A field programmable gate array controlled high-current pulsed source for airborne transient electromagnetic system

Wei Liu | Xuquan Hu | Xian Liao | Longhuan Liu | Zhihong Fu

1 Department of Electrical Engineering Theory and New Technology, School of Electrical Engineering, Chongqing University, Chongqing, People’s Republic of China
2 Chongqing Triloop Prospecting Technology Co., Ltd., Chongqing, People’s Republic of China

Correspondence
Xuquan Hu, Sixth Teaching Building, District A, Chongqing University, No.174 Shazhengjie, Shapingba, Chongqing 400044, China.
Email: huxuquan@163.com

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Abstract
A high-current pulsed source with high charging efficiency, high stability and low current ripple is proposed for airborne transient electromagnetic method application. The high voltage capacitor bank is employed to provide and store energy for generating current pulses in the inductive load. An N-channel interleaved boost module is designed to replenish the capacitor energy gap so that current pulses can be repeated in a short time. The operation principles of the proposed topology are explained and the theoretical analysis is provided. In addition, a convenient and effective control strategy to raise efficiency of the charging circuit is introduced. The concise closed-loop control method directed at the characteristics of load variations is available for changing the steady operation and control the output current. Logic input and output modules based on field programmable gate array are designed for the controller. A laboratory prototype with the maximum peak current 400 A is built for the load coil of 2.1 mH and 150 mΩ to verify the effectiveness of the proposed topology and control method.

1 | INTRODUCTION

Airborne transient electromagnetic method (ATEM) has become one of the most popular ways for geophysical prospecting due to its rapidity and high efficiency [1, 2]. As shown in Figure 1, the transmitter on the helicopter generates a bipolar pulsed current in the exciting coil, which produces the primary magnetic field in space. As a result, the eddy current under the ground generates a second magnetic field based on Ohm effect, which could cause changes in the voltage between the two terminals of receiving coil. The target body (such as mineral ore, saline-alkali soil, underground water etc.) could be found by analysing the information of the second magnetic field [3, 4].

In order to achieve a greater detection depth, the transmitting system is required to provide pulse current with higher peak value. In addition, the current pulse should be turned off quickly to acquire a higher detection resolution [5]. However, these key technical issues are not easy to solve for the following two reasons. First, the transmitting system gets power from the helicopter whose maximum power supply is generally no more than 3 kW. The limitation of input power makes it difficult to generate large current pulse. Second, the high inductive characteristic of transmitting coil restricts the rate of current change and thus causes long turn-off time of transmitting current [6].

The high-current pulsed source has been extensively employed in many fields such as cancer therapy [7], particle accelerator [8], synchrotron beam chopper system [9] etc. However, the power supply required by these current sources was quite different from that of ATEM. The multi-structure power converters achieving high-current high-precision pulsed current source are extremely attractive [10, 11]. The main problem of them is overly complex control system and large volume. The topology utilizing series connected full-bridge submodules (FB-SM) [12] and switched-capacitor (SC) units [13] could generate high-current pulses on the pure resistive load but it is not suitable for high inductive load. A current source based on positive buck-boost converters concept was presented in [14] whereas it could not operate with a suitable peak power which applied to ATEM.

In order to charge or discharge the inductive load frequently, high voltage capacitor bank seems to be a good choice for charging and accumulating energy. The pulse current source in [15] utilizes constant voltage-clamping strategy to enhance the rising speed of the pulse current without energy supplement for
the capacitor. An electromagnetic transmitter with a maximum transmitting current of 20 A was proposed in [16]. A boost element with single inductor is adopted to improve edge steepness of load current. However, the charging efficiency of this topology is insufficient to increase the peak current of power inverter. Besides, high ripple of input current leads to severe electromagnetic interference (EMI) which has a negative influence on stability of entire system. Converters with interleaved operation can provide higher output power and lower input current ripple [17–20]. Recently, the boundary-conduction-mode (BCM) boost converter with interleaving more channels has been a popular topology due to its higher output power levels [21, 22]. In general, few reports are available for capacitive energy storage based on interleaved operation.

