Optimal Transformer Design for Ultraprecise Solid State Modulators

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Abstract—In this paper, a procedure for optimal transformer design with a variable set of constraints for pulse transformers with a pulse range of 3–140 µs is presented. During the optimization procedure, the pulse shape is analyzed in the time domain, ensuring that the pulse constraints, such as rise time and overshoot are met. For accurate prediction of the pulse shape, analytical approaches are proposed to estimate the distributed capacitance and leakage inductance of the transformer. The analytical approach is verified by 2-D-FEM simulations and measurements. The optimization procedure considers pulse, core, winding, demagnetization losses, and losses of the primary switches. First, the procedure is applied to an existing pulse transformer with specifications for SwissFEL. An improvement of 16.6% in conversion efficiency is achieved in comparison with the existing design. In a second step, the procedure is applied to specifications of the compact linear collider, which demands high conversion efficiency. The resulting optimal transformer consists of three cores with five primary turns and requires a tank volume of 0.915 m³. In an optimal configuration, an overall conversion efficiency of 97.7% is achieved for the considered system including pulse losses.

Index Terms—Core losses, distributed capacitance, high voltage, leakage inductance, optimization procedure, pulse transformer, transformer design.

I. INTRODUCTION

PULSE power converter systems are used in a wide range of applications and have to comply with various system requirements, such as output power, flat-top length, and pulse repetition. In addition, the requirements for the pulse shape, rise time, pulse repetition accuracy, flat-top stability (FTS), and overshoot are highly application dependent as can be observed in two different sets of selected requirements listed in Table I, one for a short-pulse system, the SwissFEL, and one for a long-pulse system, the compact linear collider (CLIC).

In general, pulse power conversion systems can be divided into four main groups: 1) line-type modulators, such as pulse-forming networks [1]; 2) transformer-free systems based on a single high voltage switch or on the Marx-generator principle [2]; 3) concepts based on pulse transformers [3], often realized with solid state switches [4]; and 4) resonant topologies for pulses in the millisecond range [5], [6].

This paper focuses on systems with a pulse transformer and solid state switches for a pulse range 3–140 µs. This technology provides high reliability due to its simple structure and reliable short-circuit protection.

Solid state modulators with matrix transformers have been investigated in [7]–[9].

Due to the varying specifications, the pulse power conversion system must be adapted for its individual purpose. To ensure the optimal conversion system for every specification set, optimization procedures can be applied. To allow a reasonable overall computing time, one optimization cycle should not exceed 1 s. Thus, analytical approximations are used to calculate the transformer parasites. Analytical approximations have already been conducted in [4], but they are not applicable for arbitrary geometry.

Therefore, improved analytical approximations are proposed in this paper, which are suitable for a broader range of transformer geometries. They are then integrated in an optimization procedure maximizing the efficiency of the pulse transformer under the constraints of electrical and pulse requirements including switching unit.

At first, an overview of the investigated system is given in Section II. Subsequently, the optimization procedure is presented in Section III, which analyzes the pulse shape in the time domain and controls the pulse specifications. In addition, the algorithm contains several models estimating pulse, core, winding, active reset circuit losses, and losses of the primary switches.

Thereafter, in Section IV, an analytical approach for estimation of the leakage inductance and the stray capacitance is proposed, which is suitable for a variable pulse transformer geometry. Both the approaches are compared with 2-D-FEM simulations and in addition validated with a measurement of an existing pulse transformer.

TABLE I

<table>
<thead>
<tr>
<th>Specifications for Two Different Modulator Systems</th>
<th>SwissFEL</th>
<th>CLIC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Flat-top length</td>
<td>3</td>
<td>140 µs</td>
</tr>
<tr>
<td>Output voltage</td>
<td>370</td>
<td>150 kV</td>
</tr>
<tr>
<td>Output power</td>
<td>120</td>
<td>24 MW</td>
</tr>
<tr>
<td>Rise &amp; fall times</td>
<td>1</td>
<td>3 µs</td>
</tr>
<tr>
<td>Pulse repetition</td>
<td>100</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Primary circuit DC-link voltage</td>
<td>3</td>
<td>3 kV</td>
</tr>
<tr>
<td>Voltage overshoot</td>
<td>-</td>
<td>1 %</td>
</tr>
<tr>
<td>Flat-top stability FTS</td>
<td>&lt;1</td>
<td>0.85 %</td>
</tr>
<tr>
<td>Modulator global efficiency</td>
<td>-</td>
<td>90 %</td>
</tr>
</tbody>
</table>

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Finally, in Section V, the optimization procedure is conducted for specifications of SwissFEL. The resulting pulse shape is compared with the pulse shape with parameters derived from an existing pulse transformer with equal specifications. In a second step, the procedure is applied for the specifications of CLIC and the results are presented.

