# NEW APPROACH TO A SWITCHING AMPLIFIER FOR PIEZOELECTRIC ACTUATORS

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Abstract:
This paper describes a new approach to a switching amplifier for reactive loads (piezoelectric actuators). The amplifier transfers in every switching cycle exactly that portion of energy which is necessary to achieve the desired output value at the load. This approach allows unnecessary switching cycles — which merely heat the transducer and semiconductor components — to be avoided, thus permitting higher system dynamics than those of converters using traditional (PWM or current-mode) controllers. These features can be achieved by explicit evaluation of the differential equation of the converter system in every switching configuration.

## Introduction

The use of piezoelectric actuators as small, high-dynamic drives within mechatronic systems has up to now been limited by the driving amplifier. On the one hand there are small transducers with a high power density, strong achievable forces and short mechanical reaction times. On the other there are bulky, inefficient amplifiers that restrict the system dynamics to keep the power losses within tolerable limits. The reason for these power losses (which increase proportionally to the frequency $f$ ) is the large electrical field energy $W_F$ necessary for large-signal operation. However, more than 70% of the supplied energy remains stored in the actuator and can be recovered if the amplifier is able to work bi-directionally [1].
The best way to show the efficiency of an amplifier for piezoelectric actuators is the comparison with traditional analogue class-AB amplifiers. A class-AB amplifier needs, twice per period, the electric field energy, plus additional energy to compensate the inherent power losses. The power losses $p_V$ are at least

$$p_V = 2 \cdot W_F \cdot f \qquad (1)$$

By using the definition of the energy-equivalent capacitance $C_W$ and equation (1) we get

$$p_V = C_W \cdot U_{PA}^2 \cdot f \qquad (2)$$

The energy-equivalent capacitance assigns to a piezoelectric actuator the same value of capacitance $C_W$ that a linear capacitance which stored the same value of electric energy at the voltage $U_{PA}$ would have [1]. With $C_W$ = 4 µF, $U_{PA}$ = 250 V and $f$ = 1 kHz the power $p_V$ = 250 W is dissipated in the power transistors. The poor efficiency and the bulky heatsinks make such an amplifier unemployable for a mechatronic system.
A switching amplifier is able to build up the electric field in the actuator with minimal power losses and allows recovery of the stored field energy in the case of field reduction.

## Basic discussion

A switching amplifier for reactive loads generally consists of two components. A unidirectional DC/DC converter with a small input power loads a large buffer-capacitor and a second bi-directional DC/DC converter controls the energy alternating between the buffer-capacitor and the reactive load. The requirements on the unidirectional DC/DC converter are few. It only needs to compensate for the power losses of the two stages (about 5% of the applied electric field energy) plus the energy dissipated in the actuator and the connected mechanical system (about 30% of the applied field energy). For the first stage a standard DC/DC converter (flyback) is sufficient. The problem is the second stage. This stage must be dimensioned for full system power and has in the past been the focus of many studies [1,5,6]. Most of the studies use the traditional approaches to switch mode converters (pulse width modulation (PWM), or current mode control) in combination with new controller designs (state-vector control, nonlinear control or digital control). The problem is that the traditional approaches were made primarily for DC/DC converters working from a fixed input voltage to a fixed output voltage. The controller only has to regulate load changing or the changing of the input power line. These fixed frequency switch mode converters are not designed for the fast changing of the command variable (here: voltage across the piezoactuator). In general these transitions are slowed down with respect to the switching frequency (soft start). The main reason for this slowing down is that the switched inductors in the converter system must be prevented from reaching saturation. At saturation the transistor currents will rise uncontrollably and the system will fail. Thus the only way to guarantee the proper operation of a system with fast-changing command variables is to reduce the energy transferred in one switching cycle, which means an increase in the switching frequency. At the same time there must be a certain safety in the design of the magnetic components (overdimension). Therefore the switching frequency for a high bandwidth switching amplifier must be in the order of several hundred kilohertz [6]. Another problem of switch mode converters working with fixed frequencies (PWM or current mode control) is that they require a minimum ohmic load for stable operation. In the case of piezoelectric actuators the ohmic part is not sufficient and the transducers have to be actively discharged whenever the transducer voltage is higher than the desired value [6]. When the output is a DC signal, overcharging and discharging are balanced. Although the output power is zero the switching losses remain, growing with switching frequency, and therefore the efficiency of such an amplifier decreases drastically with increasing frequency of the input signal.

In [4] a circuit is presented which charges a piezoelectric actuator from zero to a fixed output voltage in a single shot. This feature is achieved by calculating the necessary switching time using the system differential equations and the energy balance.

