Pulse current generator with fast rise time based on transformers and single active switch for plasma drilling

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Abstract
This paper presents a new pulse generator topology based on the flyback converter and the XRAM generator concept. The presented topology has the advantage of only requiring a single active switch which results in a simpler control and a higher robustness. Those requirements are essential for plasma drilling, for which the pulse generator has to operate in harsh environment with a high ambient temperature and high mechanical stresses. This paper presents the design procedure of the pulse transformer of the proposed topology and results for a first design.

1 Introduction
Geothermal energy is a promising renewable energy source for continuous power generation for the base load. To access the energy, holes need to be drilled to a depth of up to 10 km. At such depths, the ambient temperature reaches up to 300 °C, which makes it possible to efficiently produce electricity [1]. However, the cost of drilling deep holes with a mechanical grinding process grows exponentially with the depth. Therefore, conventional drilling is typically not very cost-effective for holes deeper than 5 km [2].

To reduce the drilling cost, new drilling concepts have been proposed, including laser cutting and drilling [3], rock spallation with a hydrothermal flame [4] or electro pulse drilling [2]. Another promising concept is plasma drilling [5], which uses a pulsed plasma to generate fast temperature changes at the surface of the rock to disintegrate the rock by thermal contraction. For plasma drilling, current pulses with a fast rise time in the range of a few 100 ns and large pulse energy are required. In addition, a relatively low constant current for a pilot arc is required in between the pulses (fig.1), so that always a small pilot arc burns. The pilot arc is initially triggered with a high voltage igniter next to the electrode.

Due to the influence of parasitics, the pulse generator must be placed close to the load in order to achieve fast current pulse rise times. In the considered application, this means that the generator must be installed in the drill head, which is exposed to high ambient temperatures and pressure. To protect the pulse current generator, it is placed inside a protection vessel and connected via a cable (a MV transmission line) to
a power supply with a switch at the surface. Because the conditions in the protection vessel are harsh in terms of temperature and EMI, actively controlled switches are not suitable for integration in the protection vessel.

Possible concepts for the pulse current generator for the considered plasma drilling application are pulse transformers [6], meat grinders [7] and XRAM generators [8]. All those topologies have fast current rise times and offer current multiplication. Especially the XRAM generator has the additional advantage of being modular, which offers scalability in terms of power and current for the pulse drilling application. In the considered application, the XRAM modules would need to fit inside the protection vessel so that the active switches and the related control hardware would be exposed to the high ambient temperature which decreases the reliability of the system. In order to overcome this limitation, a new concept which does not require switches inside the protection vessel is presented in this paper (Fig. 1). The concept is modular and utilises transformers for current adding/multiplication. The operation is controlled with a single switch, which could be located at the surface and the new concept can inherently generate the current for the pilot arc.

Section 2 presents the new concept and its underlying operating principle. Thereafter a model of the transformer with its parasitics is introduced and the effects on the rise time of the pulse current is explained. The following section 3 introduces the design procedure of the transformer with a subsection on the calculation of the parasitics. Finally in the last section 4, a possible transformer design based on specifications for the application is presented.

2 Single switch current adder (SSCA)

The proposed single switch current adder (SSCA), shown in fig. 1, requires no active switch within the protection vessel and the power supply as well as the switch for controlling the operation can be located at the surface and connected with a cable to the modules in the bore hole. A single module \( n \) in the protection vessel consists of transformer \( T_n \) and diode \( D_n \). The concept of a module is based on a flyback converter, i.e. the transformer is used for storing energy.

The operation of the SSCA could be divided in two phases. In the pilot arc phase, i.e. in the time interval \( t_1 \) to \( t_2 \), the pilot arc burns between the electrode due to the pilot current, which also is used to store energy in the flyback transformer. In the subsequent pulse phase, i.e. in the time interval from \( t_2 \) to \( t_3 \), the energy is transferred from the flyback transformer to the (pilot) arc.

During the pilot arc phase, switch \( S \) connects all the primary windings of the transformers in series to the load and the windings of the transformers are interconnected in such a way, that the diodes \( D_n \) are reverse biased. The charging current \( i_{in} \) flows through the series connected primary windings and also across the electrode, so that the converter inherently generates the required current for the pilot arc (current path: blue doted line in fig. 1). At the end of the pilot arc phase, switch \( S \) is opened and the transformers commutate the current to the secondary winding and forward bias diodes \( D_n \). Consequently, all secondary windings are connected in parallel to the load \( R_L \) and the current in the flyback transformer is added up, i.e. multiplied by the number of modules \( n \). The commutation from a series to a parallel connection is similar in flyback converters and/or XRAM system. At the end of the pulse phase, switch \( S \) is closed again and the cycle starts over again.