II. OVERVIEW OF INVESTIGATED SYSTEM

The investigated system is a matrix transformer [8], as shown in a simplified schematics in Fig. 1. In a matrix transformer, the secondary winding encloses all the cores leading to a turns ratio \( n \), which is defined by

\[
n = \frac{v_s}{v_p} n_c
\]

where \( n_c \) is the number of cores, \( v_p \) the primary voltage, and \( v_s \) the secondary voltage. A cone winding arrangement is chosen, because it outperforms a parallel winding arrangement in terms of leakage inductance [10]. Each core has two legs leading to two identical transmission systems, which are connected in parallel. The number of cores is variable. For each core leg, a switching unit, composed of the main capacitor bank \( C_m \) and the main switch \( S_m \), is considered. Additionally, an active reset circuit per core leg is considered (\( C_r \), \( S_r \), and \( D_r \)), which will be described further in Section III-F. Due to a high ratio of \( v_s/v_p \), only the primary leakage inductance \( L_{\sigma 1} \), due to the loop comprising main capacitor and switching unit, the primary copper resistance \( R_{11} \) of this loop and the secondary capacitance \( C_{sec} \) are considered for this transformer setup. Additionally, the leakage inductance \( L_\sigma \) and the distributed capacitance \( C_d \) of the transformer are considered as well as the nonlinear klystron load \( R_{klys} \).

III. OPTIMIZATION PROCEDURE

The aim of the proposed optimization procedure is to provide a transformer design, which suits the selected requirements. This could be a limited overshoot, a limited rise time, a fixed number of primary windings, or other certain geometry constraints. As shown in Table I, the CLIC system has demanding specifications regarding modulator global efficiency, from grid to the klystron load. Therefore, in this paper, the optimization procedure is not only designed to meet the pulse requirements but also to find a solution with highest system efficiency.

In the following sections, the optimization procedure is described briefly, followed by the detailed description of each loss model.

A. Structure of the Optimization Procedure

The optimization procedure consists of several models, as shown in Fig. 2. The global optimizer runs with a set of six optimization variables: number of primary turns \( n_p \), number of cores \( n_c \), secondary winding height \( h_s \), distance between the secondary and primary windings \( d_{\omega m} \), width of the core cross-sectional area \( w_{\omega A} \), and opening angle of the cone \( \alpha \).

In addition, a set of fixed parameters are given such as constants or predefined geometric distances, e.g., the distance between the core and tank. These parameters and the optimization variables define the geometry of the transformer and its surrounding tank. Once the geometry is determined, its parameters are used to calculate all required components for the analysis of the pulse shape. Initially, the distributed capacitance \( C_d \) and the leakage inductance \( L_\sigma \) are calculated, as explained in Section IV. Second, the winding losses are computed and the ohmic copper resistance \( R_{11} \) of the primary side is transferred to the pulse shape analysis. Then, the core losses are calculated and the ohmic resistance \( R_{Fe} \) is
defined as well as the magnetizing inductance $L_h$. With $L_h$, the magnetizing current at the beginning of the pulse $I_{\text{demag}}$ is derived.

With all these parameters, the pulse shape is analyzed in the time domain. During this step, the optimizer compares the shape with a set of external constraints such as rise time or overshoot. If all the constraints are met, the pulse losses (as described in Section III-D.2) are calculated. The magnetizing current $I_{\text{demag}}$ at the end of the pulse is fed back to the active bias loss model, which then calculates the demagnetization losses. Finally, losses of each model are combined to obtain the total losses of the given parameter set.

