A New Modular XRAM-Like Inductive High-Current Pulse Generator Circuit Topology

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ABSTRACT XRAM (MARX spelt back words) is currently a very important circuit for high current pulse generators. In our previous studies, an XRAM-like circuit was proposed based on multiple pulse transformer modules and a capacitor connected in parallel. Compared with the traditional XRAM circuit, the same number of inductive energy storage modules can be used to generate higher current pulses. This circuit is also capable to recover the residual energy and generate repetitive high-current pulses. However, this circuit topology requires more power electronic switches (three IGBT switches per pulse transformer module). In order to reduce the number of switches, by adding two IGBT switches in the capacitor module, a new modular XRAM-like circuit with reduced switches (one IGBT switch per pulse transformer module) is proposed in this paper. The change of the circuit topology makes the capacitor discharge the primary inductors of the pulse transformer in series during the residual energy recovery stage, instead of the parallel discharge in the previous circuit. This requires the closing time of the IGBT switch in each pulse transformer module ahead of the discharge start time of the capacitor. Working process of the proposed circuit was described in detail and simulations of a 12-module circuit were carried out to verify the circuit operation. Finally, the preliminary experimental results of a two-module laboratory prototype are presented. The results confirm the theoretical analysis and show the validity of the converter scheme.

INDEX TERMS Pulsed power supplies, pulse circuits, electromagnetic launching, power semiconductor switches, pulse transformers.

I. INTRODUCTION High-current pulsed power technologies have been applied in the field of Electromagnetic Launch (EML) such as electrothermal-chemical guns and railguns [1]–[3], which are based on the conversion of a low-power, long-time input to a high-power, high-current, and short-time output. During recent years, many types of high current pulse generators based on inductive energy storage have been published in many literatures. There are three main kinds of basic circuit topologies for inductive pulsed power generators: XRAM based circuits, meat-grinder based circuits and pulse transformer based circuits.

The term XRAM generator is derived from the famous high-voltage generator "MARX". In the MARX generator circuit, capacitors are charged in parallel and discharged in series to generate high-voltage pulses. By reversing the letters, the term XRAM indicates that the generator is the inductive counterpart to the MARX generator and its goal is to generate high-current pulses. This kind of circuit topology is also called CMIS (Current Multiplier by Inductive Storage) in [4], [5]. In order to turn off higher charging current, a technique named inverse current commutation with semiconductor devices (ICCOS) was applied to XRAM in [6]–[8]. This ICCOS technique possesses the capability of interrupting tens of kilo-ampere current.

The meat grinder circuit was first presented in [9]. It is easier to achieve high current multiplication ratios with very few inductor stages. However, a very high voltage is generated across the main opening switch. On this issue, the slow transfer of energy through capacitive hybrid (STRETCH) meat grinder circuit is presented in [10], [11]. By adding an auxiliary capacitor in the traditional meat grinder circuit, the voltage stress of the opening switch is reduced. To further
improve the performance of the STRETCH meat grinder circuit, a series of circuit topologies have been presented in [12]–[14].

The pulse transformers can be used for current amplification when the turns of the secondary winding are less than the primary winding. In general, the primary winding of the pulse transformers is used for energy storage. Current pulses are generated in the secondary winding by using an opening switch to rapidly reduce the current in the primary. In order to reduce the energy loss and the power of the primary power supply during the charging process, superconducting coils have been used for pulse transformer primary windings by several researchers [15]–[17].

By combining the pulse transformers with the XRAM topology, a modular XRAM-like circuit topology is presented in our previous studies [18], which can obtain a high amplification factor and generate repetitive current pulses. The discharge of this circuit begins by connecting the primary windings of all the pulse transformers in parallel to discharge a capacitor. The recovery of the residual energy is achieved by discharging the capacitor in a positive direction to the parallel primary windings. However, this circuit topology requires more power electronic switches (three IGBT switches per pulse transformer module), which requires more synchronous control signals.