In this paper, a high-current pulsed source for particular inductive load applied to ATEM is proposed. The operation principles of the two parts are described in detail respectively. A digital controller based on field programmable gate array (FPGA) is designed to improve charging efficiency for capacitor bank and allow it to adjust the steady state of the current source freely. The closed-loop control method directed at the characteristics of load variations can easily achieve constant peak value of current pulses and improve the accuracy of detection results for ATEM system. Simulation and experimental results validate the capability of the proposed topology and practicality of the designed controller.

2 PROPOSED TOPOLOGY AND OPERATING PRINCIPLE

The schematic diagram of the proposed topology is shown in Figure 2 in which two typical circuit structures are connected by high voltage capacitor bank $C_o$. The coil model is simplified as a series combination which consists of an inductor $L_o$ of 2 mH or higher and a resistance $R_o$ of about 150 m$\Omega$. An H-bridge IGBT inverter is included to generate bipolar high-current pulses in the coil load. The energy required for this process decreases continuously for its provider, passive element $C_o$. To replenish the energy gap during per transmitting operation, an N-channel interleaved boost module is designed to raise the capacitor voltage when H-bridge inverter enters the sleep state. It is composed of N boost submodules which charge the capacitor in turn. In fact, multi-inductor charging circuit works in the transmitting dead zone to avoid degradation of the current pulse waveform quality caused by the switching operation of MOSFET devices, which may further affect the magnetic field and detection results.

The key waveform of the current pulsed source with steady-state operation can be observed in Figure 3. The operation principles of the positive half cycle is illustrated in detail hereinafter. To simplify the operating principles analysis, all semiconductor devices are considered ideal.

Mode 1 [$t_0$, $t_1$]: At $t_0$, $S_1$ and $S_4$ are turned-on so that $C_o$, $R_o$ and $L_o$ are connected into a loop. Part of stored energy in $C_o$ is transferred to inductive load. The output current $i_o$ rises rapidly to $I_p$ and flows in the direction shown by the solid red line in Figure 4. The capacitor voltage $u_o$ decreases from $U_1$ to $U_2$. Following equation is maintained for this interval:

$$L_o C_o \frac{d^2 i_o(t)}{dt^2} + R_o C_o \frac{di_o(t)}{dt} + i_o(t) = 0$$

(1)

The initial conditions of the second order differential equation are as follows:

$$\begin{cases} i(t_0) = 0 \\ \left. \frac{di_o(t)}{dt} \right|_{t = t_0} = U_1 \end{cases}$$

(2)

In consideration of the actual situation, the main parameters of the circuit meet the following single inequality:

$$(R_o C_o)^2 - 4 L_o C_o < 0$$

(3)
By rearranging Equation (1), the output current \(i_o(t)\) can be calculated by:

\[
i_o(t) = \frac{U_1}{\mu L_o} e^{-\delta (t-t_0)} \sin[\mu(t-t_0)]
\]  

where

\[
\mu = \sqrt{\frac{4L_oC_o - (R_oC_o)^2}{2L_oC_o}}
\]

\[
\delta = \frac{R_o}{2L_o}
\]  

To obtain a higher peak value of current pulse in a short time, the fundamental period of transmitting current pulse \(T_o\) and output duty cycle \(D\) satisfy the following relation

\[
DT_o < \frac{\pi}{\mu}
\]

Substituting the time of rising edge \(t_1 = t_1 - t_0 = \frac{DT_o}{2}\) into Equation (4), the peak current \(I_p\) is obtained

\[
I_p = \frac{U_1}{\mu L_o} e^{-\frac{\delta DT_o}{2}} \sin\left(\frac{\mu DT_o}{2}\right)
\]  

The reduced capacitance voltage \(U_2\) can be calculated (see Equation (8)).

