Letter

Multiple-Stage Converter Topology for High-Precision High-Current Pulsed Sources

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Abstract—A new high-current, low-rise-time, and high-precision pulse generator is presented. The topology is based on the use of different stages, each one specific for a particular operation range in terms of power and switching frequency. This approach allows to accomplish current, voltage, and precision requirements with standard semiconductors. Moreover, the proposed topology provides an independent and flexible adjustment of the pulse parameters (rise and fall times, flat-top duration, pulse amplitude, etc.). Experimental results are provided to validate the control of the proposed topology.

Index Terms—Accelerator power supplies, current control, pulse power system.

I. INTRODUCTION

HIGH current pulsed power converters based on a capacitor-discharge topology are the best option for an inductive load that requires fast rise time, short settling time, and extremely precise flat top [1], [2]. This topology is composed by a capacitor bank that is initially charged to be later discharged through a semiconductor switch in the inductive load [3]. The capacitor bank and the load compose a damped resonant circuit. This basic circuit can be further improved with a third harmonic circuit and an active filter to produce high current pulses with a high precision flat top [4]. However, this system presents some disadvantages.

- 1) The current-pulse parameters (flat-top duration, and rise and fall times) depend on the resonant circuit and the input voltage.
- 2) There is a strong dependency on load parameters.
- 3) Reduced efficiency. Power dissipation circuits are often required to discharge the capacitor bank between pulses.

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Fig. 1. Trapezoidal load-current waveform and corresponding applied voltage.

- 4) High rms value of the current pulse. The sinusoidal shape in the load current increases the rms value.

In order to overcome these drawbacks, this study proposes a novel topology for high current high precision pulsed current source applications.

II. MULTIPLE-STAGE CONVERTER TOPOLOGY

In pulsed power applications with an RL load, a trapezoidal waveform is the best solution to reduce the load rms current. Fig. 1 shows the load current and voltage. The current waveform implies the application of a high voltage V_H during rise and fall times, and a low voltage V_{FT} during the flat top.

The necessary input voltage to reach a current flat top I_{REF} in a rise time t_r can be calculated as:

$$V_H = \frac{I_{REF} \cdot L}{t_r} + I_{REF} \cdot R \approx \frac{I_{REF} \cdot L}{t_r} \quad (1)$$

where the load resistance has been considered negligible, due to the low influence of this term in typical pulsed loads. The flat-top voltage is given by the drop voltage on the resistive component of the load ($V_L = I_{REF} \cdot R$).

The development of semiconductor devices with increased current and voltage capabilities offers new possibilities to develop converters with improved performances [5]–[7]. However, the semiconductor technology does not allow to obtain

Fig. 2. Simplified schematic of the proposed topology.

with a same component the high currents (few kiloamperes), the high voltages (several kilovolts) and the high switching frequency necessary to achieve the high precision required. A two-stage voltage converter based on a high-voltage stage for the rise and fall times and a low voltage stage to control the flat-top current is proposed in [8]. This approach allows to reach the flat-top current in a shorter time and to decrease the switching frequency at flat top compared to a single-stage converter. However, since the flat-top structure must handle the total output current, the precision is bounded by the maximum switching frequency of its high-current semiconductors.

To overcome these problems a multiple stage current converter based on three structures with different voltage and current ranges that connect sequentially during the pulse-generation stages is proposed (see Fig. 2). The aim is to use structure 1 (high voltage and current) during rise and fall times, structure 2 (low voltage and high current) to control the flat-top mean current with moderate precision, and structure 3 (low voltage and current) to control the load current with the required precision.

The operational principle of this topology is summarized as follows.

Rise time: During this stage, the only structure that remains active is structure 1, which initiates the charge of L_1 and L through the high voltage source $V_{C1} = I_{REF} \cdot (L + L_1)/t_r$. Therefore, switches S_1 and S_2 are turned ON and S_3 and S_4 are turned OFF. If the load resistance is negligible, $i_1 = i_L$ increase with a slope given by $V_{C1}/(L_1 + L)$.

