Design of Power Supply for Driving High Power Piezoelectric Actuators

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Abstract — Subject of this contribution is a novel PWM controlled inverter of the kW power range, feeding a multi-mass ultrasonic motor (MM-USM) via a LLCC-type filter. The system is designed in a way to reduce the total harmonic distortion of the motor voltage and to locally compensate for the reactive power of the motor. Operating a three-level inverter with PWM pattern of eliminated low-order harmonics allows the use of a small and lightweight output filter showing besides improved dynamic behavior.

Keywords — High frequency power converter, piezo actuators, pulse width modulation (PWM), voltage source inverters (VSI), industrial application.

I. INTRODUCTION

High power piezoelectric actuators are used to build various kinds of piezoelectric systems like ultrasonic motors and sonotrodes for ultrasonic machining. Due to their high force and power density, they are applied more and more in aircrafts and industry. These actuators are normally constructed using solid-state piezoelectric multilayer ceramics which converts electrical energy directly into mechanical energy through linear motion. However, piezo ceramics is sensitive to positive tension. Thus, either mechanical or electrical biasing is required. In case of applying mechanical pre-stress, the actuators can be supplied only with AC voltage omitting DC-biasing. Operation at mechanical resonance frequency aims for best efficiency and movement.

Considering the advantages of piezoelectric actuators to generate high torque at low speed and this with highest force and power density, they are expected as novel technology for airborne brakes. Therefore, the EC funded project PIBRAC [1] was started to study, design and test a piezoelectric brake actuator and its involved control electronics. The yield should be distinct cuts in total weight and peak power demand compared to most suitable electromagnetic motors (PMSM). The function of the MM-USM, its assembly and detailed structure is outlined in [2] [14].

The multi-mass ultrasonic motor (MM-USM) consists of two pairs of stator rings and two rotor discs connected to the shaft. One rotor disc is sandwiched between a pair of stator rings, as shown in Fig. 1. Each stator ring consists of eight metallic blocks connected with eight piezo-ceramic multilayer stacks with alternating polarization. The structure is excited by the piezo stacks at its mechanical eigenfrequency which is designed for 35 kHz (called tangential mode) so that each metallic block would oscillate in the plane of the ring. The structure is able to oscillate also orthogonally to the surface of the stator rings and rotor disk (normal mode) excited by normal piezo stacks at the same frequency as the tangential mode, but with appropriate phase shift. Thus, elliptical movements of the metallic blocks will result that generate thrust by temporary clamping of the disks via a frictional layer (see Fig. 1(b) and 2) [1] [2]. Note, that only one normal mode actuator is required for two pairs of tangential mode actuators, since the force feedback is provided by the coupling disc.

This contribution puts focus on the power supply and its control architecture for driving high power piezoelectric actuators. A three-level LLCC-PWM inverter is presented, which is developed by combining a LLCC filter and a

![Fig. 1. Structure of multi-mass ultrasonic motor](image-url)
single-phase PWM inverter driven by an optimal pulse width modulation. The novel solution offers significant advantages to improve the performance of the power supply. The piezoelectric stacks of the motor form an equivalent parallel capacitance $C_p$, which becomes part of the LLCC filter. In each direction one piezoelement is used as sensor for obtaining the information of tangential and normal deflections $u_T, u_N$. The transformer leakage inductance and the inductance of the connection cable between inverter within the fuselage and the brake actuator at the wheel are utilized as parts of the series inductance. A power supply prototype capable of converting a DC input voltage of 270 V into frequency and magnitude controlled motor voltages is implemented for a tangential and a normal excitation system. Experimental results are presented for a standard multilayer piezoelectric actuator driven in a frequency range of 30 to 40 kHz.