\[
U_2 = L_o \left. \frac{di_o(t)}{dt} \right|_{t=t_1} + I_pR_o = U_1 e^{-\frac{\delta DT_o}{2}} \left[ \cos\left(\frac{\mu DT_o}{2}\right) \right. \\
- \frac{\delta}{\mu} \sin\left(\frac{\mu DT_o}{2}\right) + \frac{R_o}{\mu L_o} \sin\left(\frac{\mu DT_o}{2}\right)
\]

Mode 2 \([t_1, t_2]\): At \(t_1\), \(S_3\) and \(S_4\) are turned-off while the freewheeling diodes \(D_{S2}\) and \(D_{S3}\) are forward-biased. The output current \(i_o\) decreases in the direction shown by the dotted red line in Figure 4. By neglecting voltage drop of the diodes, the initial conditions are as follows

\[
\begin{cases}
i_o(t_1) = I_p \\
i_o(t_2) = 0
\end{cases}
\]  

Substituting Equation (9) into Equation (1), the output current \(i_o(t)\) \((t_1 \leq t \leq t_2)\) can be obtained

\[
i_o(t) = \frac{U_1}{\mu L_o} \sin\left[\frac{\mu DT_o}{2}\right] - e^{-\delta (t-t_0)} \sin[\mu(t-t_0)]
\]

Applying Kirchhoff’s voltage law (KVL) at the beginning of this interval provides the following equation

\[
-L_o \left. \frac{di_o(t)}{dt} \right|_{t=t_1} = U_2 + I_pR_o
\]

By replacing \(U_2\) and \(i_o(t)\) in Equation (11) from Equations (8) and (10), the time of falling edge \(t_f\) is given by

\[
t_f = t_2 - t_1 = \frac{1}{\mu} \arctan\left(\frac{\mu L_o \tau}{\mu L_o - 2\delta L_o \tau + 2R_o \tau}\right)
\]
where
\[ \tau = \tan \left( \frac{\mu DT}{2} \right) \] (13)

At \( t_2 \), when the current pulse decreases to zero, the capacitance voltage rallies to \( U_3 \)
\[ U_3 = -I_{t_1} \left. \frac{dU_o}{dt} \right|_{t=t_2} \Delta t = U_1 \sin \left( \frac{\mu DT}{2} \right) \left( e^{\delta \left( t + \frac{D}{2} \right)} \right) \] (14)

Above analysis illustrates the changing processes of capacitance voltage at the rising edge and falling edge of current pulse. Part of stored energy in capacitor is consumed by the load during these two intervals and the rest is fed back to \( C_o \).

Mode 3 \([t_3, t_4]\): Since \( t_3 \), the H-bridge inverter stops running. To repeat the transmitting pulse and guarantee the same peak current of next pulse, a charging strategy should be formulated to raise the capacitance voltage from \( U_3 \) to \( U_1 \). The energy supplied to the capacitor in the interval \([t_3, t_4]\) can be drawn as:
\[ \Delta E = \frac{1}{2} C_o U_1^2 \left[ 1 - \sin^2 \left( \frac{\mu DT}{2} \right) e^{-\delta \left( t + \frac{D}{2} \right)} \right] \] (15)

Obviously the increasing variation \( \Delta E \) has a positive influence on the initial voltage \( U_1 \) and the peak current \( I_p \).

An independent time period \([t_3, t_4]\) is intercepted from the interval \([t_2, t_4]\) to charge the capacitor so that the process of energy replenishment does not disturb the output current pulse. The \( E_{NC_1} \) in Figure 3 represents enable signal of the charger and N-channel interleaved boost module starts its energy supplying from \( t_3 \). In order to ensure the correct interleaving operation, a phase shift of \( \frac{n-1}{N} \times 2\pi \) is maintained between MOSFET gate signals of each channel. Generally, the switching frequency of \( Q_1 \) to \( Q_N \) is much faster than \( V_1 \) to \( V_4 \) so that the volume of each energy-storage inductor can be reduced.

