Optimal design of a modular 11kW series parallel resonant converter for a solid state 115-kV long pulse modulator

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Abstract

Modern accelerator driven experiments like linear colliders or spallation sources are supplied by RF amplifiers using klystrons. The cathode voltage for these klystrons can be generated by long pulse modulators generating highly accurate voltage pulses in the length of milliseconds. In this paper an optimization procedure based on an electrical and a thermal model of a series parallel resonant converter (SPRC) supplying these klystrons is presented. The efficiency of a basic SPRC-module is 95.3% with a pulsed power density of 6.64 kW/l.

I. INTRODUCTION

The SPRC topology (see Fig.1) is a modular design which needs no pulse transformer. Common designs like Bouncer Modulator topologies using pulse transformers become huge for long pulses. The series parallel resonant converter is a modular topology which avoids this drawback as the transformer is operated at high frequencies. The considered nominal pulse voltage amplitude is 115 kV with a pulse power of 2.88 MW and a pulse length of 2.8 ms. The pulse to pulse reproducibility of 0.02 % and a voltage ripple at top of less than 0.05 % are highly demanding. In order to meet these highly demanding specifications listed in Table I, the modulator is based on interleaved SPRC modules. A SPRC module [1] contains a full bridge connected to a series parallel circuit followed by a transformer a rectifier and a filter capacitor. In this paper, an optimal design of a single module according to the design considerations in [2] is presented. The basic modules of the SPRC can be connected in series or in parallel depending on the output power and ripple specifications and can also be interleaved. Due to the fact that the pulse shaping is achieved with a rectifier and a filter on the secondary side of the transformer, the pulse duration is relatively independent from the transformer size. Furthermore, the SPRC is naturally short circuit proof and provides zero voltage switching (ZVS). Additionally, the series inductance $L_S$ can be integrated by the leakage inductance of the transformer. Due to those advantages, the SPRC topology is investigated in the following. A straightforward design is difficult due to the high number of degrees of freedom and geometric parameters of the transformer. Thus, in this paper an optimization procedure based on [2] with given design constraints is shortly presented in section II. As all models are already commonly introduced in [2] section III only summarizes the electrical SPRC model and the thermal semiconductor model shortly, but put a detailed emphasis.

### Table I

<table>
<thead>
<tr>
<th>Pulse voltage $V_K$</th>
<th>115 kV</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pulse current $I_K$</td>
<td>25 A</td>
</tr>
<tr>
<td>Pulse power $P_K$</td>
<td>2.88 MW</td>
</tr>
<tr>
<td>Pulse repetition rate $P_{RR}$</td>
<td>20 Hz</td>
</tr>
<tr>
<td>Pulse width $T_P$</td>
<td>2.8 ms</td>
</tr>
<tr>
<td>Pulse duty cycle $D$</td>
<td>0.06</td>
</tr>
<tr>
<td>Pulse reproducibility $D_{REP}$</td>
<td>0.02 %</td>
</tr>
</tbody>
</table>
on the thermal transformer model as well on the leakage and isolation design procedures. Finally, section IV presents the optimization results.

II. OPTIMIZATION PROCEDURE

Based on the specifications of the pulse in Tab. I, Fig. 2 shows the optimal design procedure of the SPRC. In a first step, the optimizer varies the free parameters in the single module design with respect to the chosen design constraints in Tab. II. This single module design procedure consists of an electrical and a thermal model, and additionally of an insulation and leakage design procedure. The design of the single module is then verified with the constraints of the global specifications and this finally leads to the optimal number of single SPRC modules which are connected in series and/or in parallel.

### TABLE II

**Optimization parameters and constraints**

<table>
<thead>
<tr>
<th>Transformer core</th>
<th>Windings</th>
<th>Semiconductors</th>
</tr>
</thead>
<tbody>
<tr>
<td>l,t,h (see Tab. III)</td>
<td># of turns</td>
<td># of litz wire strands</td>
</tr>
</tbody>
</table>

**Constraints**

| Maximum magnetic flux density $B_{max}$ | 80 mT |
| Maximum winding temperatures $T_1, T_2$ | 120°C |
| Maximum temperatures in core parts $T_3, T_4$ | 120°C |
| Maximum change of junction temperature $\Delta T_J$ | 20°C |
| Defined leakage inductance $L_S$ (see Tab. III) | 5.1 µH |
| Maximum electrical field strength $E_{max}$ | 15 kV/mm |

III. SINGLE MODULE DESIGN MODELS

The following section shortly summarizes the electrical model of the single SPRC-module and the semiconductor thermal model. Afterwards the design procedures for the transformer insulation, leakage inductance and the thermal model of the transformer are given in detail.

