals, and on the other hand, to match the electrical impedance of power driver and electroceramic resonator with the aim of increasing overall efficiency.[3,4]

In general, the use of both the series and parallel resonant converters will introduce additional resonance electrical branches that can be excited during operation. In order to avoid, or at least reduce the chance of exciting parasitic resonances, the digital control signal of the inverter must be carefully chosen.

It has been studied and discussed that the parasitic resonance corresponding to the series resonant converter of Fig. 1 is usually of lower frequency than the resonance frequency of the electroceramic resonator.[3] This parasitic resonance will only affect the operation of the power driver during transient situations (when switching on the system). However, the parasitic resonance corresponding to the parallel resonant converter will in general be of higher frequency than the fundamental frequency of the electroceramic resonator. Any harmonic component in the power signal might lead to excitation of this second parasitic resonance and must be avoided.

Digital signals with low harmonic content must be selected as control signals in the inverter stage. If we consider for instance bipolar symmetrical control signals, all even harmonics are zero.

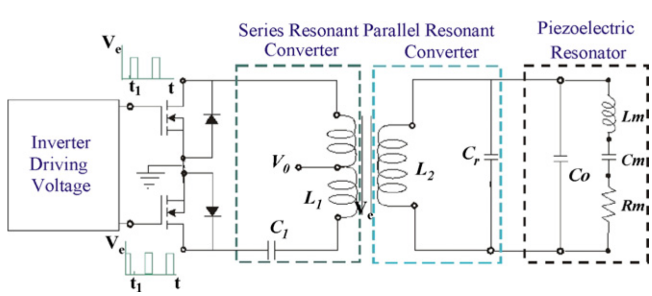

Fig. 1. Schematic representation of the signal amplification and tuning circuit.

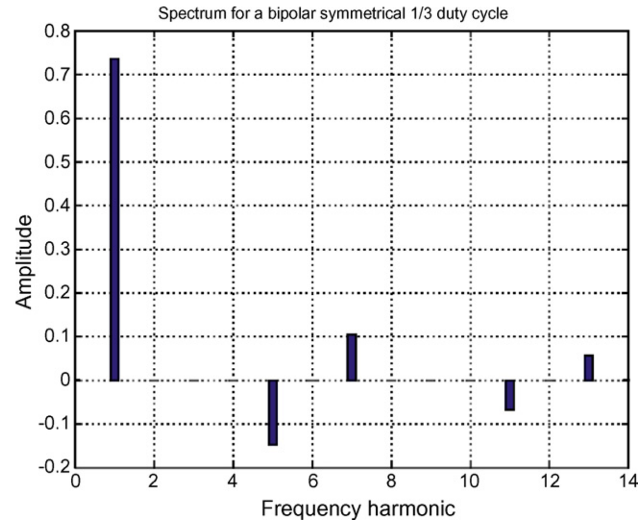

Fig. 2. Amplitude of the various different harmonics in the spectrum of a switched bipolar symmetrical signal with a duty cycle of 1/3.

In addition, if the duty cycle (ratio of high level to period) of the switched control signal is set to 33%, the frequency spectrum of the control signal can be demonstrated to be:

$$V_n = \frac{4}{\pi} V \frac{\sin n\pi\omega}{n} \frac{[1-(-1)^n]}{2} \qquad (1)$$

where V_n is the amplitude of the nth harmonic, V the amplitude of the driving voltage and ω is the duty cycle of the power signal. Fig. 2 shows the frequency content of this switched control signal.

Under these conditions, the first non-zero harmonic is the fifth one. Therefore, the combination of bipolar symmetrical switched signals with the appropriate duty cycle results in adequate control signals with a frequency content close to a pure tone tuned to the fundamental frequency of the electroceramic resonator.