This paper proposes a new modular XRAM-like circuit topology based on pulse transformers. Compared with the circuit in [18], by making the capacitor discharge the primary inductors of the pulse transformer in series during the residual energy recovery phase, the proposed circuit is more suitable for the modular design with reduced IGBT switches. Only one IGBT switch is required in each pulse transformer module. Operation principles and features of the proposed circuit are presented. The simulation results of a 12-module circuit and preliminary experimental results of a 2-module structure are provided to demonstrate the operation principle. Finally, comparative studies with the existing topologies are presented.

II. XRAM-LIKE CIRCUIT WITH THE CAPACITOR CONNECTED IN PARALLEL

The repetitive high-current pulse generator is an important device for the development of continuous EML technology. Fig.1 shows the XRAM-like circuit based on multi-module pulse transformers proposed in the previous studies, whose capacitors are connected in parallel. Each pulse transformer module contains a pulse transformer, three IGBT switches and four diodes.

The working process is mainly completed by three switch control steps. In the charging phase, the primary inductors of all pulse transformer modules are charged in series by turning on the series IGBT switches \((S_1 - S_n)\) and turning off the parallel IGBT switches \((P_{11} - P_{n1})\) and \((P_{12} - P_{n2})\). In the discharging phase, the series IGBT switches are turned off simultaneously, forcing the primary inductors of all modules to discharge the capacitor \(C\) in parallel. This makes the secondary inductors of all modules induce a large current pulse to discharge the load. When the width of the load current pulse reaches the designated value, by turning on the parallel IGBT switches simultaneously, the capacitor \(C\) discharges the parallel primary inductors positively. This allows most of the residual energy to be transferred back to the primary inductors. By repeating the above steps, continuous current pulses can be obtained.

FIGURE 1. XRAM-like circuit with the capacitor connected in parallel proposed in the previous study.

III. PROPOSED CIRCUIT WITH REDUCED SWITCHES

It can be seen from Fig. 1 that there are more switches in the circuit topology, especially parallel IGBT switches, which require more control signals. During the residual energy recovery phase, the residual energy in the secondary loop is mainly recovered by making the capacitor provide a forward voltage drop for the primary inductors. The function of parallel IGBT switches is to connect the capacitor and the primary inductors in parallel. If the capacitor provides a forward voltage drop to the primary inductors through the series switches, the parallel IGBT switches in each pulse transformer module can be omitted. Therefore, we propose a new modular XRAM-like circuit with reduced switches, as shown in Fig. 2.

Fig. 2(a) shows the n-module prototype of the proposed repetitive high-current pulse generator circuit. It includes a primary power supply module, a capacitor module and \(n\) pulse transformer modules. A DC power supply \(V_s\) and a diode \(D_s\) are used to form the primary power supply module. The capacitor module consists of a capacitor and two IGBT switches. Each pulse transformer module consists of a pulse transformer, three diodes and an IGBT switch. Typical configuration of the capacitor module and the \(j\)-th pulse transformer module \((1 < j < n)\) are shown in Fig. 2(b) and (c). Each working cycle of the proposed circuit can also be divided into three phases: the charging phase, the discharging phase, and the residual energy recovery phase.

A. THE CHARGING PHASE

Fig. 3 shows the charging phase of the proposed topology. During the charging phase, by turning on the series
The charging process can be expressed as

\[ p \frac{di}{dt} + p - s_nL = 0 \]  

(1)

where \( L_p \) and \( R_p \) are the total inductance and total resistance of the charging loop, respectively. \( L_{pj} \) and \( R_{pj} \) are the primary inductance and resistance of the \( j \)-th module, respectively.

The charging current can be obtained as follow:

\[ i_p = I_{pm}(1 - e^{-t/\tau}) = \frac{V_s}{R_p}(1 - e^{-t/\tau}) \]  

(4)

where \( I_{pm} \) is the designated maximum charging current.

Therefore, the maximum energy storage of inductors in this system is

\[ E_{pm} = \frac{1}{2}L_p I_{pm}^2 = \frac{1}{2} \sum_{j=1}^{n} L_{pj} I_{pm}^2 \]  

(5)

It can be inferred from the charging phase that when the internal resistance of the primary inductors is high, the charging process must be fast, otherwise the energy loss on the inductor coils will be very large. This requires that the primary power supply must have a higher power, especially when there are more pulse transformer modules. Therefore, to reduce the power of the primary power supply, it is necessary to select primary inductors with lower internal resistance, such as superconducting inductors.