A. ELECTRICAL MODEL

The electrical model of the SPRC basic module (see Fig.1b) is described with a first harmonic analysis [1], which also takes the output capacitor and rectifier into account. This model considers the transformer just as a voltage amplifier with turns ratio $n$ and determines all parameters of the resonant circuit ($C_S, C_P, L_S, n$) respectively delivers all voltages and currents.

B. SEMICONDUCTOR THERMAL MODEL

In this model, the resonant current calculated from the electrical model is used to determine the number of switches which leads to minimum losses with the constraint of a maximum $\Delta T_J$. The losses include conduction and switching losses. The conduction losses can be calculated directly from data sheet and the switching losses either are included from measurements or estimated also from data sheet. $\Delta T_J$ is used for lifetime estimations due to [3]. For IGBT modules, $\Delta T_J$ should not exceed 20°C under the constraint of a maximum junction temperature of 125°C to reach more than $10^8$ pulses. Some limits will be used for MOSFETs, what results in an optimal number of switches in parallel of 5. The investigated MOSFET is the STY139N65M5 from ST [4]. The required heat sink volume can be calculated with the Cooling System Performance Index (CSPI) as introduced in [5]. Fig.3 shows the built modulator prototype, each leg with 5 switches in parallel.

C. TRANSFORMER THERMAL MODEL

Based on the proposed transformer thermal model in [6], the model returns all critical temperatures (see Fig. 4). It contains all types of heat transfer represented in an equivalent thermal resistor and all losses are modelled...
by current sources. Most of the thermal resistors used in Fig.4, as presented in [2], are well described in literature [6], but there exists a lack concerning the thermal winding resistor $R_{th,Nx}$. A short explanation in the case of solid wire is given below, the full derivation of the thermal resistance of solid and litz wire can be found in [7]. Fig.5 shows a typical solid wire winding containing different heat transition resistors. First, a tangential part which represents the heat flow along the winding from layer to layer and second there are two different parts in radial direction. The tangential part can be easily calculated by

$$R_{th,tan} = \frac{i_W \cdot N_{pL}}{\lambda_{Cu} r_o^2 \pi k_L}$$

(1)

where $N_{pL}$ is the number of turns per layer and $i_W$ is the mean length per turn. In radial direction there exists an orthogonal part which is formed by both halves of the outer layers and an orthocyclic part which represents the heat transition in the inner winding layers. In orthogonal wire arrangements the wires in two neighbouring layers are laying directly side by side (see Fig.6). The thermal resistance then is [7]

$$R_{th,orth} = \frac{1}{\frac{2\lambda_{Air} l_W}{\alpha} \left[ V + \frac{1}{2} \left( \frac{2 \beta}{\alpha} \right)^2 \right]}$$

(2)

with

$$V = \arctan \left( \frac{\beta + 1}{\beta - 1} \right) \sqrt{\beta^2 - 1} - \frac{\pi}{4}$$

(3)

Fig. 3. Prototype modulator each leg with 5 switches in parallel, with two legs on one heat sink.

Fig. 4. Thermal equivalent circuit of the transformer.