3. Phase locked loop based self-tuned drivers

An electroceramic resonator is characterized by a capacitive electrical load in a frequency range below the resonance frequency and above the antiresonance frequency. In between, the electrical equivalent to the piezoelectric ceramic is inductive. This means that the piezoelectric ceramic becomes a pure resistive electrical load at resonance and antiresonance. This is true for all resonators and is consistent with the fact that at resonance and antiresonance the reactive part of the electrical impedance vanishes, thus producing peak efficiency. Therefore, the method presented in this section for tuning driving frequency to the resonance characteristics of the piezoelectric resonator is general and does not depend on electromechanical of material properties rather than the predominant resistive nature of resonators in the vicinity of resonance and antiresonance.

The resonance and antiresonance frequency of a piezoelectric actuator is generally subject to perturbations during operation. For optimal operation, a tracking electronic drive is required. The functional characteristics of such a tracking drive are

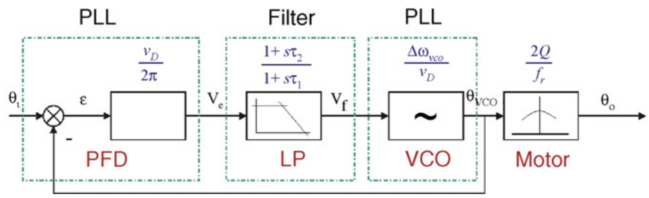

Fig. 3. Block diagram for the PPL based control loop.

depicted in the block diagram of Fig. 3. The blocks in the tracking system are:

(1) *Error phase detector*: The role of this block is to provide an error signal proportional to the phase error between driving voltage and current drawn.
(2) *Loop controller*: The loop controller receives the phase error between current and voltage as an input and provides a control signal that asymptotically tracks the resonance frequency of the piezoelectric actuator.
(3) *The plant*: This block represents the piezoelectric actuator itself. It receives the control action and performs at resonance independently of external perturbations.

In practical terms, the design of a resonance frequency tracking electronic drive is commonly based on a phase locked loop, PLL, technique. A phase locked loop consists of two main building blocks, namely the phase detector, PD, and the voltage controlled oscillator, VCO.

In a common implementation, a phase detector works as an up–down counter, in which the up count is edge triggered by the first input signal (the current: see the upper line in Fig. 4) and the down count is also edge triggered by the second input signal (the voltage: see the middle line in Fig. 4).

According to this description, the phase detector output is made up of a train of pulses, whose width is proportional to the phase lag between voltage and current drawn (see the lower line in Fig. 4).

The phase detector provides an output signal proportional to the difference in phase between current drawn and voltage:

$$v_e = K_d(\theta_i - \theta_0) = K_d\varepsilon \quad (2)$$

$$\frac{V_e(s)}{\varepsilon(s)} = K_d \quad (3)$$

where K_d is the phase detector gain.

In order to obtain the desired performance from the error phase detector in Fig. 3, a low pass filter can be added to the output of the PLL phase detector. The combination of the phase detector and the low pass filter will produce an analogue error signal proportional to the phase difference between voltage and current.

The voltage-controlled oscillator generates a switching signal with a fundamental frequency proportional to its input voltage. The duty cycle of the switching signal can generally be selected to suit the requirements of the application and will not interfere with the phase detector described above. As explained in the previous section, a suitable choice for the duty cycle would be around 1/3.