B. THE DISCHARGING PHASE

When the charging current reaches the designated value, the proposed circuit can start the discharging phase by turning off the switches \( S_1-S_0 \). The discharging phase can be divided into two steps. Equivalent circuits of the converter during these steps are presented in Fig. 4(a) and (b).

Fig. 4(a) shows the step 1 of the discharging phase. In this step, the switches \( S_1-S_0 \), \( S_{c1} \) and \( S_{c2} \) are off. The primary inductors \( L_{p1}-L_{pn} \) are connected in parallel to discharge to the capacitor \( C \) so that the current in each primary inductor quickly decays. To support the mutual flux, the secondary inductors \( L_{s1}-L_{sn} \) induce large current pulses for the load.

The voltage equation of the \( j \)-th module in the discharge step can be expressed as

\[ \begin{align*}
L_{pj} \frac{di_{pj}}{dt} + M_j \frac{di_{sij}}{dt} + R_{pj}i_{pj} + u_C &= 0 \\
L_{sij} \frac{di_{sij}}{dt} + M_j \frac{di_{pj}}{dt} + R_{sij}i_{sij} + R_L \sum_{j=1}^{n} i_{sij} &= 0
\end{align*} \]  

(6)

where \( L_{sij} \) and \( R_{sij} \) are the secondary inductance and resistance of the \( j \)-th module respectively, \( u_C \) is the voltage of the capacitor, and \( R_L \) is the resistance of the pulsed load.

In the modular design, the circuit parameters of each pulse transformer module are the same. Therefore, in this discharge step, the capacitor current \( i_C \) is \( n \) times the primary current \( i_{pj} \), and the load current \( i_L \) is \( n \) times the secondary current \( i_{sij} \). Thus, the equivalent equation of the \( n \)-module system at this discharge step can be derived as

\[ \begin{align*}
\frac{1}{n}L_{pj} \frac{di_C}{dt} + \frac{1}{n}M_j \frac{di_L}{dt} + \frac{1}{n}R_{pj}i_C + u_C &= 0 \\
\frac{1}{n}L_{sij} \frac{di_L}{dt} + \frac{1}{n}M_j \frac{di_C}{dt} + (\frac{1}{n}R_{sij} + R_L)i_L &= 0
\end{align*} \]  

(7)
FIGURE 4. Equivalent circuit of the discharging phase. (a) Step 1 of the discharging phase. (b) Step 2 of the discharging phase.

Fig. 4 (b) shows the step 2 of the discharging phase. When the primary current decays to zero, the capacitor voltage reaches its maximum value. Then, the secondary inductors $L_{s1}$–$L_{sn}$ are connected in parallel to discharge to the load. Since the switches $S_{c1}$ and $S_{c2}$ are turned off, the capacitor maintains its voltage in this step. Thus, the voltage equation of the j-th module can be expressed as

$$L_{sj}\frac{di_{sj}}{dt} + R_{sj}i_{sj} + R_{L}i_{L} = 0 \quad (8)$$

By adding the voltage equations of n modules, the equivalent equation of the n-module system at this discharge step can be derived as

$$\frac{1}{n}L_{sj}\frac{di_{L}}{dt} + \frac{1}{n}(-R_{sj} + R_{L})i_{L} = 0 \quad (9)$$

Obviously, the step 2 of the discharging phase can be equivalent to a first-order RL circuit.

C. THE RESIDUAL ENERGY RECOVERY PHASE

When the load current pulse reaches the required width or decays below the required value, the residual energy recovery phase can be started by turning on the switches $S_1$–$S_n$, $S_{c1}$ and $S_{c2}$, simultaneously. The equivalent circuit of this phase is shown in Fig. 5. In this period, the capacitor charges the primary inductors in series and the diode $D_s$ prevents the capacitor current passing through $V_s$. Because the capacitor provides a forward voltage drop for the primary inductors, the residual energy in the secondary circuits is quickly transferred back to the primary inductors.