For the convenience of controller design, the conduction time in one switch cycle of all submodules are equal:
\[ t_{on1} = t_{on2} = \cdots = t_{onN} = t_4 - t_3 = t_{on} \] (16)

Assuming that there is no difference in the main parameters of boost elements for each submodule and the internal resistance of \( L_1 \) to \( L_N \) is negligible, the boost inductor current \( I_L \) can be obtained
\[ I_L = \frac{U_S}{L_{on}} \] (17)

where \( L_1 = L_2 = \cdots = L_N = L \).

After this all the energy stored on the inductors is transferred to the capacitor \( C_o \). Important equation during this interval is as follows:
\[ P_c = \frac{1}{2} C_o U_1^2 - \frac{1}{2} C_o U_3^2 \] (18)

where \( P_c \) is the average charging power of interleaved boost module in the charging time period \( t_c \). Let the duty cycle of each submodule be same
\[ d_1 = d_2 = d_3 = \cdots = d_N = d \] (19)

An energy equation of the whole system can be established without considering the switching loss:
\[ \Delta E = \frac{N d k}{t_{on}} \times \left( 0.5 L I_o^2 \right) \] (20)

By replacing \( \Delta E \) and \( I_o \) in Equation (20) from Equations (15) and (17) above, the relationship between output peak current \( I_p \) and input voltage \( U_S \) of the power source can be obtained:
\[ I_p = \sqrt{\frac{N d k t_{on}}{L C_o (\delta DT - e^{-\delta DT})}} \frac{\sin \left( \frac{\mu DT}{2} \right) U_S}{\mu L} \] (21)

After the completion of charging, the capacitor voltage \( \mu_o \) returns to \( U_1 \) at \( t_5 \). The operation principles of the negative half cycle for the output current is similar to the positive one so that the relevant statements are omitted.

3 | DESIGN CONSIDERATION

According to the above analysis, the energy-storage capacitor bank and charging circuit with N-channel interleaved boost module play a crucial role in generating high-current pulses for the inductive load. The parameters used include the transmitting periodic time \( T_o = 40 \text{ ms} \) and output duty cycle \( D = 0.08 \), that is, the rise time of current pulse \( t_3 = \frac{DT}{2} = 1.6 \text{ ms} \). The selection of corresponding passive elements and parameter design of boost submodule are discussed in this section. Then the circuit’s dynamic performance is researched.

3.1 | Energy-storage capacitor bank

The high-voltage capacitor bank serves as an independent energy-supply device for the H-bridge inverter during the rise time and fall time of current pulses. The volume and weight of \( C_o \) depend on its maximum pressure endurance and the capacitance. Furthermore, based on the requirements of geophysical prospecting, a current pulse with higher peak value and shorter fall time is more suitable for the transmitting system. So the effect of different capacitance on the output current pulse and charging circuit is investigated.

Figure 5 depicts the output peak current and fall time of current pulses when the capacitance varies from 0.5 to 4 mF without considering the charging circuit of N-channel interleaved boost module. It can be seen from Figure 5 that, the peak current rises as the capacitance is increased whereas the fall time keeps a sustained downward trend. For the capacitance larger
than 2 mF, both the peak current and fall time change slowly with increasing the capacitance. A capacitor of about 1.5 mF can achieve current pulses with peak value of more than 400 A and fall time of less than 1.5 ms. Larger capacitors do little to raise the peak current and reduce the fall time and may increase the size of system structure.

The charging power that the transmitting circuit needs is given by Equation (18). Figure 6 shows the relationship between the charging power $P_c$, the initial voltage of capacitor $U_1$ and the capacitance $C_o$ when the load stays constant. It is observed that the charging power increases when the initial voltage or the capacitance is risen. For a front power supply that can provide a maximum charging power of 2–3 kW, an initial voltage of 700–900 V and an energy-storage capacitor of 1.5–3 mF may be the best choice to output high-current pulses. It helps to reduce the weight and volume of the system and take well advantage of the power provided by the helicopter generator.

### 3.2 Parameter design of boost submodule

As described earlier, the N-channel interleaved boost module charges the capacitor intermittently and it is important to ensure that the module operates stably. It is noticed that each boost submodule must not enter continuous conduction mode (CCM) which may cause damage to the devices. Discontinuous conduction mode (DCM) or boundary conduction mode (BCM) can ensure that all the stored energy of inductors ($L_1$ to $L_N$) is transferred to capacitor in each switching cycle.