Fig. 5. Cross section of a solid wire winding.

$$Z = \frac{\beta (\beta^2 - 2)}{(\beta^2 - 1)^{3/2}} \arctan \left( \frac{\beta + 1}{\beta - 1} \right) - \frac{\beta}{2} \frac{\pi}{4}$$

(4)

and

$$\alpha = 1 - \frac{\delta}{R_{th,Nso}/\lambda_{Air}} ; \quad \beta = \frac{1}{\alpha} \left( 1 + \frac{h_Z}{2 R_{th,Lay}/\lambda_{Air}} \right)$$

(5)

In orthocyclic windings the wires in the neighbouring layers are laying directly in the gap of the previous layer (see Fig.7). $R_{th,cyc}$ can be obtained as [7]

$$R_{th,cyc} = \frac{1}{4 \lambda_{Air} l_W \left( M_{Air} + M_{Iso} \left( \frac{4 \lambda_{Air} r_o^2}{\lambda_{Iso}} \right) \right) \left( r_o - \frac{\delta}{2} \right)}$$

(6)

with

$$M_{Air} = \int_0^\pi \cos^2 \psi - \cos \psi \sqrt{\cos^2 \psi - 0.75} - 0.5 \cos \psi - \alpha \sqrt{\cos^2 \psi - 0.75} + 0.5 \right)^2 d\psi$$

(7)

$$M_{Iso} = \int_0^\pi \sin^2 \psi + \cos \psi \sqrt{\cos^2 \psi - 0.75} - 0.75 \cos \psi - \alpha \sqrt{\cos^2 \psi - 0.75} + 0.5 \right)^2 d\psi$$

(8)

This finally leads to the general form of $R_{th,Nx}$

$$R_{th,Nx} = \left( R_{th,tan} || R_{th,cyc} \right) \frac{N_L - 1}{N_{pL}} + 1 \frac{N_{pL}}{N_{pL}}$$

(9)

Fig. 6. (a) Two orthogonal arranged wires with ideal assumed thermal heat flow lines. (b) Thermal heat flow between two wires simulated with COMSOL.
D. INSULATION DESIGN PROCEDURE

High electrical field strengths can harm the insulation of the transformer permanently and lead to arcs between the windings or the core. Therefore, a proper insulation design is unavoidable. Following assumptions are made for the insulation design. The borders of the winding window are grounded and the space between the windings and the rest of the winding window is completely filled with a homogenous isolation material. To calculate the electrical field inside the winding window the so called charge simulation method (CSM) (also called mirror charge method or image charge method) described in [8] is applied and is shortly explained in the case of one conductor. In Fig.8 (a) the conductor in the original window is represented by \( n \) image charges \( Q_j \) located inside the conductor on a radius \( d \) and belonging contour points on the radius \( r \) of the conductor each with the same conductor potential \( \Phi_C \). \( \Phi_C \) is formed by the superposition of these image charges in

\[
\Phi_C = \sum_{j=1}^{n} p_j \cdot Q_j \tag{10}
\]

where \( p_j \) are the potential coefficients which contain the geometric locations of the images charges. By mirroring the original window three times counterclockwise gives the original box depicted in Fig.8 (b). To build the basic box it would be also possible to mirroring around the original window but then the automated extension of mirrored basic boxes as seen in Fig.8 (c) is much more complicated. Applying (10) to all windows in all basic boxes and corresponding image charges leads to a system of linear equations which has to be solved for the unknown charges \( Q \)

\[
[Q] = [p]^{-1} \cdot [\Phi_C] \tag{11}
\]

Afterwards is it possible to calculate the electrical field strength \( E \) at every location \( x \) and \( y \) inside the original window.

\[
E = \sqrt{E_x^2 + E_y^2} \tag{12}
\]

with

\[
E_x = \sum_{j} \frac{Q_j}{2\pi \epsilon} (A1 + A2 + \ldots + Am) \tag{13}
\]

Fig. 7. (a) Three orthocyclic arranged wires with ideal assumed thermal heat flow lines. (b) Thermal heat flow between three wires simulated with COMSOL.