The circuit equation of the residual energy recovery phase can be expressed as

$$\begin{align*}
\sum_{j=1}^{n} L_{pj}\frac{di_{p}}{dt} + \sum_{j=1}^{n} M_{j}\frac{di_{sj}}{dt} + (\sum_{j=1}^{n} R_{pj} + R_{line})i_{p} - u_{C} &= 0 \\
L_{sj}\frac{di_{sj}}{dt} + M_{j}\frac{di_{pj}}{dt} + R_{sj}i_{sj} + R_{L}i_{L} &= 0
\end{align*} \quad (10)$$

Because the parameters of each module are the same, the equivalent equation of the n-module system in this phase can be expressed as

$$\begin{align*}
L_{p}\frac{di_{p}}{dt} + M_{j}\frac{di_{L}}{dt} + R_{p}i_{p} - u_{C} &= 0 \\
\frac{1}{n}L_{sj}\frac{di_{L}}{dt} + M_{j}\frac{di_{pj}}{dt} + (\frac{1}{n}R_{sj} + R_{L})i_{L} &= 0
\end{align*} \quad (11)$$

When the capacitor voltage drops to be equal to the voltage of the primary power supply, the circuit automatically re-enters the charging phase and a new working cycle begins. Obviously, after each working cycle, the capacitor has a small amount of residual energy determined by the voltage of the primary power supply.

Assuming that the current recovered by the primary inductors at the end of the residual energy recovery phase is $I_{rec}$, the total recovered residual energy is

$$E_{rec} = \frac{1}{2}L_{p}I_{rec}^2 \quad (12)$$

In order to analyze the energy distribution during the discharging process of a cycle, the ratio ($\eta_{load}$) of the energy obtained by the load to the maximum inductive energy storage, the ratio ($\eta_{rec}$) of the recovered energy by the primary inductor to the maximum inductive energy storage, and the ($\eta_{loss}$) ratio of the energy consumed by semiconductor switches and primary resistors to the maximum inductive energy storage can be calculated as

$$\begin{align*}
\eta_{load} &= \frac{\int_{0}^{T} P_{load}dt}{\int_{0}^{T} P_{max}dt} \\
\eta_{rec} &= \frac{E_{rec}}{E_{max}} \\
\eta_{loss} &= \frac{E_{loss}}{E_{max}}
\end{align*} \quad (13)$$

where $P_{load}$ is the power dissipated in the load, $P_{max}$ is the maximum power dissipated in the load, $E_{max}$ is the maximum energy stored in the inductors, $E_{rec}$ is the recovered energy, and $E_{loss}$ is the energy dissipated by semiconductor switches and primary resistors.
energy storage are defined as follows:

\[
\eta_{\text{load}} = \frac{E_{\text{load}}}{E_{\text{pm}}} = \frac{\int_{T_{n-1}}^{T_n} R_L i^2 \, dt}{E_{\text{pm}}} \times 100\% \quad (13)
\]

\[
\eta_{\text{rec}} = \frac{E_{\text{rec}}}{E_{\text{pm}}} \times 100\% \quad (14)
\]

where \(T_{n-1}\) and \(T_n\) are the start and end times of the \(n\)-th cycle, respectively.

\[
\eta_{\text{loss}} = (1 - \eta_{\text{load}} - \eta_{\text{rec}}) \times 100\% \quad (15)
\]

Since the residual energy is recovered, the actual energy transfer efficiency from the primary side to the load is defined as the ratio of the energy obtained by the load to the actually reduced energy of the primary inductors.

\[
\eta = \frac{\eta_{\text{load}}}{1 - \eta_{\text{rec}}} \times 100\% = \frac{\eta_{\text{load}}}{\eta_{\text{load}} + \eta_{\text{loss}}} \times 100\% \quad (16)
\]

IV. SIMULATION AND EXPERIMENTAL VERIFICATION

A. SIMULATION

In order to verify the principle of the proposed circuit and compare it with the previous circuit in [18], a 500-kJ system with twelve high-temperature superconducting (HTS) pulse transformer modules was used for simulation. The specific circuit parameters are shown in Table 1, among which the parameters of pulse transformers refer to the literature [19]. Assuming that the cooling temperature of the HTS pulse transformers is 40-K, the calculated primary critical current is greater than 1.1-kA. Within this current value, the primary inductors can be assumed to be in superconducting state.