The maximum current during the pilot arc phase is reached by the end of the interval and depends mainly on the value of the load resistor \( R_L \), the primary winding resistance \( R_{pri} \) and the input voltage \( V_{in} \). In the following section, the detailed current and voltage waveform for the modules during the two phases are presented.

2.1 Operation principle

In the SSCA primary windings of the transformer are connected in series in phase 1 and all secondary windings are connected in parallel to the load in phase 2. This results in different input \( V_{in,n} \) and output \( V_{out,n} \) voltage waveforms for each module. In contrast the primary and secondary current waveform are the same for all modules. Fig. 2a shows the first module 1 which is connected to the electrode and fig.
2b the last module n which is connected to the cable. On the left, the input voltage $v_{in}$ and the primary current $i_{pri}$ waveforms are shown and on the right, the output $v_{out}$ and the secondary current $i_{sec}$ waveforms are presented.

In the following description of the basic current and voltage waveforms, the transmission line and parasitic components are neglected for the sake of simplicity and it is assumed that the SCCA is connected to an ideal voltage source at the input.

**Pilot arc phase**

At the beginning of the pilot arc phase, switch $S$ is closed at $t = t_1$, so that all the primary windings and the load are connected in series to the input voltage $V_{in}$. The primary current $i_{pri}$ in the $RL$ circuit with the time constant $\tau = \frac{L_m}{R_L}$, increases exponentially and all the diodes $D_n$ are reverse biased.

The input voltage $v_{in,n}$ at module $n$ is equal to the input voltage $V_{in}$ as shown in fig. 2b. Due to the series connection of the primary windings, the magnetizing inductor acts as an inductive voltage divider during the pilot phase, what defines the input and output voltages of all modules. Since the inductive voltage drop across the magnetizing inductors $L_m$ decays, the voltage $v_{out}$ across $R_L$ becomes equal to $V_{in}$ at the end of the pilot arc phase.

**Pulse phase**

At the beginning of the pulse phase, switch $S$ is opened and in case of an ideal transformer ($L_\sigma = 0$), the primary current $i_{pri}$ commutes immediately to the secondary winding. Due to the parallel connection of the secondary windings, the load current rises to:

$$i_L = \sum_{\zeta=1}^{n} i_{sec,\zeta}$$  \hspace{1cm} (1)

The load voltage $v_L$ as well as the secondary voltages $v_{sec}$ of modules increase with the rising current through the load i.e. the arc. The induced voltages onto the primary winding add due to the series
connection of the modules. Therefore, the maximum voltage value results for input voltage $v_{in,n}$ of module $n$ which is equal to the sum of all primary winding voltages and the load voltage.

2.2 Impact of parasitics on the pulse performance

In order to evaluate the behaviour of a non-ideal SSCA, a model of the transformer including the parasitics is essential. Generally, a transformer has two main parasitics which significantly impact the pulse behaviour of the system. Those parasitics are the leakage inductance and the stray capacitance shown in fig. 3. The magnetic behaviour of the transformer is modelled with a T-equivalent circuit assuming that the leakage inductance is transferred to the primary. In general, the stray capacitances of the transformer can be represented with the “six capacitor model (6C)” presented, for example, in [10]. The capacitors of the 6C-Model depend on the geometry of the core and the winding. In the case of the SSCA topology, the model is simplified to one interwinding capacitor $C_{stray}$ (Fig. 3a), since they have the largest effect on the current rise time as explained in the following.

For the proposed application, the main goal is to achieve a fast current rise time of the pulse and therefore, the current and voltage waveforms before and after the start of the pulse at $t = t_2$ are analysed in detail (see fig. 3b/c).

Voltage $v_{in}$ and voltage $v_{out}$ across $R_L$ become equal to $V_{in}$ and voltage $v_c$ is zero assuming that shortly before $t = t_2$, the primary winding voltages $v_{pri}$ decayed to zero and the current $i_{L,t}$ is equal $I_{max}$. At the moment $t = t_2$, switch $S$ opens and the pilot arc current $i_{pilot}$ from the voltage source becomes zero, shown as a blue dotted line in fig. 3a/c. Leakage current $i_{L,t}$ must be continuous and commutes to the stray capacitance $C_{stray}$. The stray capacitor current $i_{C,stray}$ is shown as green dotted line in fig. 3a/c. Due to the current, the stray capacitance voltage $v_{C,n}$ (Fig. 3b) increases and starts oscillating since the leakage inductor and the stray capacitor form a series RLC circuit. The capacitor current $i_{C,stray}$ is in the opposite direction of the pulse current $i_{pulse}$ shown in the second graph in fig. 3c. Therefore, the resulting
load current $i_L$, shown as red line on the right graph in fig. 3c, is the difference between the capacitor current $i_{C, stray}$ and the pulse current $i_{pulse}$. The current rise time of a single module is therefore limited to the resonance frequency of the RLC circuit. Moreover, the stray capacitance is charged to a higher voltage moving from module 1 close to the electrodes to module $n$ close to the cable, because $v_{in,n}$ is the sum of all $v_{L\sigma,n}$, $v_{Lm,n}$ and the load voltage $v_L$ which further affects the current rise time.