At the end of the optimization procedure, the algorithm supplies a transformer configuration, where all demanded pulse requirements are met and highest efficiency is achieved.

**B. Geometric Transformer Setup**

In the first step of the optimization procedure, the geometry of the transformer depending on the set of optimization parameters is defined as a basis for the analytical calculations. The correlation of geometric parameters is shown in Fig. 3. The entire geometry is defined by a few parameters to build it as compact as possible while avoiding field enhancement. Therefore, the minimal distance $d_{w,\text{min}}$ between the conductor with the highest voltage potential of the secondary and the primary windings must be derived. The same distance is applied between this conductor and the grounded core. The conductor with the highest secondary voltage potential is usually realized as a field shape ring with radius $r_r$. To estimate the resulting electrical peak field, the geometry is approximated as a cylindrical capacitor (Fig. 3), where the highest electrical field occurs at the inner radius $r_r$ and is defined by

$$E(r_r) = \frac{V_s}{r_r} \frac{1}{\ln \left( \frac{d_w}{r_r} \right)}$$

where $V_s$ is the applied secondary voltage, $d_w$ the outer, and $r_r$ the inner radius of the cylinder. Therefore, the minimal allowed distance $d_{w,\text{min}}$ between the windings can be calculated as

$$d_{w,\text{min}} = r_r \exp \left( \frac{V_s}{r_r E_{\text{peak}}} \right)$$

where $E_{\text{peak}}$ is the maximal tolerable peak electric field in oil. This peak field occurs only at the surface of the field shape ring. An average electrical field between the windings $E_{\text{av}}$, when a homogenous field is assumed, is defined by

$$E_{\text{av}} = \frac{(V_s - u_p)}{(d_w - rr)}.$$

Calculating the average electrical field $E_{\text{av}}$ of the analyzed transformer geometries would result in much lower values than the peak electric field $E_{\text{peak}}$, which is shown in Figs. 10 and 11. Because the electrical field in case of the analyzed geometries is inhomogeneous, only the peak electric field is considered and set as constraint. A value of $E_{\text{peak}} = 20 \text{kV/mm}$ is assumed for short pulses of a few microseconds and for longer pulses in the 100 μs range the value is set to $E_{\text{peak}} = 12 \text{kV/mm}$.

The cross-sectional area of each single core can be defined by

$$A_c = \frac{v_p l_{\text{flat}}}{2 B_{\text{max}} n_p F_c}$$

where $F_c$ is the filling factor of the cut tape-wound core and $B_{\text{max}}$ the desired flux density.

The other geometry parameters, shown in Fig. 3, are

$$b_c = 2r_r + 2d_w + 2d_p + 2d_{\text{prim}}$$

$$h_{s,\text{min}} = n_s d_{s,\text{min}} \cos(\alpha_{\text{max}})$$

$$h_c = d_w + h_s + d_{\text{top}}$$

$$d_{A_c} = \frac{A_c}{w_{A_c}}$$

The area outside the core is equally structured as the core window, except that the distance between the field shape ring and oil tank is enlarged by $d_{\text{add}}$, leading to smaller capacitance values of the transformer, but to higher tank volume. Therefore, the value of $d_{\text{add}}$ is set externally.

1) Transformer Cooling: The transformer is considered to be in a tank filled with standard transformer oil, to allow a more compact design. The cooling is realized by grounded cooling pipes. The positioning of these pipes is important, as they contribute to the distributed capacitance of the transformer. To minimize their influence, they are positioned at the level of the lower voltage turns. To allow an effective cooling, the transformer is placed upside down, as shown in Fig. 3 into the tank. Consequently, the cooling pipes are positioned at the top close to the tank wall, where the heated oil flows due to convection. Even though the ground plate needs to be enforced to carry the transformer weight, this configuration is chosen as it has the additional advantage of wiring being led through the top, thereby avoiding oil leakage.
C. Winding Losses

The winding losses, due to skin and proximity effects were subjected in a number of publications, e.g., [11]. To estimate the skin depth in the conductor, the pulse current is approximated as a trapezoid and a FFT analysis is conducted. For calculating the proximity effect losses, the secondary circular windings are transformed to a sheet conductor, as shown in [11]. A simplified geometry is assumed, in which the windings cover the entire height of the core window.