In recent studies [2] the necessary charge and the resulting switching times (maximum currents) are calculated for every switching cycle with an actuator model [3]. The time-critical calculations are implemented in a free-programmable gate array (FPGA).

This paper presents a new approach to switching amplifiers without an arithmetic unit, which nevertheless solves the problems mentioned, namely magnetic saturation and overcharging of the transducer in connection with unnecessary switching cycles.

## Converter Configuration

Fig. 1 shows the basic configuration of the power stage. The converter type described in this paper is a buck-boost converter. However, the switching scheme can by adapted to all kinds of switch mode topologies, including transformer converters. This topology was chosen for the prototype because it

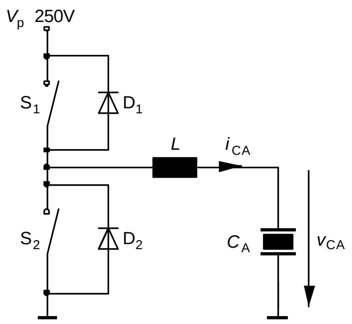


**Fig. 1:** Basic configuration of the converter system

can handle large switching currents (10-16 A) with a small semiconductor stress.

A primary voltage supply supplies the circuit with the constant voltage $V_P$. Two switches $S_1$ and $S_2$ can either connect this voltage with the *LC*-series resonant circuit or short the resonant circuit.

To charge the actuator ($C_A$) the switch $S_1$ must be closed. The current $i_{CA}$ now flows from the supply $V_P$ via $S_1$ and the inductor $L$ to $C_A$ and the voltage $v_{CA}$ rises. At time $t_{off}$ $S_1$ is opened but because of the inductance $L$ a freewheel current through the diode $D_2$ will continue to charge the actuator until time $t_{end}$ ,when $L$ is demagnetized and the diode cut off.

Fig. 2 shows the voltage and current transients for a voltage jump from 100 V to 150 V.

To discharge the actuator, $S_2$ is closed and the actuator stores a portion of its field energy in the inductance $L$. Before the desired voltage $v_{CA}$ at the

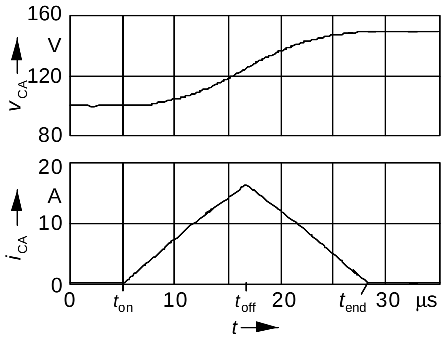


**Fig. 2:** Voltage and current transients for one switching cycle

actuator is reached, $S_1$ must be opened (at $t_{off}$). The freewheel current through the diode $D_1$ now allows the energy to return to the primary supply.

Because of the freewheel currents the voltage at the actuator continues to rise (fall) until the diodes cut off at $t_{end}$. The control unit therefore has to open the switches before the desired voltage at the actuator is reached.

## Calculation of switching times

Normally the nonlinear behavior of a switch mode converter presents the controller designer with a problem. Although the total system is nonlinear, it can be described with a set of linear differential equations. Every switching configuration ($S_1$, $S_2$ closed or $D_1$, $D_2$ conducting) has its own linear differential equations describing the transient behavior of the system. The following calculation is valid for the switching configuration $S_1$ closed. For a more detailed description see [7]. At $t_{on} = 0$ the switch $S_1$ is closed and the voltage at the actuator $v_{CA}(0) \neq 0$, the primary voltage $V_P$ is constant and the current through the inductor $i_{CA} = 0$. The differential equations for this switching configuration are:

$$V_{\mathrm{P}} = L \cdot \frac{\mathrm{d}i_{\mathrm{CA}}(t)}{\mathrm{d}t} + v_{\mathrm{CA}}(t) \tag{3}$$

$$i_{\mathrm{CA}}(t) = C_{\mathrm{A}} \cdot \frac{\mathrm{d}v_{\mathrm{CA}}(t)}{\mathrm{d}t} \tag{4}$$

The Laplace transformation with the initial conditions mentioned above yields:

$$\frac{1}{s} \cdot V_{\mathrm{P}} = s\,L \cdot I_{\mathrm{CA}}(s) + V_{\mathrm{CA}}(s) \tag{5}$$

$$I_{\mathrm{CA}}(s) = sC_{\mathrm{A}} \cdot \left( V_{\mathrm{CA}}(s) - \frac{1}{s} \cdot v_{\mathrm{CA}}(0) \right) \tag{6}$$