3 Design procedure

In the following the design procedure of the SSCA is presented, which is split into two parts. In the first part the load model of the plasma for the circuit simulation is introduced. In the second part, the general design procedure of the transformer is shown.

3.1 Plasma load model

The pulse waveform subdivided in two phases shown in fig. 1. For both phases, the load current $i_L$ flows through the arc plasma. Therefore, the behaviour of the pulse waveform is highly depended on the V-I characteristic of the plasma. The V-I characteristic can be approximated with a conductance model used for DC arc discharges in gas environments presented in [11]. In this application, the considered gas behaves like steam. In the model the conductance $G$ of the gas is assumed to be proportional to the electron density $n_e$ of the plasma (2) and the constant $F$ describes an experimental constant which describes the relation between conductance and gas properties.

$$G = \frac{i}{v} = F n_e$$ (2)

The rate of production of electrons (3) depends on the electric power subtracted by two electron loss terms. The first loss term describes the recombination at the outer wall of the plasma and the second term expresses the recombination losses inside the plasma. Wall losses are modelled to be proportional to the electron density $n_e$ whereas the recombination losses depend exponentially on the electron density $n_e$. Each term of the rate equation has its own linear proportionality factor ($\bar{A}, \bar{B}, \bar{C}, \bar{D}$) to describes the gas properties.

$$\frac{dn_e}{dt} = \bar{A} i v - \bar{B} n_e - \bar{C} e^{\bar{D} n_e}$$ (3)

![Figure 4: (a) Arc plasma model. (b) Winding arrangement $W_1$ to $W_3$.](image)
Replacing the electron density in (3) with (2) results in (4), in which the mathematical manipulation leads
to the constants $A = \bar{A}F$, $B = \bar{B}$, $C = \bar{C}F$ and $D = \bar{D}/F$.

$$\frac{dG}{dt} = A i^2_\text{G} - BG - Ce^{DG} \quad (4)$$

The constants are fitted with experimental data presented in [12], [13] and [14]. In these publications, a
water gap was pulsed with a high DC voltage until break down was initiated. The arc through the water
gap, heats up the water and then generates plasma from the steam. The current and the voltage values
for the breakdown were measured. Publication [13] presents arcs with currents values up to 4.5 kA. The
data extracted from the measurements in [12] show a conductance between 0.5 S and 5.8 S with a time
constant between 4.8 $\mu$s and 5.1 $\mu$s. In [14], the authors increased the applied voltage step wise from
7 kV to 13 kV while varying the external pressure between 0 and 4 MPa. Furthermore, the experiment
was performed with a coaxial electrode with a water gap of 5 mm. The resulting current waveform shows
the behaviour of an underdamped RLC circuit. The arc resistance decreased with higher pulse voltage
and increased with increasing hydrostatic pressure. The measured values were in a range from 4 S to
8 S for the arc without external applied pressure. The tested peak current value was between 15 kA and
35 kA. The authors do not specify an exact current rise time but measurements on the same setup in [13]
indicate a current rise time below 50 $\mu$s.

Based on this experimental data, the parameters $A, B, C, D$ are fitted such that the conductance value
varies between 2.68 S and 0.57 S from phase 2 to phase 1. The highest conductance value is reached at
the end of the pulse interval after the plasma has heated up. The modelled time constant is set to the value
of 23 $\mu$s. The plasma model (Fig. 4a) is implemented in PLECS as capacitor in parallel to an inductor
and a controlled voltage source, in which the voltage source represents the voltage drop of the plasma.
The voltage value results from the differential equation for the conductance multiplied by the impressed
current. Besides the plasma load model, the SSCA requires a transformer design procedure to calculate
the design parameter to full fill the pulse specification which is shown in the following.