D. Pulse Shape Analysis

For the design of a pulse transformer, it is crucial to predict the resulting pulse shape. Therefore, in this paper, the time-domain circuit model for pulse prediction is introduced, followed by the pulse constraint implementation and description of the algorithm’s penalty arrangement.

1) Time Domain Circuit Model: Because the klystron load $R_{\text{kly}}$ is nonlinear and influences the pulse shape significantly [10], the pulse shape is evaluated in the time domain. The applied circuit model is derived from [12], based on the secondary voltage potential and shown in Fig. 4.

This circuit model includes the mentioned parameters $C_d$, $L_d$, $L_h$, $R_{\text{Fe}}$, $R_{\text{kly}}$, $L_{\sigma 1}$, and $L_{\sigma 2}$. On the secondary side, a capacitance $C_{\text{sec}} = 100 \, \text{pF}$, due to klystron load and measurement equipment, is considered.

Due to the time domain analysis, the model is able to cope with any given nonlinear load function. However, for the klystron the load function is defined by

$$i_k(t) = kv_k^{1.5}(t) \quad \text{and} \quad k = \frac{i_{k,\text{nom}}}{V_k^{1.5}} \quad (10)$$

where $k$ is the perveance, $i_{k,\text{nom}}$ is the nominal, $i_k(t)$ and $v_k(t)$ are the time-dependent values of current and voltage of the klystron.

To consider the rise and fall times of the primary switches as well as the voltage drop in the primary capacitances $\Delta v_{\text{cap}}$ during flat top, the voltage signal $v_1(t)$ is applied as a time-dependent function. This circuit model is not applicable during demagnetization of the transformer. However, it is a valid assumption that the energy stored in the transformer can be retrieved by the active bias circuit, except losses occurring in its components [4]. Therefore, losses during the pre- and demagnetization periods are considered in the active bias circuit loss model.

2) Pulse Losses: Because the klystron load cannot be initiated until the voltage complies with the FTS criteria, a part of the transferred energy of the pulse is lost. In [13], the ratio between the ideal and real pulses has been defined as pulse efficiency. Since in this paper, the pulse shape is analyzed in the time domain, the energy lost before the klystron can be activated and the energy lost during the fall time of the pulse can be computed. These losses will be referred to as pulse losses in the following.

3) Implementation of the Pulse Constraints: During the optimization procedure, the pulse constraints are surveyed. If the pulse exceeds a constraint, a penalty will be set.

The pulse behavior is analyzed in the time domain with a nonlinear differential equation system, which is solved analytically. A possible pulse shape with constraints is shown in Fig. 5.

The resulting pulse signal is compared in every time step with the predefined overshoot voltage and the voltage band. If the signal crosses the overshoot voltage boundary, the calculation will be terminated. As long as the signal exceeds the upper voltage or falls below the lower voltage of the voltage band, the time vector $t_{\text{ref}}$ is increased. Thus $t_{\text{ref}}$ indicates the beginning of the flat-top period $t_{\text{flat}}$. To minimize the computation time, the analysis is only conducted for the period of

$$t_{\text{controll}} = t_{\text{rise}} + t_{\text{settle}} + t_{\text{add}} \quad (11)$$

where $t_{\text{rise}}$ is the rise time, $t_{\text{settle}}$ the settling time and $t_{\text{add}} = 2 \, \mu\text{s}$. If $t_{\text{ref}}$ is higher than $t_{\text{rise}} + t_{\text{settle}}$, a penalty will be set. This will also be the case, if $t_{\text{rise}}$, which is derived from the pulse signal, is exceeded.

To ensure that the algorithm converges to the predefined design space, the penalties increase with their difference from the set point value.