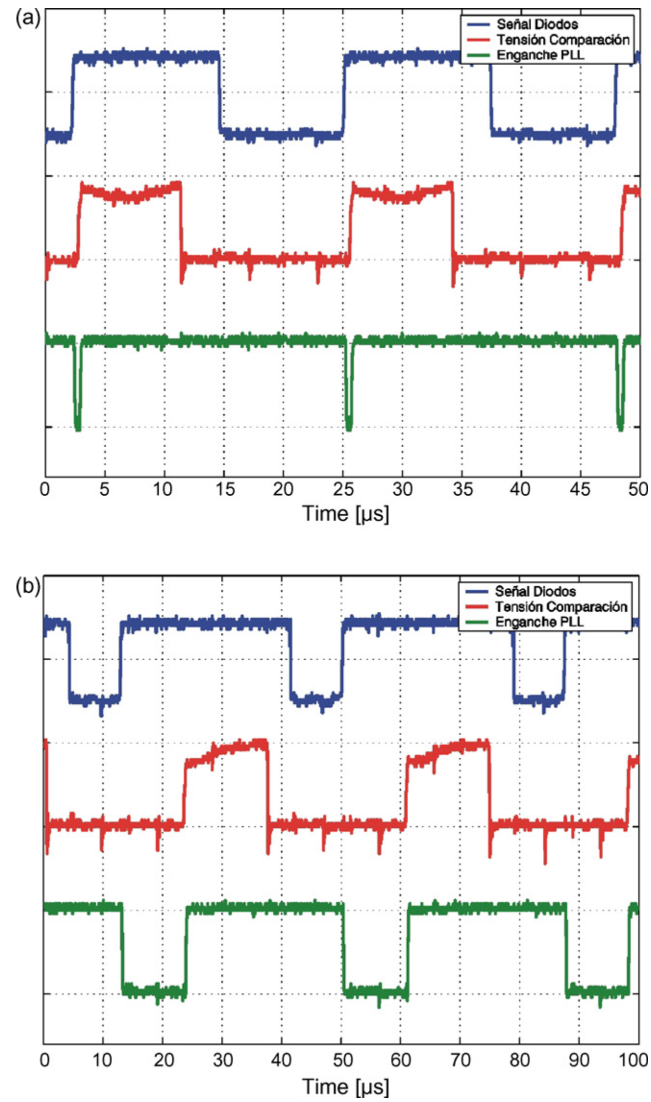

Fig. 4. Frequency tuning through the use of PLL close control loops for ultrasonic motors.

The voltage-controlled oscillator will provide a deviation from the central frequency that is proportional to the analogue error signal from the phase detector and loop filter, v_f. This is mathematically expressed as:

$$\Delta\omega = K_0 v_f \quad (4)$$

Since frequency is a derivative of phase, the above formulation for the functional characteristics of the voltage-controlled oscillator can be described as:

$$s\Theta_{vco}(s) = \frac{K_0}{V_p(s)} \quad (5)$$

It can readily be appreciated that Eq. (5) describes the voltage-controlled oscillator as a pure integrator. As a result, the close loop transfer function of the tracking loop is always a Type I loop.

If the tracking circuit implements an appropriate controller together with the voltage-controlled oscillator in a cascade configuration, this will give the functionality of the loop controller in Fig. 3. To be able to reject permanent dc components in the error signal, the appropriate low pass filter must be selected.

In the simplest configuration, a classic passive low pass filter can be implemented. As discussed by Gardner,[5] a passive low pass filter gives good results in most tracking applications.

The transfer function for the low pass filter is:

$$\frac{V_f(s)}{V_e(s)} = \frac{1+s\tau_2}{1+s\tau_1} \tag{6}$$

The open loop transfer function for the tracking circuit can be written as:

$$G(s) = K_d \frac{K_0}{s} \frac{1+s\tau_2}{1+s\tau_1} \tag{7}$$

The closed loop transfer function corresponding to the tracking circuit comprising the phase detector (Eq. (3)), the loop low pass filter (Eq. (5)) and the voltage controlled oscillator (Eq. (5)) is:

$$H(s) = \frac{G(s)}{1+G(s)H(s)} = \frac{K_d K_0/\tau_1(1+s\tau_2)}{s^2 + s(1+K_0K_d\tau_2)/\tau_1 + K_0K_d/\tau_1} \tag{8}$$

The stability of the tracking system can readily be shown using the Final Value Theorem.[1]

For the particular case of classic low pass passive filters, the system is Type I, i.e. there is one perfect integrator in the tracking loop—the one provided by the voltage-controlled oscillator. When directly applied, the Final Value Theorem predicts that the system will track step changes in the phase error without permanent phase error in the steady state.