Solving for $I_{\mathrm{CA}}(s)$ and $V_{\mathrm{CA}}(s)$ and inverse transformation yields the time response:

$$i_{\mathrm{CA}}(t) = \frac{V_{\mathrm{P}} - v_{CA}(0)}{Lw} \cdot \sin(wt) \tag{7}$$

$$v_{CA}(t) = V_{\mathrm{P}}\left(1 - \cos(wt)\right) + v_{\mathrm{CA}}(0) \cdot \cos(wt) \tag{8}$$

with $= 1/\sqrt{L\,C_{\mathrm{A}}}$

At $t_{\mathrm{off}}$ the switch $S_1$ is opened and the energy transfer from the primary voltage supply to the resonant circuit stops. Now a freewheel current through the diode $D_2$ flows until $t_{\mathrm{end}}$ , when the diode cuts off and the actuator has the end voltage $v_{\mathrm{CA}}(t_{\mathrm{end}})$. The exact moment when $S_1$ must be opened to reach the desired voltage $v_{\mathrm{CA}}(t_{\mathrm{end}})$ can be calculated using the energy balance equation. Most of the magnetic energy stored in the inductance at $t_{\mathrm{off}}$ will be converted into electric energy at $t_{\mathrm{end}}$ .

$$\frac{1}{2} L\, i_{\mathrm{CA}}^2(t_{\mathrm{off}}) = \frac{1}{2} C_{\mathrm{A}} \cdot \left( u_{\mathrm{CA}}^2(t_{\mathrm{end}}) - u_{\mathrm{CA}}^2(t_{\mathrm{off}}) \right) \tag{9}$$

Equations (7), (8) and (9) allow calculation of the exact time $t_{\mathrm{off}}$ needed to reach every desired voltage $v_{\mathrm{CA}}(t_{\mathrm{end}}) < V_{\mathrm{P}}$ at the transducer. The same calculations can be made for discharging and the result is:

$$t_{\mathrm{off}} = \frac{1}{\phantom{x}} \left| \arcsin\left( \frac{K_1}{K_2} \right) - \frac{\phantom{x}}{2} \right| \tag{10}$$

with $K_1$ and $K_2$ being different for charging and discharging and only dependent on $v_{\mathrm{CA}}(t_{\mathrm{on}})$, $v_{\mathrm{CA}}(t_{\mathrm{end}})$ ($v_{\mathrm{CA}}(t_{\mathrm{end}}) = v_{\mathrm{in}}$ ) and $V_{\mathrm{P}}$. In operation the nonlinear function $t_{\mathrm{off}}(v_{\mathrm{CA}}, v_{\mathrm{in}})$ (Eq. 10) has to be computed at $t_{\mathrm{on}}$ (before every switching cycle). To achieve the desired dynamics this computing has to be done in a few microseconds. This hard time limit would require the fastest and most expensive arithmetic units available at present. For this reason another method was chosen. For all combinations of $v_{\mathrm{CA}}$ and $v_{\mathrm{in}}$ the calculations for $t_{\mathrm{off}}(v_{\mathrm{CA}}, v_{\mathrm{in}})$ were made offline and stored in an EPROM memory. During operation this ROM table is addressed directly by the outputs of two A/D converters sampling $v_{\mathrm{in}}$ and $v_{\mathrm{CA}}$ and the output signal of the memory is the desired time $t_{\mathrm{off}}$. With Eq. (7) the ROM table can be modified so that for every configuration of $v_{\mathrm{in}}$ and $v_{\mathrm{CA}}$ a certain maximum current $i_{\mathrm{CA}}(t_{\mathrm{off}})$ is not exceeded. Thus a current limit can be implemented without current sensing.

## Description of the amplifier as implemented

Fig. 3 shows the block diagram of the amplifier as implemented. The whole control logic of the amplifier is programmed in a single low-cost CPLD (Complex Programmable Logic Device). Every time the current through the inductance is zero, the demagnetization detection starts a new operation cycle. Each operation cycle consists of the following steps: sampling $v_{\mathrm{in}}$ and $v_{\mathrm{CA}}$, converting to digital, addressing the memory directly with the digital values, loading the switching time $t_{\mathrm{off}}$ from the memory into the CPLD, starting the timer with the value $t_{\mathrm{off}}$ and switching the relevent MOSFET for the duration of the time $t_{\mathrm{off}}$.

