EXPERIMENTAL VALIDATION OF A SERIAL PARALLEL RESONANT CONVERTER MODEL FOR A SOLID STATE 115-kV LONG PULSE MODULATOR

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Abstract—Medium and high beta cavities used in linear colliders or spallation sources are supplied by klystrons or inductive output tubes (IOT) amplifiers. The cathode voltage for these amplifiers can be generated by long pulse modulators generating highly accurate, high voltage pulses in the length of milliseconds. With existing modulator topologies all the demanding requirements like fast pulse rise time, high accuracy and low voltage ripple hardly can be satisfied at the same time. Common designs like bouncer modulator topologies using pulse transformers become huge for long pulses. The series parallel resonant converter (SPRC) avoids this drawback as the transformer is operated at high frequencies. In this paper, the comprehensive multi domain model of a SPRC converter including an electrical model of the inverter, a magnetic model, and an isolation design procedure of the transformer is verified with a prototype of a single module operated under full load conditions. In addition, a comparison between the predicted parasitics like leakage inductance and stray capacitance of the transformer and measured values is given. An evaluation of the isolation of the transformer, which is especially crucial for a series connection of the modules is also performed. Additionally, different possibilities to realize the desired series inductance are discussed.

I. INTRODUCTION

Based on the optimization procedure presented in [1] a single SPRC module has been designed. A single SPRC module [2] contains a full bridge connected to a series-parallel resonant circuit followed by a transformer, a rectifier, and a filter capacitor (see Fig.1). To obtain the required pulse voltage (see specifications in Tab. I) eight single SPRC modules have to be connected in series forming a stack. Two of these stacks are connected in parallel to achieve the required pulse power (see Fig.1). In this paper, the comprehensive multi domain model of a single SPRC module including an electrical model of the inverter, a magnetic model, and an isolation design procedure of the transformer are verified with a prototype of a single module operated under full load conditions. Section II shortly discusses the results of a single SPRC module with its optimized parameters including the resonant tank, the transformer and the output rectifier. In section III a comparison between the analytical approach which is used in the optimization procedure and FEM calculations of the leakage inductance and the stray capacitance is given and compared to measurements. Section IV presents the E-field conform isolation design of the transformer. Different possibilities to realize the desired series inductance (integrated or separately) are presented in section V. Finally, section VI presents pulse measurements of the prototype.

II. OPTIMIZATION RESULTS

Because of the high number of degrees of freedom during the design process as e.g. geometric parameters of the transformer or the design of the resonant tank, the optimization procedure presented in [1] has been developed for optimally designing the modulator. Tab. II summarizes the optimization results of the resonant tank and the transformer. It turned out that due to the high resonant current $I_{LS}$ the best way to realize the series inductance $L_s$ is an air toroid [1] plus the leakage inductance of the transformer. Therefore, it is necessary to accurately know the leakage of the transformer. An accuracy

![Fig. 1. Full system with 8 single SPRC modules in series forming a stack. Two of this stacks are connected in parallel.](image-url)
evaluation of the transformer parasitics calculations compared to measurements is given in the next section.

III. TRANSFORMER PARASITICS VALIDATION

A. Leakage inductance

As presented in [1] the calculation of the leakage inductance is based on the mirror current method. To increase the accuracy of this method different mirror planes are used inside and outside the core. Inside the core window mirroring on all 4 core walls is applied (see Fig. 2(b)) and for the region outside the core mirroring just to the left wall is used (see Fig. 2(c)). Therefore, (15) and (16) in [1] can be simplified in the left mirroring case and using (17) from [1] leads to the desired leakage inductance, where \( l_{W} \) is the integration length inside the core \( l_{W,\text{in}} \) and outside the core \( l_{W,\text{out}} \) (see Fig. 2(a)). Finally, with this method it is possible to calculate the magnetic energy \( W_{m} \) inside the transformer window (see Fig. 2(b)), in the air box outside the transformers’ front (see Fig. 2(a)), but it is not possible to consider the energy above and under the transformer core. It is assumed that there is a negligible amount of energy inside the core due to the high permeability of the core material. Table III compares the error between FEM and the mirror current method related to the measured leakage inductance by varying the parameter \( n_1 \). If the areas in the core, above and below the transformer are also considered in the FEM evaluation the leakage inductance results in 1.59\( \mu \)H and leads to an error of \(-0.94\%\) related to the measured value.

B. Stray capacitance

As presented in [1] the calculation of the stray capacitance is based on the mirror charge method, with mirroring inside the core at all 4 walls and separately mirroring at the left wall outside the transformer. A comparison is given between 2D FEM, the mirror method and measurements in Tab.IV. No values are given for 3D FEM because little deviations of the turns arrangement or the radiiuses of the winding edges in the 3D model lead to high deviations in the electrical field, respectively in the capacitance.

IV. INSULATION DESIGN PROCEDURE

The following section describes an insulation post design check which is performed after the optimization procedure. It is not possible to integrate a full analytical model of the transformer in all details e.g. bobbins and winding fastenings (see Fig. 3) in the procedure. A basic insulation design is included in the optimization procedure by calculating the maximum electrical field and varying the distances between primary and secondary [1].