#### For a boost inductor operating in steady state, the following formula can be derived from volt-second equilibrium principle

$$U_S t_{on} = (\mu_o - U_S) t_{off}$$  \hspace{1cm} (22)

where $t_{off}$ is the fall time of inductor current from the peak value $I_L$ to 0. Given the power limitation, the following single inequality should be satisfied:

$$\frac{0.5 N I_L U_S (t_{on} + t_{off})}{T_c} \leq P_{c_{max}}$$  \hspace{1cm} (23)

where $T_c$ is the time interval between two turn-on times of MOSFET. When the boost submodule works in BCM mode, $T_c$ has its minimum value $t_{on} + t_{off}$.

Substituting Equation (17) into Equation (23), the inductance of $L$ can be obtained:

$$L \geq \frac{0.5 N t_{on} U_S^2}{P_{c_{max}}}$$  \hspace{1cm} (24)

For MOSFET in each boost submodule, the voltage stress is $U_{1_n}$ and the maximum current flowing through $Q_n$ is $I_n$. Considering adequate safety allowance, the design is 1.5 times of the maximum value in the actual scheme.

### 3.3 Circuit's dynamic performance for load variation

To clearly illustrate the circuit characteristics, the series RL circuit is used to simulate the real load. However, the load parameters are not completely unchanged during the system operation. As we know, the resistance of the coil will increase with the rising of its own temperature, while its inductance is nearly a fixed value. On the other hand, in order to achieve high accuracy of the detection results, it is necessary to ensure that the peak value of the current pulses does not change with the load variation, so as to generate magnetic fields with the same strength.

When constant-frequency control method is adopted for the charging circuit, Figure 7 shows the relationship between the
duty cycle of boost submodules $d$ and the coil resistance $R_c$ when the peak value of current pulses remains constant.

As coil resistance increases, the duty cycle of boost modules needs to be raised to maintain the stability of peak current. Considering the ambient operating temperature outside the helicopter and metal thermal balance, coil temperature will not exceed a certain upper limit. The green areas in Figure 7 is the actual variation range of the coil resistance. It can be seen that the peak value of the current pulse can be kept stable by changing the duty cycle only in a small range. For current pulses with a peak value of less than 400 A, the adjustment range of duty cycle, $\Delta d$ is generally not more than 0.15.

4 | SYSTEM CONTROL SCHEME

In order to analyse the controller design of the current pulsed source, a transmitting system with four-channel interleaved boost module is discussed as an example.

4.1 | Control strategy of charging circuit

To avoid the damage of MOSFET caused by the continuous increase of charging current, it is necessary to establish a soft-start mechanism during the initiation of the current source. When the conduction time $t_{on}$ reaches its default value after the soft-start process, there are two methods to maintain steady-state charging. Constant-frequency control (CFC) method simply sets fixed turn-on and turn-off times for each MOSFET and no additional analogue or digital circuitry is needed.

As shown in Figure 8(a), when all boost submodules work with the fixed switching frequency $f_s$, it is observed that the inductor current of every single channel would drop to zero before the MOSFET is turned on again. For every switching cycle, there are non-charging time period of $\frac{1}{f_s} - t_{on} - t_{off}$ which leads to efficiency degradation. Increasing the frequency $f_s$ would help solve the problem but it is difficult to determine the maximum $f_s$ both efficiently and safely.

Constant-on-time control (COTC) changes the switching frequency while maintaining the conduction time $t_{on}$. It is widely used in the design of DC-DC converter due to its high light-load efficiency. Figure 8(b) depicts total input current and branch current of each single channel when the charging circuit works in COTC mode. Comparing the two control methods, it can be found that COTC mode increases the charging current by about 10 A without changing the current ripple.