Over the last decade electronic power supplies for piezoelectric system have been well studied and applied using different kinds of resonant converter concepts. A resonant inverter with rectangular voltage shape combined with LLCC-type output filter presented in [3], [4] shows advanced characteristics and best suited properties in respect to efficiency, stationary and dynamic behavior, as well as to control and commissioning efforts. The drawbacks of these resonant inverters are the large volume, heavy weight and high cost of the magnetic components of the resonant filter such as transformer and inductor, especially in case of driving high power piezoelectric actuators of the kW range [9]. The aforementioned disadvantages can be avoided, if a PWM inverter is used in conjunction with the LLCC filter, since the filter transfer function is very robust to variations of the parallel capacitance. In addition the power supply enables a broadband characterization of the MM-USM [13].

The fundamental frequency of the power inverter is determined by the mechanical resonant frequency which is in a range of 20 to 40 kHz. This implies that the inverter switching frequency is in the range of 140 to 300 kHz, which produces consequently more switching losses and eventually ends up with EMC issues. To overcome these issues related to two-level PWM converters, either an optimal pulse width modulation or a multi-level inverter technology are employed for the inverter enhancement [7] [8] [11] [12].

The proposed three-level LLCC-PWM inverter [13] shown in Fig. 3 is investigated to excite and control the tangential mode of the high power piezoelectric actuator. The novel solution is designed to attain an optimal performance of the power supply for driving a high power piezoelectric actuator, attributed by the following features:

1) The total harmonic distortion (THD) of the actuator voltage is reduced to enhance piezoelectric stack lifetime without increasing the switching frequency compared with a LC-PWM inverter [5].

2) The reactive power of the piezoelectric actuator is compensated locally, by placing inductor $L_p$ close to the actuator. Hence, cables of considerable length linking the transformer output and actuators can be rated only with respect to the real power, which saves considerable weight and costs [6].

3) Due to the PWM methods and reactive power compensation, the output filter shows an optimized performance at minimized volume and weight, if compared to a resonant operated inverter, see [3] [4].

4) At the same time the approach allows wide bandwidth characterization of the actuator for means of commissioning without changing the filter components.
B. Three-level optimal PWM

Suitable switching angles can be calculated to eliminate selected harmonics of the output voltage [11] [12]. The output voltage $u_{\text{filter}}(t)$ of the three-level converter is assumed as quarter-wave symmetrical (Fig. 4 lower part). As a result of this symmetry, only harmonics of odd orders can occur. The according complex Fourier coefficients result as

$$\hat{U}_{\text{filter}} = \frac{2}{j\nu T} \sum_{i=1}^{n} (-1)^{i+1} \cos (\nu \omega t) \quad \nu = 1, 3, \ldots (1)$$

Due to the quarter-wave symmetry, all coefficients consist only of imaginary parts. In order to eliminate the $3^{rd}$, $5^{th}$ and $7^{th}$ harmonics and to ensure a certain fundamental content, four independent switching angles $0 \leq \alpha_1 \leq \ldots \leq \alpha_4 \leq \frac{\pi}{2}$ are required. Please note that transistors $S_5$ and $S_6$ are used only to change the polarity of the output voltage. This leads to a system of four nonlinear equations that can be solved applying Newton’s method as follows [12].

$$\alpha_{j+1} = \alpha_j - (\hat{U}'_{\text{filter}}(\alpha_j))^{-1} (\hat{U}_{\text{filter}}(\alpha_j) - M) \quad (2)$$

$$M = \left[ \begin{array}{ccc} U_{\text{filter}}^* & 0 & 0 \\ \alpha_1 & \alpha_2 & \alpha_3 \\ \alpha_4 \end{array} \right] \quad \text{reference values} \quad (3)$$

$$\alpha = \left[ \begin{array}{c} \alpha_1 \\ \alpha_2 \\ \alpha_3 \\ \alpha_4 \end{array} \right] \quad \text{switching angles} \quad (4)$$

$$\hat{U}_{\text{filter}}(\alpha) = \left[ \hat{U}_{1,\text{filter}} \ \hat{U}_{3,\text{filter}} \ \hat{U}_{5,\text{filter}} \ \hat{U}_{7,\text{filter}} \right] \quad \text{harmonics} \quad (5)$$

$$\hat{U}'_{\text{filter}}(\alpha) = \frac{\partial \hat{U}_{\text{filter}}(\alpha)}{\partial \alpha_1, \ldots, \partial \alpha_4} \quad \text{Jacobian Matrix} \quad (6)$$

E.g., for the positive halfwave with positive current, $S_6$ is turned on and pulses are generated only using $S_1$ and $S_2$, while $S_3$, $S_4$ and $S_5$ keep completely turned off.