E. Core Losses

The considered material for the pulse transformer is Metglas (amorphous alloy 2605SA1) because it offers a good compromise between the low losses, high saturation flux density, and costs. For a rectangular voltage, resulting in different gradients of the magnetic flux density $dB/dt$, losses can be predicted by the improved generalized Steinmetz equation (GSE) [14]. In case of a pulse transformer with active bias circuit, which allows to double the magnetic flux density swing, the energy
loss per pulse $E_o$ is described by

$$E_o = k_1 A_{c,eff} l_{eff} (2 B_{max})^{\beta - \alpha} \left( t_{premag} \left( 2 B_{max} \right)^{\alpha} + t_{flat} \left( 2 B_{max} \right)^{\alpha} + t_{demag} \left( 2 B_{max} \right)^{\alpha} \right)$$

where $\alpha$, $\beta$, and $k_1$ are the Steinmetz parameters, $A_{c,eff}$ is the effective cross-sectional area, $l_{eff}$ is the effective magnetic length, $t_{premag}$ and $t_{demag}$ are the pre- and demagnetization time, $t_{flat}$ is the flat-top length, and $B_{max}$ is the desired magnetic flux density, which was set for the design procedure to $B_{max} = 1.2 \ T$.

Steinmetz parameters are based on curve fitting and the considered pulse range from 3 to 140 $\mu$s is wide. Therefore, parameter sets for three different ranges, as shown in Table II are derived from the measurements. The test core is an AMCC 367S core, which has an effective cross-sectional area of 5.29 cm² and a magnetic length of 43.78 cm.

G. Losses of Main Switches

Because of high reverse voltage and high current ratings are required, there are only insulated gate bipolar transistors (IGBTs) as main switches considered. The loss energy $E_{IGBT}$ of an IGBT is defined by

$$E_{IGBT} = V_{CE,sat} I_{prim} t_{flat} + E_{on} + E_{off}$$

where $E_{on}$ is the energy loss during turn-on, $E_{off}$ during turn-off and $V_{CE,sat}$ the saturation collector emitter voltage. Typical values from datasheet are assumed [16].

IV. CALCULATION OF PARASITICS

In Section III-D, it is pointed out that for analyzing the pulse shape, the distributed capacitance $C_d$ and the stray inductance $L_s$ are critical parameters. In [4], simplified analytical equations were proposed for both parameters. These equations are, however, only valid for certain geometries and contain assumptions such as a centered field shape ring between the transformer and oil tank. Therefore, these are not suitable for a flexible geometry in an optimization procedure.

Consequently, in this section analytical methods are proposed to calculate the distributed capacitance and the leakage inductance for a matrix transformer with flexible geometry. The only assumption made is that the core as well as the primary and secondary windings is grounded at one end. This is the case for most of the pulse transformers to obtain a defined potential between the windings [4].
First, the method for calculating $C_d$ is introduced, followed by the calculation method of $L_d$. Second, these methods are compared with 2-D finite element method (FEM) simulations. In addition, the simulations are verified with measurements of an existing pulse transformer.

### A. Calculation of Distributed Capacitance

To estimate the distributed capacitance $C_d$ of a pulse transformer, the geometry is analyzed in the 2-D space to obtain the capacitance per length $C'_d$, which is then multiplied with its associated length.

The $n_s$ secondary turns are considered as line conductors with a voltage potential $v_k$ of the $k$th conductor of

$$v_k = v_s \frac{k - 1}{n_s - 1}, \quad 1 \leq k \leq n_s$$  \hspace{1cm} (15)

where $v_s$ is the secondary voltage potential of the transformer. The primary windings are realized as foil conductors and are considered for the analytical calculation as entirely grounded. This simplification leads to a slightly higher capacitance, but the error is negligible (e.g., for Fig. 10 smaller than 2%) because

$$V_s / V_p \gg 1.$$  \hspace{1cm} (16)

In a multiconductor system in 2-D space, which is shown in Fig. 8(a) for four conductors, the relation between the potential $\Phi'$ and charge $Q$ is described by

$$[Q] = [p]^{-1} \cdot [\Phi'] = [c] \cdot [\Phi'].$$  \hspace{1cm} (17)

In case of the pulse transformer, the conductors are surrounded by the grounded core and the oil tank. The influence of these surfaces is considered applying the charge simulation method [17]. Each conductor is mirrored in each direction to consider the influence of the grounded surfaces. To describe the resulting electrical field correctly, the mirror charges have alternately positive and negative signs, as shown in Fig. 8(b).