<table>
<thead>
<tr>
<th>2D FEM</th>
<th>Mirrored</th>
<th>Measured</th>
<th>error FEM</th>
<th>error mirror</th>
</tr>
</thead>
<tbody>
<tr>
<td>45.9</td>
<td>46.5</td>
<td>59.9</td>
<td>-23.4</td>
<td>-22.4</td>
</tr>
</tbody>
</table>

Fig. 2. Integration areas and paths for 3D, 2D FEM and 2D numerical parasitics calculations of the transformer.
times a proper insulation design is necessary and a detailed analysis of the electrical field distribution along creepage paths and long oil paths inside the transformer was carried out with the help of the Weidmann design curve method [4]. This design method is based on oil design curves which are derived from homogenous electrical breakdown tests [5]. These are then compared with the averaged cumulated electrical field strength $E_{\text{avg,cum}}$ along certain paths. Fig. 3 shows in (a) an evaluated oil gap $l_1$ and a creepage path $l_f$. For a valid design $E_{\text{avg,cum}}$ both paths have to be below the design curve in (b). The design curves of the used transformer oil MIDEL 7131 [6] are derived from [7]. With this method, insulation designs with homogenous as well as inhomogenous field distributions can be investigated. Finally, applying this method leads to an electrical field conform design, which means that the potential lines just have a tangential component along insulation fastenings (see Fig. 3(a)). Hence, the insulator is stressed in normal direction by the electrical field and has its maximum electrical strength. Fig. 4(a) shows the built transformer prototype without any mountings between primary and secondary bobbins inside the transformer to avoid creepage paths between them as depicted in Fig. 4(b). The bobbins are fixed outside the transformer, leading to a longer creepage distance between the windings and little deviations of the arrangement between reality and 3D model would lead to high deviations in the electrical field

V. ALTERNATIVE WAYS TO REALIZE THE SERIES INDUCTANCE

The series inductance in the optimization procedure is modelled as an air toroid. Alternative ways to realize the series inductance and a comparison of them by volume and losses is shown next. The following discussion is based on the results in Tab.II and all transformers are designed to the same isolation distances and turn ratios. A possible way to realize the series inductance $L_S$ is to integrate it completely with the transformer leakage inductance (see Fig. 5 (a)). By adapting $\Delta x$ and $\Delta y$ in the case of the U-core transformer or only $\Delta y$ in the case of the E-core transformer, the leakage inductance can be modified. This leads to a high boxed volume. Another possibility is to use an additional stray core wound by the primary or the secondary winding [8] (see Fig. 5(b)). The inductance is set by varying the air gap of the stray core. If it is possible to use all turns of the secondary winding the isolation distance $x_{\text{min}}$ has a minimum. Otherwise, if not all turns of the secondary could be used to keep the air gap as short as possible two times of the insulation distances have to be added to $x_{\text{min}}$ and increases the volume. Additionally, using a stray core leads to higher losses due to the high resonant current $I_{L_s}$. The third option to realize the series inductance $L_S$ is an air toroid (see Fig. 5(c)). The transformer is designed with respect to minimum insulation distances and the air toroid causes no core losses. Thus, using a transformer and an air
toroid it is the best compromise between volume and losses to realize the series inductance. In Tab. V a comparison of the discussed possibilities between volume and losses is listed.

**TABLE V**

<table>
<thead>
<tr>
<th></th>
<th>Losses [W]</th>
<th>Volume [l]</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a) U-Core</td>
<td>82</td>
<td>48</td>
</tr>
<tr>
<td>(a) E-Core</td>
<td>66.1</td>
<td>46</td>
</tr>
<tr>
<td>(b)</td>
<td>361.7</td>
<td>25.1</td>
</tr>
<tr>
<td>(c)</td>
<td>97.3</td>
<td>12.8</td>
</tr>
</tbody>
</table>

VI. MEASUREMENTS

Finally, this section presents the measurement results. The green curve $V_{\text{Out,meas}}(t)$ in Fig. 6 depicts the measured output voltage pulse $V_{O11}$ of a single SPRC module. The simulated blue curve $V_{\text{Out,sim}}(t)$ shows good accordance with the measured one and the red curve $V_{\text{Out,avg}}(t)$ is the mean average of $V_{\text{Out,meas}}(t)$, used to determine the rise and the fall time. Rise and fall times are well below the given limits in Tab.I. Just a $300\mu$s pulse is shown in Fig. 6 to get the pulse rise and fall time into one picture.

VII. CONCLUSION

In this paper, the comprehensive multi domain model of the SPRC converter is verified with a prototype of a single module by measurements. The simulated data shows good accordance with the measured one. In addition, a comparison between the predicted parasitics like leakage inductance and stray capacitance of the transformer and the measured ones is given. The error between calculated and measured parasitics is low. An evaluation of the isolation of the transformer is also carried out. Additionally, different possibilities to realize the desired series inductance are discussed and compared by volume and losses.

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