Fig. 5 presents the harmonics of $u_{\text{filter}}$ as they result from the harmonic elimination modulation (HEM) compared with a conventional carrier based modulation (CBM). As expected, the harmonics of orders 3, 5 and 7 are zero with HEM while non-zero with CBM. For higher orders, however, the CBM may generate lower harmonics than HEM. The harmonics of the actuator voltage $u_{\text{actuator}}$ can be calculated from the inverter harmonics by multiplication with the LLCC filter transfer function also shown in Fig. 5. Because the high-order harmonics are strongly attenuated by the filter, the actuator THD will result much lower with HEM compared to CBM.

C. Balancing of capacitive voltage divider

As shown in Fig. 6, the energy buffer consists of a capacitive voltage divider. It is essential that the voltage across the capacitors must be balanced. Otherwise the output voltage would get a DC component and even harmonic components could be introduced with the consequence of increased THD.

Thus, a controller must take care of DC balancing so that $\Delta u$ results as zero.

$$U_{\text{in}} = u_1(t) + u_2(t) \quad (7)$$

$$\Delta u = u_2(t) - u_1(t) \quad (8)$$
Three different cases are defined depending on the voltage deviation:

- **Balanced case:** \(|\Delta u| = 0\) to \(5\) V
  Modulation used here is same as explained in the section II-B.
- **Unbalanced case:** \(|\Delta u| = 5\) to \(20\) V
  In this case \(\Delta u\) has to be actively balanced by a controller; a detailed explanation is followed in this section.
- **Failure case:** \(|\Delta u| = 20\) V to \(U_{\text{in}}\)
  Protection shutdown.

The voltage difference \(\Delta u\) is influenced by the \(i(t)\) (Fig. 6) according to eq. (9b). In the final state, the voltage difference \(\Delta u\) is expected to be zero.

\[
\begin{align}
  i(t) &= i_1(t) - i_2(t) \quad (9a) \\
  i(t) &= C_1 \frac{du_1(t)}{dt} - C_2 \frac{du_2(t)}{dt} \quad C = C_1 = C_2 \quad (9b) \\
  \Delta u(t) &= \frac{\Delta u_0}{\text{final state}} - \frac{1}{C} \int_0^t i(t) dt = k T_m I_w \\ 
  (9c)
\end{align}
\]

The current \(i(t)\) can be varied, if the switching states connecting the middle point of the capacitors to the output are omitted. (These are the states, when only \(S_2\) or \(S_3\) is turned on.) Doing that way during the positive halfwave, \(\Delta u\) would increase, skipping these states during the negative halfwave decreases \(\Delta u\). In practice, such balancing operation may take several periods. So it is necessary to calculate new switching angles for that case.

Fig. 7. Typical voltage characteristic

The shape of the output voltage is no longer an odd function. Thus, also the coefficients of even harmonics of orders 2 to 8 have to be forced to zero, introducing four switching angles \(\beta_1\) to \(\beta_4\) in addition to \(\alpha_1\) to \(\alpha_4\) resulting in a total set of eight independent variables. The angles \(\alpha_5\) to \(\alpha_8\) and \(\beta_5\) to \(\beta_8\) will again result from the waveform symmetry. As a result of this new pulse pattern, the output voltage is not free of a DC bias. However, the DC content will be sufficiently suppressed by the LLCC filter.

The set of eight nonlinear equations is again solved applying Newton’s method yielding results as shown in Fig. 8.