In a 2-D space, the potential coefficient $p_{ij}$ between the conductor $i$ and $j$ can be described based on the superposition principle [18] as

$$p_{ij} = \frac{1}{2 \pi \epsilon_0 \epsilon_r} \ln \left( \frac{r_{ij}}{r_{i,mn}^{'} r_{j,mn}^{''}} \right) + \sum_{m p j=mn j=1}^{N/2} \ln \left( \frac{r_{i,m p j}^{'} r_{j,mn j}^{''}}{r_{i,mn j}^{'} r_{j,mn j}^{''}} \right)$$  \hspace{1cm} (18)

where $r_{ij}$ is the distance between the two conductors, $r_{i,m p j}^{'}$ is the distance of conductor $i$ to all positive mirror charges and $r_{i,mn j}^{''}$ to all negative mirror charges of conductor $j$.

In case of $i = j$, the potential coefficient is obtained by

$$p_{ii} = \frac{1}{2 \pi \epsilon_0 \epsilon_r} \ln \left( \frac{r_r}{r_{i,m p i}^{'} r_{i,mn i}^{''}} \right) + \sum_{m p i=mn i=1}^{N/2} \ln \left( \frac{r_{i,m p i}^{'} r_{i,mn i}^{''}}{r_{i,mn i}^{'} r_{i,mn i}^{''}} \right)$$  \hspace{1cm} (19)

where $r_r$ is the radius of the conductor, $r_{i,m p i}^{'}$ the distance of conductor $i$ to all its positive mirror charges and $r_{i,mn i}^{''}$ to all its negative mirror charges.

To obtain the partial capacitances between the conductors, the potential coefficient matrix $[p]$ has to be inverted to obtain the capacitance coefficients $[c]$. The partial capacitance per unit length between the two conductors $C_{ij}'$ and the capacitance per unit length to the surrounding potential $C_{i,\infty}'$, can be obtained by

$$C_{ij}' = -c_{ij} \quad \text{and} \quad C_{i,\infty}' = \sum_{j=1}^{n} c_{ij}.$$  \hspace{1cm} (20)

The total distributed capacitance per unit length $C_d'$ based on the secondary voltage potential $v_s$ is derived from the total stored electric energy $W_{el,\text{tot}}'$ of the geometry, which is calculated by summing up the electric energy of each partial capacitance:

$$W_{el,\text{tot}}' = \sum_{j=1}^{n} 0.5 C_{i,j}' (v_i - v_j)^2 \quad \text{and} \quad C_d' = \frac{2W_{el,\text{tot}}'}{v_s^2}.$$  \hspace{1cm} (21)

### B. Calculation of Leakage Inductance

The leakage inductance $L_d$ is derived by multiplying of the inductance per unit length $L_d'$ in the 2-D space with its associated length. The approach used in this paper is based on a multiconductor system. The geometry is simplified, as shown in Fig. 9, where the primary winding is approximated by $n_s$ circular conductors, equally distributed over the primary winding height $h_p$. The field shape ring is approximated by a smaller circular conductor. The secondary winding conductors...
are then rearranged, to keep the secondary winding height\( h_s \) constant.

In a multi-conductor system, the magnetic flux per length is defined by

\[
\Phi_j' = \sum_j L_{ij}' \cdot i_j. \tag{22}
\]

Due to the 2-D analysis, it is not of interest how the conductors are connected as turns [19]. Therefore, one primary and one secondary turn can be combined to a double circuit line. If the radius \( r_{rr} \) is small compared with the distance \( d_1 \) between the two conductors, the self-inductance per unit length \( L_{11}' \) of a double circuit line can be defined by [20]

\[
L_{11}' = \frac{2\Phi_1'}{i_1} = \frac{\mu_0}{\pi} \ln \left( \frac{d_1}{r_{rr}} \right). \tag{23}
\]

The mutual inductance per unit length, which is caused by current \( i_1 \) of circuit line \( l_1 \) in \( l_2 \), as shown in Fig. 9(b) can be described by [20]

\[
M' = L_{21}' = \frac{\Phi_{21}'}{i_1} = \frac{\mu_0}{2\pi} \ln \left( \frac{r_{12}r_{21}}{r_{11}r_{22}} \right), \tag{24}
\]

Because in the chosen arrangement it is assumed that all the conductors carry an equal current, the total inductance per unit length of the geometry can be obtained by

\[
L' = \sum_i \sum_j L_{ij}' . \tag{25}
\]

The core has a very high permeability \( \mu_r \) and therefore can be observed as a magnetic mirror, where the conductors are mirrored, as shown in [15]. For fast computation, only eight mirrored conductors are considered for each turn. Outside the core, the conductors are only mirrored in one direction at the axis of the core, because the surrounding oil tank has a lower permeability and the distance to the windings is much higher than between the windings and core.