Traditional open-loop controller in COTC mode for N-channel interleaved boost converter needs N zero-current-detection (ZCD) circuits to ensure that each switch is turned on at the zero-inductor-current time and this makes analogue peripheral circuitry much more complex. Considering the charging process and variation trend of capacitor voltage $u_c$, a simplified controller is designed to raise the charging efficiency.

Figure 9 shows the changes of $u_c$ as the charging currents of four phases are injected to capacitor bank in a prearranged sequence. The $u_{o1}$, $u_{o2}$, $u_{o3}$, $u_{o4}$, which are the voltages when the inductor currents $i_{L1}$, $i_{L2}$, $i_{L3}$, $i_{L4}$ start to shut off, are lifted step-by-step during each charging cycle. The analysis results in Section 3 manifest that the actual turn-off time of each inductor current is decreasing as the capacitor voltage is raised, that is, $t_{off1} > t_{off2} > t_{off3} > t_{off4}$. Therefore, as long as the inductor current of the first phase is guaranteed to work in BCM mode and the same delay time $1/N(t_{on} + t_{off1})$ is given between every two phases, the other submodules can theoretically work in light DCM mode. Since the inductor current turn-off time of different phases is pretty close, the non-charging time of each channel is negligible.

Based on the method mentioned above, auxiliary winding of $L_1$ and a ZCD circuit is needed to trap the zero-crossing point
4.2 Voltage control method

As described in Section 3, with the fixed parameters of H-bridge inverter and load coil, the output peak current $I_p$ is mainly related to the initial voltage of the capacitor and further determined by supplemental energy of the charging circuit.

In fact, setting different reference voltage $U_{ref}$ for charging stage leads to different steady state and output peak current. Since the relationship between capacitor voltage and peak current has been shown in Section 2, the reference voltages could be classified into several levels according to the required output peak currents.

As shown in Figure 11, a hysteresis comparator is used to track the moment when the capacitor voltage reaches the reference value during charging time $t_c$. This control mode enables the peak current to be adjusted in a wide range. What is more, its comparison bands could be also adjusted to achieve the required precision.

The charging module stops running for the rest time $t_{c_{max}} - t_c$ until next current pulse is generated. The operational waveforms of the voltage control method are shown in Figure 12. Compared with the simple charging time control method, voltage control can avoid the problem of inconsistencies in periodic recharge caused by slight difference between the inductance of different channels.

4.3 Closed-loop control for constant peak current

The circuit’s dynamic performance for load variation has been discussed earlier. Through a large number of open-loop tests, it can be found that the coil temperature changes very slowly with time. In addition, the time interval between the two current pulses is more than 15 ms, during which all the boost sub-modules are not actually connected to the load but charging the capacitor instead. Therefore, in combination with the load characteristics, a lower adjustment speed can meet the requirements of stable peak current.

Here we adjust the conduction time $t_{on}$ by look-up table as shown in Figure 13. The Hall sensor is used to capture the peak value $I_p$ of current pulse and the equivalent digital voltage is transferred to the logic controller by AD converter.

When the conspicuous difference between the peak value $I_p$ and reference current $I_{ref}$ exceeds the preset value $\Delta I$, the time interval $\Delta t = t_a - t_{on}$ is fed into a multiplexer, which consists of different options about the conduction time. The digital controller will select the appropriate conduction time $t_b$ to charge the capacitor in the next cycle according to the time interval $\Delta t$ and $I_p$. 

FIGURE 10  Controller architecture of 4-channel charging circuit

FIGURE 11  Logical circuit design for the proposed voltage control method

FIGURE 12  Operational waveforms of the proposed voltage control method
It can be seen that the smaller the difference $\Delta t_s$ between the preset conduction times, the higher adjusting precision of the controller will be. Obviously, increasing the number of choices $m$ can achieve the same goal. The topology and the load characteristics make the above simple control method effective for the stable output of current pulses. Furthermore, the control process mentioned above only requires a small amount of logical resources for FPGA so that its stability is improved.