Fig. 8. Charge simulation method based on mirror charges. (a) Conductor represented by image charges, (b) basic box which contains the original window and three mirroring of it and (c) full set of used mirrorings.

\[
A1 = \frac{x-x_j}{(x-x_j)^2 + (y-y_j)^2} \left( x+x_j \right) + \frac{y+y_j}{(x-x_j)^2 + (y-y_j)^2} \tag{14}
\]

\[
B1 = \frac{y-y_j}{(x-x_j)^2 + (y-y_j)^2} \left( y+y_j \right) + \frac{y+y_j}{(x-x_j)^2 + (y-y_j)^2} \tag{17}
\]

\[
A2 = \frac{x-x_j-2X}{(x-x_j-2X)^2 + (y-y_j)^2} \left( x+x_j-2X \right) + \frac{y+y_j}{(x-x_j-2X)^2 + (y-y_j)^2} \tag{15}
\]

\[
B2 = \frac{y-y_j}{(x-x_j-2X)^2 + (y-y_j)^2} \left( y+y_j \right) + \frac{y+y_j}{(x-x_j-2X)^2 + (y-y_j)^2} \tag{18}
\]

where \( A1, B1 \) are the series of the positions of the image charges depending on the considered point \( x \) and \( y \) in the original box and \( A2 - Am, B2 - Bm \) are according to the shifted boxes (see Fig.8 (c) doted brown boxes). The variables \( x_j \) and \( y_j \) are the given positions of the image
Fig. 9. E-field distribution in grounded winding window.

charges whereas X and Y are the width and the length of the original window. For computational reasons $m$ is chosen 9 and the number of image charges is 16. This results in a quite high accuracy. Fig.9 shows the E-field distribution of the final optimized design with an average field of 5.75 kV/mm and a maximum field of lower than 15 kV/mm. Depicted design has been also compared to FEM and gives a deviation of lower than 8%.

E. LEAKAGE DESIGN PROCEDURE

Basically, the transformer is designed to minimum isolation spaces which leads to a given leakage inductance that is calculated with the current mirror method. Here, each conductor is represented by a single current in the original window (see Fig.10 (a)) which then is mirrored in the same way as described in the previous section (see Fig.10 (b) and (c)). In contrast to the mirror charge method where in a first step all unknown charges have to be calculated, the current is inherently given. Therefore

$$H = \sqrt{H_x^2 + H_y^2}$$

with

$$H_x = \sum_{j} \frac{I_j}{2\pi}(A1' + A2' + \ldots + Am')$$

$$H_y = \sum_{j} \frac{I_j}{2\pi}(B1' + B2' + \ldots + Bm')$$

The terms $Am$, $Bm$ and $Am'$, $Bm'$ are almost the same with the difference that all fractional terms have plus-signs. Finally to get the leakage inductance referred to the primary side

$$L_{leak,P1} = \frac{2W_m}{I_P}$$

with $W_m = \frac{l_W \mu}{2} \int \int H^2 dx dy$

where $I_P$ is the total primary current and $l_W$ the mean winding length. In order to get the final requested series inductance $L_S$ an auxiliary core as shown in Fig.11 delivers the missing leakage inductance $L_{leak,Aux}$ by adjusting the air gap.

$$L_S = L_{leak,P1} + L_{leak,Aux}$$

With this method influences concerning the leakage inductance due to manufacturing tolerances can be easily balanced.

IV. OPTIMIZATION RESULTS

Fig.12 and Tab.III summing the results of a single SPRC module optimization and the final number of modules of the overall system. The main losses are equally distributed
between transformer and switches whereas due to the much more complex structure of the transformer which results in more complicated cooling efforts, the volume of the transformer is quite higher than that of the modulator switches (see Fig.12 (a) and (b)). Comparing the given constraints in Tab.II with the results in Tab.III all values concerning temperatures, flux density and electrical field strengths fulfilling the limits. For the entire system 16 SPRC-modules are required.

![Comparison of losses and comparison of volume of a single SPRC-module.](image)

Fig. 12. (a) Comparison of losses and (b) comparison of volume of a single SPRC-module.

V. CONCLUSIONS

Due to a large number of degrees of freedom (e.g. leakage inductance and geometric parameters of the transformer), an optimal straightforward design for the resonant converter system is difficult. Therefore, in this paper an optimization procedure is presented, which is based on an electrical and a thermal model of the SPRC converter. Additionally, an insulation design procedure for the transformer and a thermal model for the switches is provided. With this approach an optimal design for minimum losses is achieved. The overall system consists of 2 SPRC modules connected in parallel and 8 of them in series. Each modulator module has 5 MOSFETs connected in parallel. The efficiency of a single SPRC-module is 95.3% with a pulsed power density of 6.64 kW/l.

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**References**