**C. Validation by 2-D FEM Simulations**

To validate the analytical approaches of \( C_d' \) and \( L_{\sigma}' \), the results are compared with a 2-D-FEM simulation. In Fig. 10, a matrix transformer geometry for long pulses and in Fig. 11 a geometry for short pulses is shown, each with the resulting magnetic and electric field. 2-D-FEM simulations of these two different geometries are compared in Table III with \( C_d' \) and \( L_{\sigma}' \) for the core window and the space between the core and tank. It can be observed that the difference between the 2-D-FEM simulations and calculation is smaller than 6%. This indicates that although simplifications are applied, the accuracy is still high. For further validation, analytical calculation results are compared with the measurement data in the following section.

**D. Verification With Measurements**

To validate the proposed analytical calculations of \( C_d \) and \( L_{\sigma} \), the calculations are compared with measurements of an existing pulse transformer. The pulse transformer is realized for the specifications in Table I(b) and corresponds to Figs. 11 and 13.

At first, it is described in detail how \( C_d \) and \( L_{\sigma} \) of the pulse transformer are obtained from analytical calculations, followed by a comparison of the approach with 2-D FEM and measurements.

1) **Parasitics in Dependence on Geometry:** Because the distances between the core and tank wall differ at the front, the rear, and the side of the pulse transformer, the distributed capacitance per unit length \( C_d' \) is calculated for three different regions and the core window, which are shown in Fig. 13. These values are then multiplied with the average length of the corresponding region.

The stray inductance does not depend on the distance between the core and tank wall, because almost of the magnetic energy is comprised in the area between the primary and secondary windings. Therefore, only two different regions are considered: 1) inside the core window and 2) outside of the core. As an average length of the second region, the
central point between the primary and secondary windings is multiplied with the corresponding length, which is indicated with $l_{L/\sigma}'$.

The results for $C_d$ and $L_\sigma$ in dependence on the area are shown in Table IV.

2) Comparison: To validate the proposed analytical calculations further to pulse measurements of an existing pulse transformer were performed.

In a first measurement, the primary inductance of a switching unit is determined to be $L_{11} = 90 \text{nH}$, which can be transferred to the secondary side with respect to the number of cores and the turn ratio to $L_{11}' = 119 \mu\text{H}$.

The transformer is then measured with a resistive load of $R_l = 1000 \Omega$ at a pulse voltage of $V_s = 189 \text{kV}$. The measured pulse shape is then compared with a second-order delay element, which can be used to describe the pulse shape at the beginning of the pulse [4]. The second-order delay element considers the calculated parasitics and in addition the rise time of the switches of $T_r = 200 \text{ns}$.

In Fig. 12, the measurement data are compared with the pulse shape with results of 2-D-FEM analysis and calculation. Both results show high correspondence to measurements and to each other.

V. APPLICATION OF THE PROPOSED OPTIMIZATION PROCEDURE

A. Specifications of SwissFEL

The proposed optimization procedure is conducted with specifications of the SwissFEL (Table I). At first, the pulse shape is predicted using the geometry parameters of the investigated pulse transformer, shown in Fig. 13. The parasitics are taken from Section IV-D.2, assuming a klystron load and an additional secondary capacitance of $C_{sec} = 50 \text{pF}$.

In a second step, the algorithm is conducted with equal assumptions, but with unconstrained optimization parameters. Only two optimization parameters, $n_c$ and $w_{Ac}$ are preset: Because of the high required power of 120 MW at least 12 pulse switches are required. Therefore, $n_c = 6$ is used in the algorithm, since there are two switching units per core. The minimum core width is preset to have an equal width as the used IGBT modules $w_{Ac} = 140 \text{mm}$, which is also the case for the existing transformer, to allow a direct connection of the modules and to minimize the primary inductances.