### 3.2 Transformer design procedure

The optimal selection of the core size and its core material is crucial to achieve minimal losses and the
required pulse energy $E_{\text{pulse}}$. Constraints for the optimization are the available space, the pulse energy.
The current rise time value is feed into an algorithm which determines analytically the pulse energy and
total losses.
The total losses consist of conduction losses of the winding and core losses. As the pulse rate is low (300 Hz), skin effect losses are only considered for the secondary winding which is conducting the high frequency pulse. The DC resistance is calculated with the current density of $6 \text{ A/mm}^2$. The core losses are calculated with the iGSE method [15] which uses the extracted Steinmetz parameters from the data sheet. Moreover, to the non-linear behaviour of the core material is modelled with a magnetization function given by the manufacturer for estimating the correct pulse energy.

The algorithm stores all solutions which reached the minimal required pulse energy. All the solutions are then normalized by the result with either the lowest losses, inductance or number of turns. The goal of keeping the inductance and the turn number low is to attain a low input voltage $V_{in}$ respectively achieve low parasitic values. Those three normalized values are summed up as an weighting function and weighted equally. The sum with the lowest value represents the optimal core size.

3.2.1 Calculation of parasitics

The values of parasitics components in the parasitic model in fig. 3 are determined with finite element method (FEM) simulations. The parasitics depend mainly on the geometrical arrangement of the primary and secondary winding. Therefore, three different winding topologies are evaluated. The three topologies are presented in fig. 4b. The first winding arrangement ($W_1$) has two round shaped single conductors placed next to another. Thus, the single turns are on the same radial position. For the second arrangement ($W_2$), the primary is positioned at a different radial position than the secondary winding. The last winding arrangement ($W_3$) is based on a foil winding. The conductor has a rectangular shape and is as wide as it does not touch the adjacent turn. Moreover, the primary and secondary winding are placed on top of each other.

The design process is reduced to three main parameters which determine the position of the windings. The first parameter is the distance to the core in radial direction $S_{air}$. The second one is the distance in between the two windings $S_{win}$. In case the two winding are on the same layer this parameter is linked to the third and last parameter as a function, which is the number of turns $N_{turns}$.

The circuit model of the transformer requires the values of the stray capacitance and leakage inductance of the windings. Therefore, two simulations of the it at i.e. electric and magnetic field were conducted. In case of the electric field, one winding is grounded and the other one is set to a specific potential. With the electric energy in the simulation domain the capacitance value is derived. For the leakage field both windings are set to equal contrary current values. The leakage inductance follows analogue to the stray capacitance with the energy in the magnetic field and the current value of the winding.

The calculation of the parasitics are performed for all combinations of core sizes and winding arrangement. The resulting capacitance and inductance values are used in the parasitic transformer model to calculate the pulse performance.

4 Resulting design parameters

In this section, results of the algorithm for the core design, the winding design and the pulse simulation are presented. For each section of the design procedure, the design parameters and the constrains are introduced. The solutions of each section are used as input parameter for the subsequent section. This leads finally to a feasible implementation of the transformer.

4.1 Results of core selection algorithm

The input parameters for the core optimization are the core material properties in addition to the chosen core geometry and current waveform. The two core geometries which fit into the protection vessel are presented in fig. 6b. The focus is put on powder core materials since the offer a high temperature stability, low losses and relatively soft saturation. The evaluated materials are MPP, High Flux, Kool Mu Max and XFlux. All of them have a curie temperature higher than 460°C and a relative permeability in the range of 14 to 550.
The current waveform considered for the optimization is presented in fig. 6a. The minimum pilot current to avoid extinction of the arc is set to 10 A. Furthermore, the maximum pilot current is set to 300 A. The current waveform shows a time constant for the rise and fall slope. Those value are derived from the specification of the pulse frequency and pulse duration, given in table below the core geometry. Furthermore, the design parameters are limited to the range from 6 to 16 numbers of turns, a minimum pulse energy of 10 J and 10 to 30 cores in parallel.

The lowest losses are achieved for the High Flux material. For which the solution space contains 272 possible implementations in the case of core number 1. Specifically, the core implemented with a relative $\mu_r$ of 150 is achieving the lowest losses. Fig. 7 shows the solution space for four different values of $\mu_r$ in a total losses versus number of cores plot. The core material is distinguished by different symbols whereas the number of turns is shown by a grey scale. The optimal solution is highlighted with an red arrow. The table on the right hand side of the plot shows the detailed solution for both core sizes.