Flat top: When currents $i_L = i_1$ reach the reference value I_{REF} , structure 1 is disconnected and structures 2 and 3 are connected by means of S_1 turning-off and S_3 turning-on. The voltages of these structures, V_{C2} and V_{C3} , must be higher than $I_{REF} \cdot R$. The turning-on of S_3 allows the connection of the active filter and the capacitor C to the node (A). C is connected in parallel to structure 3 to avoid overvoltage in this node.

Since structure 2 must handle the high load current, a medium S_4 switching frequency, f_{I1} , must be adopted. As a consequence, a peak-to-peak ripple current ΔI_1 , much higher than

TABLE I
SEMICONDUCTOR DEVICES REQUIREMENTS

Devices	Voltage	Current	Operation frequency
S_1, S_2, D_1, D_2	High (V_{C1})	High (I_{REF})	Very low (f_P)
S_3, D_5	High (V_{C1})	Low (ΔI_1)	Very low (f_P)
S_4, D_4	Low (V_{C2})	High (I_{REF})	Medium (f_{I1})
$H_1 \dots H_4$	Low (V_{C3})	Low ($\sim \Delta I_1$)	High (f_{IF})

Fig. 3. Current waveforms and operation principle.

the required one, is obtained. Assuming that the forward voltage of the devices is negligible, the expression of ΔI_1 is given by

$$\Delta I_1 = \frac{v_C}{L_1 \cdot f_{I1}} \cdot \left(1 - \frac{v_C}{V_{C2}}\right). \quad (2)$$

In order to cancel ΔI_1 , structure 3 must operate at a switching frequency f_{IF} higher than f_{I1} . This structure generates a ripple current ΔI_f as follows:

$$\Delta I_f = \frac{V_{C3}}{2 \cdot L_f \cdot f_{IF}} \cdot \left(1 - \frac{v_C^2}{V_{C3}^2}\right). \quad (3)$$

Fall time: To decrease the load current, all switches are turned OFF. The energy stored in the load and inductor L_1 returns to the capacitor bank C_1 through D_1 , D_2 , and D_4 . The current difference between i_{L1} and i_L when S_3 is turned OFF flows through D_5 .

Table I resumes the requirements for the semiconductor devices, where f_P is the pulse repetition frequency.

Fig. 3 shows the current waveforms of the proposed system and the state of the different semiconductor switches.

III. CONTROL SYSTEM

As the system is composed by different interconnected structures, a transient is generated each time a structure is

Fig. 4. Flat-top equivalent circuit.

connected/disconnected. This aspect is critical at the beginning of the flat top, as the generated transient can exceed the system specifications of duration and amplitude. Fig. 4 shows an equivalent circuit of the system in the flat top, where G_{I1} and G_{IF} current generators model i_1 and i_F , respectively.

The capacitor and the load form a second-order circuit with a low damping factor. If the initial conditions of the circuit are not appropriate ($i_1 = i_L \neq I_{REF}$ or $v_C \neq I_{REF} \cdot R$), an undesired oscillatory response will be generated. In practice, the appropriate conditions are very difficult to obtain due to possible delays in the devices operation (i.e., S_1) and measurement errors. Therefore, the control system must be able to handle the transient response. In order to achieve this, a state-feedback control is proposed to place the closed-loop poles of the second-order circuit at ω_P , so as to obtain a critical damped response. The feedback gains K_V and K_I are given by

$$K_V = \frac{C}{K_f} (2 \cdot \omega_P - R/L)$$

$$K_I = \frac{1}{K_f} \left(\frac{\omega_P^2}{\omega_0^2} - K_f \cdot K_V \cdot R - 1 \right) \quad (4)$$

where ω_0 is the resonant frequency of the RLC circuit and K_f is the active filter gain. This gain is assumed constant as the bandwidth of the active filter is much larger than the bandwidth of the system.