TABLE 1. Parameters of the twelve-module system for simulations.

| Designation            | Value          |
|------------------------|----------------|
| Primary inductors \(L_p\) | 74.47 mH       |
| Secondary inductors \(L_q\) | 730.83 \(\mu\)H |
| Mutual inductance \(M_{pq}\) | 7.156 mH       |
| Primary resistance \(R_L\) | Ignored        |
| Secondary resistance \(R_q\) | 2 m\(\Omega\)   |
| Calculated primary critical current | > 1.1 kA (at 40 K) |
| Number of stages \(n\) | 12             |
| Capacitor \(C\)         | 800 \(\mu\)F    |
| Primary supply \(V_t\)  | 60 V or 90 V   |
| Load resistance \(R_L\) | 2 m\(\Omega\)   |
| Operation frequency    | 0.1 Hz         |

In addition, since there are no diodes connected in series with the series IGBT switches \(S_1-S_n\), compared with the circuit in [18], the proposed circuit has relatively lower voltage requirements for the primary power supply. In order to obtain similar charging currents at the same operating frequency, the primary power supply voltage of the proposed circuit is set to 60-V in the simulation, and the primary power supply voltage of the circuit in [18] is set to 90-V. The on-state voltage drop and on-state resistance of the semiconductor switches (IGBTs and Diodes) is assumed to be 2-V and 1 m\(\Omega\), respectively.

FIGURE 6. Control signals of switches.

FIGURE 7. Simulated waveforms of the two circuits with 12 modules and similar charging currents in continuous operating cycles.
Fig. 7 shows the simulated waveforms of the two circuits in continuous operation mode. Fig. 8 shows the simulated waveforms of the two circuits in a stable discharge phase between 119.99 and 120.01 s. It can be seen that the simulated waveforms of the two circuits are basically the same in the discharging phase, but there are obvious differences in the residual energy recovery phase. Since the capacitor discharges the primary inductors in series in the proposed circuit, the recovery time of the residual energy is relatively long. This is reflected in the longer the cut-off time of the load current pulse and the longer the recovery time of the primary current. However, the voltage of the primary inductor during the residual energy recovery phase is also significantly reduced. This reduces the number of high-voltage shocks that the primary inductor experiences during each work cycle. The comparison of some performance indices of the two circuits is listed in Table 2, where $P_{\text{Vin}}$ is the maximum output power of the primary power supply, $I_{\text{load}}$ is the amplitude of the load current pulse, and $\gamma$ is the amplification factor of the load current pulse to the charging current.

The simulation results confirm the principle of the proposed circuit. First of all, under the same continuous current pulse output, the number of IGBT switches of the proposed circuit is significantly reduced. Secondary, in the case of similar charging currents, the proposed circuit has lower requirements for the primary power supply voltage. As a result, the maximum output power of the primary power supply can be significantly reduced. Finally, the energy transmission efficiency of the system has also been improved. However, it should be noted that the duration of the remaining energy recovery phase of the system is also significantly increased.

## B. PRELIMINARY EXPERIMENTAL VERIFICATION

To further verify the proper operation of the proposed circuit, an experimental setup with two small HTS pulse transformer modules was assembled and tested. The experimental circuit is shown in Fig. 9. Due to the manufacturing process, there is a small difference between the parameters of the two HTS pulse transformer modules. The detailed parameters are listed in Table 3. The experimental waveforms are shown in Fig. 10 and 11, respectively.

Fig. 10 shows the experimental waveforms of the two HTS pulse transformer modules in continuous operation mode. Fig. 11 shows the experimental waveforms in the stable discharge phase between 0.794 and 0.802 s. In the experiment, the maximum charging current obtained