A simulation is carried out to evaluate the performance of these modulation method. For both approaches (HEM and CBM) the THD of the voltage waveform versus \(u_{\text{CP}}\) is calculated and plotted in Fig. 9. From Fig. 5 it is clear that the THD of the HEM is always lower compared to the CBM. This is true for the whole operating region. On a closer look to Fig. 9 it is obvious that the lowest frequency harmonics (in case of CBM) are not strongly attenuated. These harmonics are the major contributor for the THD. However, in case of HEM the 2nd to 8th harmonics are zero, and hence the lower THD results.

**D. Control scheme of power supply**

The power supply control loop forms the inner control loop of the whole piezoelectric brake actuator control system. Therefore the task of the inner loop is to control the voltage amplitude and operating frequency of the tangential and normal modes of the motor; additionally the phase angle must be controlled between these two modes in order to ensure efficient thrust generation by proper superpositioning of the tangential and normal mode movements. It is controlled in a way to generate a phase shift of 90° between tangential deflection and the normal force.
A cascaded voltage and current control scheme is designed to satisfy aircraft brake system requirements and provide a proper amount of flexibility for commissioning. As shown in Fig. 10 the reference variables are motor voltage $u_{Cp}$, $(u_{Cp,s}$ and $u_{Cp,c})$, frequency $f_p$ and phase angle $\Delta \varphi$. The feedback signals are current $i_{filter}$, voltage $u_{Cp}$ and voltage $u_{Pi}$ of a piezoelectric element used as sensor indicating the position of the oscillating mass.

By using a demodulation, the amplitude of these feedback signals is decomposed into sine and cosine components, which yield the same results as the first order Fourier coefficients. The PWM generates switching signals based on reference voltage $u_{filter}$ and provides also the reference of sine and cosine values for the demodulation algorithms.

III. IMPLEMENTATION

The modulation was subdivided in several components on a FPGA, which are described by the block diagram shown in Fig. 11. One component is the Direct Digital Synthesis block (DDS) [10]. The output signal from the DDS is a sawtooth of variable frequency. The trigger block enables changing of nominal values (voltage, frequency) at the end of a period. The switching angles depending on the voltage reference as well as sine and cosine values are stored in look-up tables (LUT), where the sawtooth signal is used as phase input. The sine and cosine signals are used for phase sensitive demodulation of the original measurement signals as described above. Temperature and error monitoring is also included to ensure reliable operation.

An experimental inverter prototype of 1.5 kW power was built to verify the principle of operation, see Fig. 12. The rated output is 270 V (amplitude) at a frequency of 35 kHz, the DC input voltage of 270 V is supplied from the aircraft power grid. Due to the fact that the target motor is still under construction, an equivalent load was used for testing instead. Resulting experimental waveforms of the tangential mode are shown in Fig. 13, showing that the voltage across the piezo elements $u_{Cp}$ are nicely sinusoidal as in the simulation, only the phase between $i_{filter}$ and $u_{filter}$ is slightly different. However, the power factor of the power supply is nearly one. The measured spectrum of the inverter output voltage is shown in Fig. 14(a). As expected, the harmonics of orders 3, 5 and 7 are close to zero with the harmonics elimination modulation, contrary to conventional carrier-based PWM. High-order harmonics, however, may be larger with HEM compared to CBM (e.g. orders 11 and 15). These harmonics have only minor effect on the THD, because they are sufficiently suppressed by the LLCC filter (Fig. 14(b)). Hence, the THD of the actuator voltage $u_{Cp}$ is smaller with HEM (Fig. 14(b)), which was the original objective for gaining a higher lifetime of the piezo stacks.
IV. CONCLUSION

A number of design aspects of a power supply for a high power piezoelectric actuator are outlined in this contribution. The highlights of this power supply are low weight and volume together with minimized THD of the actuator voltages to ensure maximum lifetime of the piezo stacks. In particular this is achieved by combining several techniques such as:

- Three-level inverter technology
- Elimination of low-order harmonics from the PWM
- LLCC filter with matching characteristics
- Compensation of reactive power by a parallel mounted inductor in proximity of the actuator

The design is verified by a prototype setup for a power rating of 1.5 kW.

REFERENCES

[10] Xilinx, LogicCore DDS v5.0, Product Specification, April 2005