However, the system will lead to permanent constant error in the steady state if a ramp phase error is introduced. If we wish to ensure zero tracking error upon application of ramp phase errors, the loop tracking circuit should be completed either by a classic PI controller or by making use of active filters with a transfer function:

$$\frac{V_f(s)}{V_e(s)} = \frac{1+s\tau_2}{s\tau_1} \tag{9}$$

In that case, the tracking loop would be converted to a Type II system comprising two perfect integrators. Consequently, ramp changes in the phase error would be perfectly tracked without permanent steady state errors.

4. Experimental results

The experimental verification of the design method of power electronic drivers for electroceramic resonators has been conducted on the well-known USR-60 Shinsei's ultrasonic motor. A control and power electronic driver has been designed and prototyped. The schematic representation of Fig. 1 shows the various different circuit design parameters. The values of the inductance and capacitors of the series and parallel resonant converter have been adjusted to the USR-60 equivalent circuit so that the electrical impedance of the power stage and the load (the USR-60 ultrasonic motor) are matched. The electromechanical and material properties of these ultrasonic motors can be found in a previous paper from the authors.[4]

The control stage (the inverter in Fig. 1) has been implemented on a HCF4046B phase locked loop from STMicroelectronics. This PLL circuit includes both a voltage controlled oscillator and a phase detector.

The phase detector works on the voltage and the current input to the ultrasonic motor. It is an up–down counter and delivers a pulse of duration equal to the time between up flanges in the voltage and current in the motor. This switched frequency is filtered so that a voltage proportional to the phase lag between both signals is obtained.

This phase lag analogue signal controls the voltage controlled oscillator: a phase lag will lead a frequency change in the direction of minimizing this phase lag, and thus taking the ultrasonic motor to its resistive frequency range, which corresponds to resonance or antiresonance.

The behaviour of the self-tuning power electronic driver has been analyzed on the USR-60 ultrasonic motor. Fig. 4 shows the situation when the PLL is used in the closed loop control of the driving frequency. In Fig. 4a we show the situation in which the PPL drives the motor at resonance. The upper and middle switched signals correspond to the voltage and current drawn. It is clearly seen that the phase lag is very low and this results in a high level value of the switched signal corresponding to the phase detector output (lower signal in Fig. 4a).

On the other hand, when the PLL is off and there is a slight change in the external conditions, the frequency will not follow these changes and the situation depicted in Fig. 4b will take place. Here, current drawn and voltage show a clear phase lag corresponding to a capacitive behaviour of the ultrasonic motor, i.e. the motor being driven out of resonance.

5. Conclusions

This paper presented the selection of optimal control switched signals to drive electroceramic resonators. In addition, the paper proposed a closed loop implementation based on phase locked loops to obtain a self-tuned power driver for these resonators. The design takes into consideration the particular resistive electrical behaviour of piezoelectric ceramics at resonance to set up a tuning procedure based on the minimization of phase lag between voltage and current drawn.

A paradigmatic case of piezoelectric resonator, the USR-60 ultrasonic motor, has been used to experimentally verify the self-tuning characteristics of the power drive. It is shown that the PLL based control loop tracks frequency changes in the resonance characteristics of the electroceramic resonator and contributes to ensuring a tuned operating point.

References

1. Pons, J. L., *Emerging Actuator Technologies. A Micromechatronic Approach*. John Wiley & Sons, Chichester, 2005.
2. Pollard, H. F., *Sound Waves in Solids*. Pion Ltd., 1977.
3. Schaaf, U. and van der Broek, H., Piezoelectric motor fed by a PLL-controlled series resonant converter. In *European Conference on Power Electronics and Applications*, 1995.
4. Pons, J. L., Fernández, J. F., Villegas, M., Ochoa, P., Ceres, R., Calderón, L. and Rocon, E., Tuned driving of piezoelectric resonators: impedance matching. *Bol. Soc. Esp. Ceram. V*, 2006, **45**(3), 178–183.
5. Gardner, F. M., *Phaselock Techniques*. John Wiley & Sons, New York, 1979.