5 | EXPERIMENTAL RESULTS

Experimental works are provided for verifying the effectiveness of the proposed high-current pulsed source in this section. The main component parameters used in the experimental platform are $U_S = 75$ V, $L = 100$ $\mu$H, $L_o = 2.1$ mH and $R_o = 150$ m$\Omega$. A capacitor bank of 1.5 mF and withstanding voltage 1200 V was employed as an energy-storage device based on the analysis of Section 3. The H-bridge inverter consisted of four IGBT switches $S_1$–$S_4$ (FF450R17ME4). Four MOSFET switches $Q_1$–$Q_4$ (IMW120R030M1H) were adopted for the interleaved charging submodules.

To implement the proposed control strategy, a Cyclone device EP3C16E144 was used as the central controller. Its abundant logical resources and architectural event counters made it an outstanding candidate of system control for the proposed current source. A fast and precise voltage comparator AD790 was employed to realize the voltage control scheme. In order to have an appropriate safety margin, 1000 V was designated as the ultimate voltage level of the capacitor bank within the program.

Experimental waveforms of the load current $i_o$ and the input current $i_{in}$ when the source enters steady state are shown in Figure 14. It can be observed that the charging current maintains a stable output when H-bridge inverter goes into a sleep mode. The large output current pulses and the voltage of energy-storage capacitor can be seen from Figure 15. The peak value of current pulse is more than 400 A and the capacitor voltage varies in a controllable range under the action of charging circuit.

The detailed waveforms of charging current $i_{in}$ and single-channel inductor current $i_{L2}$ are presented in Figure 16. As we can see, the boost submodules is close to working in BCM mode by the designed controller described in Section 4.

The proposed controller is operated at different conduction time, $t_{on}$. The output peak current $I_p$ for the maximum charging time is measured to evaluate its effect on the efficiency of the current source. Another controller with CFC method is also
CONCLUSION

This work has presented a high-current pulsed source suitable for specific inductive load. An N-channel interleaved boost module is designed to replenish the energy consumed for current pulse generating per cycle. The relationship between main steady-state parameters has been analysed in detail and it helps to determine the capacity of the energy-storage capacitor and selection of other parts. To improve efficiency of the charging circuit, a control strategy with simplified peripheral circuitry is proposed. Then, the steady-state output peak current could be adjusted on a large scale by the voltage control method. The closed-loop constant-peak-current method can easily realize stable output current for load variation. A digital controller based on FPGA has been designed for the aforementioned control scheme. Compared with several topologies with similar purposes, the proposed scheme has high charging efficiency and low current ripple. Moreover, it simplifies the control process and improves the stability of the system. A laboratory prototype with the maximum output peak current of 400 A is implemented, which confirms desirable functionality of the proposed pulsed source. Experimental results have verified that the topology and control strategy are effective and suitable for ATEM application.

TABLE 1 Comparison of three schematics for TEM application

| Scheme | Method | Whether transformer is included | Theoretical $I_p$ | Overall efficiency |
|--------|--------|-------------------------------|------------------|-------------------|
| [16]   | CVM    | No                            | <200 A           | <60%              |
| [5]    | CPM    | Yes                           | 250–500 A        | 80–90%            |
| This paper | CPM     | No                            | 100–650 A        | >92%              |

This paper tested for comparison. As the results in Figure 17 indicate, the proposed method lifts the peak current by 30–50 A, which contributes to stronger magnetic field and more accurate detection in ATEM application.

Contrasting the approach with the charging mode by isolated converter introduced in [5], it can be found that the way to adjust the charging voltage and current pulse is more flexible. The inductor current is pumped into the capacitor and charge it to a certain level defined by digital controller. What is more, the dynamic regulation of the topology for load variation is more simple and efficient. Compared with the charging scheme of single boost module proposed in [16], the ripple of input current is significantly reduced. In addition, its charging power only changes on a small scale which leads to a higher efficiency to raise the capacitor voltage. Specific comparison results are shown in Table 1. In one word, the N-channel interleaved boost module is especially suitable for the charging circuit of ATEM application because of its flexible adjustment and high efficiency.

6 | CONCLUSION

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