As the resulting transfer is a type-0 system, there will be a steady-state error in the mean value of i_L , which is compensated by an external loop with an integral-type controller. The gain of this controller is adjusted to obtain a cutoff frequency of ω_C . In order to obtain a transient response mainly dependant on the external loop bandwidth, the poles of the state feedback are placed above the bandwidth ($\omega_P > \omega_C$).

The precision of the flat-top current depends mainly on the rejection of the system to current ripples ΔI_f and ΔI_1 . Regarding ΔI_f , it can only be attenuated by the RLC circuit transfer function (5), since the feedback loops do not reject it

$$\frac{\Delta I_L}{\Delta I_f} = \frac{\omega_0^2}{s^2 + 2 \cdot \xi \cdot \omega_0 \cdot s + \omega_0^2} \approx \frac{\omega_0^2}{\omega_{IF}^2} \quad (5)$$

TABLE II
SEMICONDUCTOR DEVICES REQUIREMENTS

Devices	Voltage	Current	Operation frequency
S_1, S_2, D_1, D_2	$V_{C1} = 3.3 \text{ kV}$	$I_{REF} = 2 \text{ kA}$	$f_P = 1 \text{ Hz}$
S_3, D_5	$V_{C1} = 3.3 \text{ kV}$	$\Delta I_1 = 48 \text{ A}$	$f_P = 1 \text{ Hz}$
S_4, D_4	$V_{C2} = 500 \text{ V}$	$I_{REF} = 2 \text{ kA}$	$f_{I1} = 5 \text{ kHz}$
$H_1 \dots H_4$	$V_{C3} = 500 \text{ V}$	$\sim \Delta I_1 = 48 \text{ A}$	$f_{IF} = 100 \text{ kHz}$

ΔI_1 rejection depends on feedback loops and is given by

$$\frac{\Delta I_L}{\Delta I_1} = \frac{1}{K_f} \cdot \frac{K / ((s/\omega_P) + 1)^2}{1 + (\omega_C/s(s/\omega_P + 1)^2)} \quad (6)$$

where $K = K_f / (1 + K_V \cdot K_f \cdot R + K_I \cdot K_f)$ is the gain of the transfer function resulting from the state feedback of the RLC circuit. The worst-case ripple rejection occurs when $\omega_C < \omega_{I1} < \omega_P$, then ΔI_1 is attenuated as

$$\frac{\Delta I_L}{\Delta I_1} \approx \frac{K}{K_f} = \frac{1}{1 + K_V \cdot K_f \cdot R + K_I \cdot K_f} \quad (7)$$

Equation (7) shows that the ripple rejection on the flat top depends inversely on R . Hence, as the increase of the heating power dissipation tends to increase the load resistance, ΔI_L is reduced and the rejection is improved. Operating with (4) and (7) yields a simpler expression

$$\frac{\Delta I_L}{\Delta I_1} \approx \frac{\omega_0^2}{\omega_P^2} \quad (8)$$

Equation (8) shows that a better rejection requires a larger ω_P , i.e., a larger active filter frequency. In order to avoid increasing the active filter frequency, a feedforward loop is incorporated. This compensation is obtained by adding $i_1 - I_{REF}$ to the external loop reference. In this case, ΔI_1 in (8) is replaced by the residual ripple of the feedforward compensation, which allows to obtain a higher rejection.

IV. POWER SYSTEM

To evaluate aspects related to the power implementation, a pulsed source with the following specifications was considered: $I_{REF} = 2 \text{ kA}$, $L = 1 \text{ mH}$, $R = 0.15 \Omega$, $t_r = t_f = 1 \text{ ms}$, $f_P = 1 \text{ Hz}$, and flat-top precision of 5×10^{-4} . Using (1)–(3), the requirements given in Table II were obtained.

f_{I1} and f_{IF} were adopted to obtain a low stress on the semiconductors S_4 , D_4 , and $H_1 \dots H_4$, which is defined by the required lifetime as regards thermal cycling capability. Concerning S_1 , S_2 , S_3 , D_1 , D_2 , and D_5 , their stress will depend on f_P , which is very low.