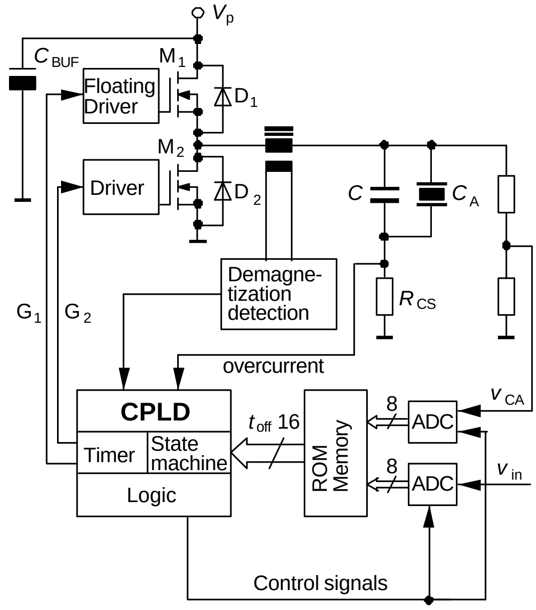


**Fig. 3:** Block diagram of the amplifier as implemented

This leads to a free-oscillating system which adapts its switching frequency to the requirements of the input signal. The overcurrent detection is merely an additional security function and is not necessary for normal operation. The capacitor parallel to the actuator is about twice as big as the highest value of the nonlinear actuator capacitance. Thus the amplifier always has a fixed load capacitance. At the same time this capacitor will reduce the overvoltage at the amplifier output in the case of an external mechanical stimulation. The power losses will not increase dramatically due to this measure because the efficiency of the amplifier is about 95%.

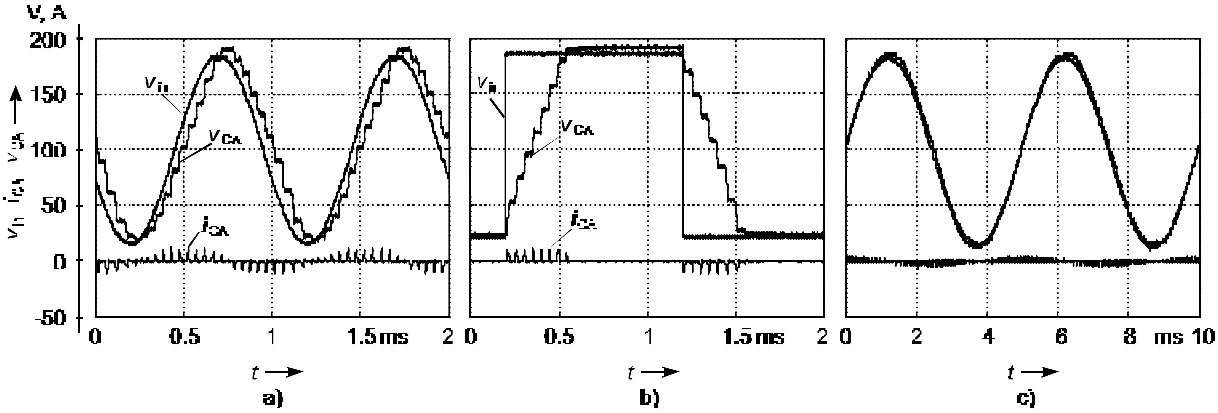


**Fig. 4:** Measured signals of the amplifier as implemented (4µF capacitive load)
a) 1kHz sinus signal b) 500Hz rectangular signal c) 200Hz sinus signal

## System performance

Fig. 4 shows some characteristic transient plots of the amplifier as implemented. Fig. 4 a) demonstrates the working principle and the special features of the amplifier. The output signal $v_{CA}$ always lags only one switching cycle behind the input $v_{in}$. A new switching cycle can only be started if the inductor current $i_{CA}$ is zero. The load capacitance will never be overloaded, so additional switching cycles to correct the output voltage will not occur. For lower-frequency input signals (Fig. 4 c) the output signal accuracy is excellent in comparison to other switching amplifiers. The plots of $v_{in}$ and $v_{CA}$ can hardly be distinguished. It is only near the signal maximum that the curves differ from each other and the switching ripples can be detected. Fig. 4 b) shows the response to a 500 Hz rectangular signal. The cycle-by-cycle current limit (12A), implemented directly in the switching table, causes the voltage to rise in several steps. With the last step the desired output voltage is exactly reached and the amplifier stops switching.
These features would be considerably more difficult to achieve with traditional switching amplifiers using PWM or current-mode control.

## Conclusion

In this paper a new approach to a switching amplifier for piezoelectric actuators has been presented. The amplifier as implemented is a low cost solution, but nevertheless it displays excellent output characteristics. Its high dynamics, its simple construction and its robustness will make this solution interesting for many industrial applications.
For proper system operation the value of the load capacitance used for calculation is not a critical parameter, as long as it is smaller than the exact value.

The tests presented in this paper were only made with capacitive loads. Additional tests have shown that with some changes this approach is also suited for magnetic loads e.g.magnetostrictive actuators.

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