### TABLE 2. Comparison of some performance indices of the two circuits.

| Parameters      | Circuit in [18] | Proposed circuit |
|-----------------|-----------------|-----------------|
| $I_{\text{Vin}}$| 997.4 A        | 1033.3 A        |
| $E_{\text{Vin}}$| 444.7 kJ        | 477.3 kJ        |
| $P_{\text{Vin}}$| 89.82 kW        | 61.98 kW        |
| $I_{\text{load}}$| 115.9 kA       | 120.2 kA        |
| $\gamma$        | 116             | 116             |
| $E_{\text{load}}$| 160.5 kJ        | 177.5 kJ        |
| $\eta_{\text{load}}$| 36.09%       | 37.19%         |
| $E_{\text{Rec}}$| 767.2 A        | 789.1 A        |
| $\eta_{\text{Rec}}$| 59.14%       | 58.29%         |
| $\eta_{\text{Loss}}$| 4.77%       | 4.52%         |
| $\eta$          | 88.34%         | 89.17%         |
TABLE 3. Parameters of the 2-module experimental setup.

| Parameters          | Module 1            | Module 2            |
|---------------------|---------------------|---------------------|
| Primary inductance  | 6.45 mH             | 6.42 mH             |
| Secondary inductance| 34.2 μH             | 34.5 μH             |
| Coupling coefficient| 0.95                | 0.95                |
| Primary resistance  | 1.3 mΩ              | 1.3 mΩ              |
| Secondary resistance| 2.2 mΩ              | 2.8 mΩ              |
| Primary critical current | 110 A (at 77 K)  | 110 A (at 77 K)    |
| Line resistance     | 60 mΩ               |                     |
| Capacitor C         | 200 μF              |                     |
| Load resistance     | 2 mΩ                |                     |
| Primary dc supply   | 13.5 V              |                     |
| Operation frequency | 5 Hz                |                     |

is 103-A, and the output current pulse amplitude is about 2.34-kA. The current amplification factor is 22.7. The primary current waveforms of the two HTS pulse transformer modules are basically the same. This shows that the small parameter difference between the two actual HTS pulse transformer modules has a negligible effect on the working process. In addition, since there are only two HTS pulse transformer modules and their inductance values are small, during the residual energy recovery phase, the primary inductor voltage reaches half of the capacitor voltage, and the cut-off time of the load current pulse and the recovery time of the primary current are both relatively short. On the whole, the change trend of the experimental waveforms is consistent with the principle analysis of the working process.

V. COMPARATIVE STUDY

The proposed current pulse generator circuit can also be compared with the other three existing topologies that were presented in [5], [8], and [13]. In these circuits multi-module inductors are used for energy storage and the generation of higher current pulses. The specifications of these circuits are summarized in Table 4. The number of primary power supplies, switches, diodes, capacitors, output current, as well as residual energy recovery and repetitive pulse output capability are used to compare different topologies.

The current pulse generator circuits in [5] and [8] are two improved XRAM circuits. Their current amplification factor is relatively small. The circuit in [13] is an improved
TABLE 4. Comparison of multi-stage current pulse generator circuits.

| References | Input current | Iₘ | I₀ | Iₙ | Iₘ⁺² |
|------------|---------------|---|---|---|---|
| [5]        | Iₘ | I₀ | Iₙ | Iₙ |
| [8]        | n   | n  | n  | n  |
| [13]       | 1   | 1  | 1  | 1  |
| Proposed   | 2n  | 0  | n² | 1  |

STRETCH meat grinder circuit. The circuits in [8] and [13] use thyristors as the opening switches. In order to turn off these thyristors, each thyristor must be equipped with a capacitor with a pre-charge voltage. Therefore, these two circuits require more capacitors, and the generation of repetitive current pulses is not mentioned. The circuit proposed in this paper belongs to the XRAM composed of multi-module pulse transformers. Its current amplification factor is the product of the amplification factor (γ) of the pulse transformer module and the number of modules. The overall comparison shows that the proposed circuit topology has good performance in generating repetitive high-amplitude current pulses with reduced components.

VI. CONCLUSION
A new modular XRAM-like repetitive high-current pulse generator circuit is proposed in this paper. Compared with the previous similar circuits, the presented circuit topology is composed of pulse transformer modules that reduce the number of IGBT switches and the number of series diodes. This simplifies the circuit topology and reduces the maximum output power of the primary power supply. The working procedure of the proposed circuit is described in detail. In order to verify the proper operation of the proposed circuit, simulations were carried out with a 12-module system and preliminary verification experiment was carried out with two HTS pulse transformer modules. The simulation and experimental results confirm the principle of the proposed circuit.

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