The delays of S_4 and $H_1 \dots H_4$ generate an increase of ΔI_1 and ΔI_f ripples. Hence, they must be considered when the ripple rejection is adjusted by (7) and (8).

The reverse-recovery time of D_4 produces a reverse current proportional to the diode forward current and its slope during turn off, which flows through S_4 in addition to I_1 . Moreover, during this recovery time, both S_4 and D_4 are forward biased simultaneously. Hence, the S_4 current is only limited by stray inductances. Since i_{D4} forward current and its slope are high, the maximum value of the reverse current can damage S_4 ; therefore,

Fig. 5. Load current with capacitor-discharge converter.

it should be taken into account for S_4 ratings. Additionally, stray inductances could generate an overvoltage across the diode that can damage it. This effect can be solved by using a low reverse-recovery time diode with a conventional snubber circuit. Although this phenomenon impacts on the converter efficiency, it does not affect the precision of the flat-top current.

In the flat top, the most influential parasitic component is the capacitor C equivalent series resistance, ESR, which modifies the poles and adds a zero at $\omega_z = 1/(\text{ESR} \cdot C)$ of the RLC function transfer. The poles modification is not a drawback, since they can be relocated by means of the feedback gains adjustment (K_V and K_I). On the other hand, the zero could decrease the ΔI_f ripple rejection calculated in (5), which affects the flat-top precision. To avoid this, the zero location must be $\omega_z > \omega_{IF}$, which can be achieved by adopting a capacitor with an appropriate ESR or through a parallel array of capacitors.

Concerning L_1 and L_f , their possible variations (e.g., by hysteresis or by saturation) will change either the frequencies f_{I1} and f_{IF} (if hysteresis current control is used) or ΔI_1 and ΔI_f (if fixed frequency current control is used). In both cases, the ripple rejection will be modified. Therefore, the lowest inductance values of L_1 and L_f should be considered either to adjust the hysteresis bands or the ripple rejection by means of (5) and (8). Another characteristic of L_1 and L_f is their series resistances, which are low. In general, as the switching times are much lower than the associated time constants, the current waveforms are practically linear.

V. SIMULATIONS

In order to compare the proposed topology with the capacitive discharge one, both converters have been evaluated using MATLAB/Simulink. The simulation integrates the different effects of digital control, like communication and calculation delays or A/D conversions. The load and current pulse characteristics are the same as the ones in Section IV, considering a t_{ft} of 1–3 ms.

Fig. 5 shows a current pulse obtained with the capacitor-discharge converter for a $t_{ft} = 1$ ms. The resonant frequency and the input voltage were adjusted to 133 Hz and 2914 V, respectively. It can be noted that t_r and t_f are close to 1.4 ms.

If a flat top of 3 ms is considered, the resonant frequency and charging voltage should be modified to 44 Hz and 980 V. The

Fig. 6. Load current (solid line: capacitor-discharge converter, dashed line: proposed topology).

Fig. 7. Detail of the load current during flat top.

most critical-parameter modification is the resonance frequency as it involves modifying the circuit elements. Moreover, the decrease in the resonance frequency leads to higher rise and fall times, which increases the associated current effective value [see Fig. 6 (solid line)].

In the case of the proposed topology, the t_r and t_{ft} adjustments are independent and are defined by V_{C1} (1) and C_2 , respectively. C_2 is sized in order to be able to deliver the required energy to the load (which is defined by t_{ft} and I_{REF}), with a voltage drop that does not affect the i_1 control, i.e., $V_{C2min} > v_C$. In this case, C_2 is a 60 mF/500 V capacitor bank. Fig. 6 (dashed line) shows a current pulse of $t_{ft} = 3$ ms using the proposed topology. It can be seen that, t_r and t_f are significantly lower, with a 30% reduction of the effective value.