### 4.2 Results of parasitic components simulations with FEM

A parametric CAD model of the transformer geometry is implemented for the FEM simulation which simplifies the parametric sweep. Furthermore, the geometry is reduced to a symmetrical element of the toroidal transformer to reduce computation time. The critical material parameters for the evaluation are the permittivity in which the transformer is operating and the core material. The transformer is placed inside an oil filled chamber and therefore the relative permittivity is selected to be 2.2.
The parametric sweep is conducted for the range from 1 mm to 10 mm for the parameter $S_{\text{win}}$ which is the distance between the windings. Furthermore, the parameter $S_{\text{air}}$ is swept from 1 mm to 3 mm. In some cases, the geometry does not offer sufficient space in the protection vessel, and therefore, these solutions are not considered further. For each solution, the electrostatic and magnetic energy of the whole domain is calculated.

The results are presented in fig. 8, which shows three graphs with the stray capacitance on the left and the leakage inductance on the right vertical axis. The top table shows the solution for core nr. 1 for which only the winding arrangement $W_1$ fits inside the protection vessel. This is due to the fact that core nr. 1 has a bigger diameter and therefore the winding has not enough space to fit in between the outer diameter of the core and the inner diameter of the protection vessel. Therefore winding arrangements $W_2$ and $W_3$ are not feasible with core nr. 1. Therefore, the only feasible solution with core nr. 1 results in 279 pF for $C_{\text{stray}}$ and 5.52 μH for $L_{\text{leakage}}$. All winding arrangements can be implemented with core nr. 2 and the resulting parasitics are shown in the three graphs.

Winding arrangement $W_3$ results in the highest stray capacitance values and the lowest leakage inductance which is expected for foil windings. The other two arrangement result in both low values for the stray capacitance as the facing area between the winding is smaller for the round conductor and the distance is higher. The lowest stray capacitance is attained with the core nr. 2 and winding arrangement $W_2$.

### 4.3 Resulting pulse rise time

The values for the parasitics resulting from the FEM simulation are evaluated with PLECS. In a first step, only the extreme values of $S_{\text{win}}$ or $S_{\text{air}}$ are simulated. By that measure, the computational time can be reduced. The results shows that the rise time is more sensitive to the capacitance than to the leakage inductance value. Thus, only the solution with the smallest stray capacitance for both core sizes are simulated. Fig. 1 shows on the X-axis the current rise time whereas on the primary and secondary Y-axis, the stray capacitance respectively leakage inductance is shown.

The plot shows the general trend that with a higher stray capacitance the rise time increases. Therefore, the best result is obtained with the lowest stray capacitance value resulting for core nr. 2 and winding arrangement 2. With this combination a pulse current rise time of 1500 ns could be achieved. Furthermore, it shows that with the winding arrangement 2 faster current rise times can be achieved as with the other two arrangements.

Figure 8: Values for the parasitics depending geometrical parameter $S_{\text{win}}$ and $S_{\text{air}}$. The circles and asterisks represent the calculated values.
Table I: Fastest achievable rise time for each core and winding combination

<table>
<thead>
<tr>
<th>Core</th>
<th>Winding arrangement</th>
<th>$C_{stray}$</th>
<th>$L_{leakage}$</th>
<th>$t_{rise}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nr. 1</td>
<td>$W_1$</td>
<td>279 pF</td>
<td>5.52 μH</td>
<td>559 ns</td>
</tr>
<tr>
<td>Nr. 2</td>
<td>$W_1$</td>
<td>346 pF</td>
<td>4.94 μH</td>
<td>575 ns</td>
</tr>
<tr>
<td>Nr. 2</td>
<td>$W_2$</td>
<td>185 pF</td>
<td>11.72 μH</td>
<td>494 ns</td>
</tr>
<tr>
<td>Nr. 2</td>
<td>$W_3$</td>
<td>314 pF</td>
<td>5.91 μH</td>
<td>575 ns</td>
</tr>
</tbody>
</table>

5 Conclusion

This paper proposes a new topology for generating fast current pulses, which is based on a combination of the XRAM generator and flyback converter. The main advantage of the presented topology is the increased robustness by lowering the number of active switches to one and the fact that the switch can be placed at the surface. The proposed topology has a modular design and is therefore scalable. To achieve the required fast current rise time, a detailed analysis and optimisation of the parasitic components of the pulse transformer are required. The core material is selected with an algorithm to achieve a minimal solution regarding losses, inductance, and numbers of turns. The calculation is performed for two different core geometries, which lead to similar optimal results. In a second step, the parasitic components of three different winding arrangements are determined with a FEM simulation. The resulting parasitic values are feed into a PLECS model to attain the current rise time. The fastest achievable current rise time is below 500 ns for core nr. 2 and winding arrangement 2.

References