The results of the comparison are shown in Table V. The first column lists the results when the optimization parameters correspond to the investigated pulse transformer. In the second
Fig. 14. Section of pulse shape with specifications of SwissFEL with parameters of the investigated pulse transformer and pulse shape resulting from the optimization procedure with unconstrained parameters. Both displayed pulses have an equal starting point and a pulsewidth of 3 μs.

Table VI

<table>
<thead>
<tr>
<th>Primary turns</th>
<th>S</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of cores</td>
<td>3</td>
</tr>
<tr>
<td>Secondary winding height</td>
<td>38.93 cm</td>
</tr>
<tr>
<td>Core window width</td>
<td>10.94 cm</td>
</tr>
<tr>
<td>Distance core tank</td>
<td>14.94 cm</td>
</tr>
<tr>
<td>Tank volume</td>
<td>0.9596 m³</td>
</tr>
<tr>
<td>Rise time</td>
<td>2.85 μs</td>
</tr>
<tr>
<td>Conversion efficiency</td>
<td>97.7%</td>
</tr>
<tr>
<td>L_p</td>
<td>721.27 μH</td>
</tr>
<tr>
<td>C_s</td>
<td>380.96 pF</td>
</tr>
<tr>
<td>L_a</td>
<td>41.6 μH</td>
</tr>
<tr>
<td>C_a</td>
<td>100 pF</td>
</tr>
</tbody>
</table>

The optimization results are displayed, when the parameters are optimized to high conversion efficiency.

It can be observed that the optimization procedure results in a similar core geometry, but increases the secondary winding height, reduces the distance at the bottom of the windings and increases the number of primary turns to two. Due to a higher number of turns, the leakage inductance and therefore the rise time as well as the damping of the pulse is increased. On the contrary, the time to flat top is reduced, because the pulse with parameters of the investigated transformer overshoots and complies later with the flat-top criteria, as shown in Fig. 14, where a section of both pulse shapes is displayed. Thus, by applying the optimization procedure, an increase in conversion efficiency including pulse losses from 53% to 61.8% can be achieved corresponding to a relative improvement of 16.6%.

The proposed optimization procedure is therefore well suited to improve the design process of a pulse transformer.

B. Specifications of CLIC

The proposed optimization procedure is in a second step executed for the specifications of the CLIC system (Table I). The optimization results are shown in Table VI. A section of the pulse shape is shown in Fig. 15(a). In an optimal configuration, the pulse does not exceed the voltage band of the FTS and therefore immediately reaches the flat-top criteria.

The distribution of the losses per pulse is displayed in Fig. 15(b), showing the dominance of the pulse losses with 61%, followed by the core losses with 22% and a share of the other loss components of 18%. The 2-D-FEM analysis is shown in Fig. 10, and the resulting transformer is shown in Fig. 16.

VI. CONCLUSION

In this paper, a general procedure for pulse transformer optimization is presented, considering a given set of pulse specifications.

The optimization parameters include number of primary turns, number of cores, secondary winding height, distance between the secondary and primary windings, width of the core cross-sectional area, and opening angle of the winding cone. In the procedure, the total losses consisting of pulse, core, winding, active bias circuit losses, and losses of the primary switches are minimized.

Due to the nonlinear klystron load, the pulse shape is analyzed in the time domain, ensuring that the given pulse constraints are met. For core loss estimation, measurements on a test core are performed for pulse lengths in the range of 3–300 μs.

In addition, analytical calculations of the transformer parasitics are proposed and verified with 2-D-FEM simulations as well as pulse transformer measurements.
The procedure is then applied to investigate an existing pulse transformer with the specifications of SwissFEL, to validate the method and to optimize the existing design. The resulting transformer parameters of the optimization procedure lead to a shorter time to flat top and improve the system conversion efficiency by 16.6%. Finally, the procedure is conducted for a long-pulse system with the specifications of CLIC optimizing the overall efficiency. An overall conversion efficiency of 97.7% including pulse losses is achieved.

REFERENCES


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