Fig. 7 shows a zoom of the load current during flat top. The transitory response at the beginning of flat top is adjusted to 200 μ s by taking $\omega_C = 2\pi 2.7$ kHz. The bandwidth and precision of the sensors have been assumed adequate for a high-precision measurement, which allows to obtain a precision close to 200 ppm. Generally, the precision will depend on both the sensors and the semiconductors switching frequencies.

TABLE III
PROTOTYPE PARAMETERS

L	1 mH	$V_{C2} = V_{C3}$	10 V
R	0.15 Ω	I_{REF}	15 A
L_1	500 μ H	t_r, t_f	1 ms
L_F	50 μ H	Settling time	< 200 μ s
V_{C1}	25 V	Flat-top duration	3 ms

VI. EXPERIMENTAL RESULTS

Experimental tests were carried out on a low-power-scale prototype. The aim of these tests was to validate the controlled interconnection of stages and the associated control of the proposed topology, without focusing on the resulting precision. The current and voltage values of this prototype have been scaled down to laboratory levels, while the switching frequencies have been kept to maintain the system dynamics. The main parameters of the system are listed in Table III.

The current control of i_1 and i_F is performed by means of hysteresis current control. The hysteresis bands of i_1 and i_F are adjusted to obtain a maximum switching frequency of 5 and 100 kHz, respectively. In order to fulfil the settling time requirement, the external loop bandwidth was adjusted to 2.7 kHz, while the closed-loop poles of the state feedback were located in 15 kHz so as not to affect the stability of the external loop or to influence the transient response.

The proposed control system was implemented by a specific electronic board composed of a Xilinx FPGA XCS40XL and a 16-bit analog/digital acquisition module. The main block of the control system performs the following tasks:

- 1) structure interconnection management;
- 2) i_1 and i_F digital hysteresis current controls,
- 3) control loop algorithms,
- 4) management of control and communication with the analog/digital acquisition module.

The acquisition module consists of four 16-bit ADC with their corresponding differential input filters, while the sampling frequency is 1 MHz. The measurement of i_1 and i_L was performed with Hall effect current sensors (accuracy $\pm 0.5\%$) and the measurement of i_F through a Pearson-type current sensor (accuracy $\pm 1\%$).

A current pulse with no significant transient response during the coupling of structures (beginning and end of the flat top) was obtained; times t_r , t_f , and t_{ft} were within specifications (see Fig. 8). t_r was greater than t_f because, during the rise time, the forward voltage of the devices and V_{C1} have an opposite polarity, while during the fall time they have the same polarity. In the full-scale prototype, this effect will be negligible since V_{C1} is much greater than the voltage drop of the devices.

The transitory response at flat top beginning is shown in Fig. 9, where its duration is 184 μ s and its amplitude is 0.7% of I_{REF} . The measured precision in the flat top is close to 500 ppm.

VII. CONCLUSION

A new multiple-stage converter topology for pulsed current sources was proposed. The proposed topology allows to fulfil

Fig. 8. Experimental results of the scaled-down prototype. Current pulse measurement (2.5 A/division, 1 ms/division).

Fig. 9. Experimental results of the scaled-down prototype. Flat-top transient response (20 mA/division, 200 μ s/division).

the requirements of flat top and rise/fall times independently. This feature provides flexibility in the adjustment of the pulse parameters, which allows the reduction of the pulse current effective value. Furthermore, the proposed system allows the recovery of the load energy during the fall time.

The aforementioned characteristics make the proposed topology suitable for high current pulses with high precision flat top and short rise and fall times applications. Besides, it allows to extend the converter operation boundaries parallel to the semiconductor technology improvements.

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