Design Procedure of a 14.4 kV, 100 kHz Transformer with a High Isolation Voltage (115 kV)

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ABSTRACT
In this paper, the design procedure of a 14.4 kV output voltage, 100 kHz transformer with an isolation voltage of 115 kV using Litz wire is presented. All design models, including a generalized magnetic model for the leakage and the loss calculations as well as an electrical model for the parasitic capacitance estimation for the transformer are derived and proven by measurements. For designing the insulation, a comprehensive design method based on an analytical maximum electrical field evaluation and an electrical field conform design is used. The resulting design is verified by long and short term partial discharge measurements on a prototype transformer.

Index Terms — High voltage techniques, HF transformers, insulation, isolation technology.

1 INTRODUCTION

For the new linear collider at the European Spallation Source (ESS) in Lund, 2.9 MW pulse modulators with pulsed output voltages of 115 kV and pulse lengths in the range of a few milliseconds are required (pulse specifications see Table 1). For generating such pulses, a long pulse modulator based on a modular series parallel resonant converter (SPRC) topology has been developed [1]. This converter is operated at high switching frequencies (100-110 kHz) in order to minimize the dimensions of the reactive components and the transformer. To achieve the required output voltage of 115 kV, 8 SPRC basic modules (SPRC bm) each with an output of 14.4 kV are connected in series [2], see Figure 1. Due to the series connection of the SPRC basic modules, the insulation of the last oil insulated transformer in the row has to withstand the full pulse voltage. Therefore, in this paper additionally a comprehensive transformer design and optimization procedure is presented. This design procedure includes a generalized calculation method for the leakage, the parasitic capacitance and all high frequency losses.

In the literature, several approaches are presented for designing high voltage, high frequency transformers. In [3, 4, 5] designs with nominal output voltages between 50 kV - 60 kV and a switching frequency of 20 kHz, but no partial discharge tests have been performed. The same is true for the transformer presented in [6], which is designed with respect to an isolation voltage of 15 kV, a nominal output voltage of 3.8 kV and 3 kHz operation frequency. In [7] the design is carried out for a nominal output voltage of 3 kV, a switching frequency of 10 kHz and only provides partial discharge measurements for short term tests (1 min, test voltage 28 kV).

Table 1. Pulse specifications.

<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>2.9 MW</td>
</tr>
<tr>
<td>Pulse repetition rate $P_{reb}$</td>
<td>14 Hz</td>
</tr>
<tr>
<td>Pulse width $T_p$</td>
<td>3.5 ms</td>
</tr>
<tr>
<td>Pulse duty cycle $D$</td>
<td>0.05</td>
</tr>
<tr>
<td>Pulse rise time (0.99% $V_K$) $t_{rov}$</td>
<td>150 µs</td>
</tr>
<tr>
<td>Pulse fall time (100..10% $V_K$) $t_{ofv}$</td>
<td>150 µs</td>
</tr>
</tbody>
</table>

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However, all of these transformers are either tested only under nominal field conditions [3]-[6] and/or no values for long term partial discharge measurements which are an essential life time parameter for high voltage components [8], are given [7]. In addition, the isolation voltage of 115 kV and the switching frequency range of 100 kHz -110 kHz exceed by far the designs presented in [3]-[7]. Therefore, in this paper a comprehensive design procedure of a 14.4 kV nominal output voltage, 100 kHz transformer, is presented. The procedure consists of a detailed insulation design method for an isolation voltage of 115 kV, which is verified by long term (60 min) 115kV test voltage and short term (5 min) extended test voltage (up to 136 %) partial discharge measurements. It further contains a generalized magnetic model for the leakage and loss calculations, as well as an electrical model for the parasitic capacitance estimation. Finally, this procedure is employed in an optimization algorithm, which leads to an optimal design and minimized development times and costs of the high voltage high frequency transformer.

In Section 2, first, the transformer design procedure including the models for the parasitics, the high frequency loss calculations and a comprehensive insulation design procedure is given in detail. In Section 3, the resulting insulation design is evaluated by long term nominal voltage and short term over-voltage partial discharge measurements. For verifying the losses and the parasitics, in addition, pulse measurement of a single SPRC bm are presented.

**2 TRANSFORMER DESIGN PROCEDURE**

Due to the high number of degrees of freedom encountered in the transformer design process as for example the geometric parameters of the core or the windings, an optimization procedure has been developed for optimally designing the transformer (see Figure 2) [1]. In the first step, with an electrical model of the SPRC bm the input parameters (e.g. primary current $I_{prim}$ or secondary voltage $V_{sec}$) and constraints (e.g. maximum temperature $T_{max}$ or flux density $B_{max}$) for the transformer optimization are determined for the given pulse specifications.
In the next step, a specific core and winding geometry has to be chosen before the transformer optimization is started. In order to minimize the magnetic stray field outside the transformer, an E-type core is used as core geometry, where two windings are wound around the center leg.

For the winding geometry, there are five possible basic winding configurations (see Figure 3). These are investigated in the following, with respect to the maximum electrical field, lowest electrical energy per length \( W' E \) and maximum wire to wire withstand voltage \( V_{WS} \) by varying the distance \( \Delta x \).

The standard and the flyback winding configurations (see Figures 3a and 3b) result in high \( E_{max} \), \( W' E \) and \( V_{WS} \) values [9]. The s-winding configuration (see Figures 3c and 3d) has the advantage of a minimum withstand voltage, but still high \( E_{max} \) values occur. Adding a field shape ring to the configuration in Figure 3d results in the winding arrangement given in Figure 3e. The first and the last turn of the winding are placed inside field shape rings leading to a reduced \( E_{max} \). Figures 4a and 4b show the electrical field distribution of the s-winding configuration with and without field shape rings. Comparing the maximal electrical fields for cases (d) and (e) given in Table 2, it could be seen that the occurring peak field is reduced by 43.3 %. The field shape rings are on the same potential as the respective turn and one end of the turn is soldered to the corresponding field shape ring, see Figure 7d.

Due to this arrangement the high frequency losses do not increase much because most of the load current is still conducted by the Litz wire inside the field shape ring and not by the field shaping ring.

Based on the chosen core and winding geometry (see Figure 3e) all losses and parasitics are calculated and also the maximum electrical peak field is estimated. Afterwards, a detailed model of the transformer is evaluated regarding oil gap widths and creepage paths in a FEM based post design check.

The first step of the transformer design are the core losses, which are discussed in the following.

### 2.1 Core Losses

Due to the high switching frequency \( f = 100 \text{ kHz} \), ferrite K2008 is used as core material. Since the flux has an approximately sinusoidal waveform, the volume dependent core losses are calculated with the standard Steinmetz equation [10]

\[
P_V = k f^\alpha B^\beta.
\]

Solving equation (1) at three operating points (see Table 3), extracted from the losses versus frequency curve in [11], gives the Steinmetz parameters \( \alpha, \beta \) and \( k \) [12]. The Steinmetz parameters for K2008 at a core temperature of 100 °C are listed in Table 3. The core losses have to be additionally scaled by a correction factor \( c_0 \). This factor is the ratio between the core losses, measured on R34 toroids, taken from the material (K2008) data sheet [11] and the losses taken from the chosen core shape (U126/20) data sheet [13], given for an identical operating point. In the case of the U126/20 core, \( c_0 = 2.56 \). The total averaged core losses for the E-core made of 16 U-cores

\[
P_C = P_V V_C c_0 D
\]
where $V_C$ is the core volume.

In the next step, the high frequency losses are calculated as explained in the following.

### 2.2 WINDING LOSSES

In order to minimize the high frequency losses, which include skin and proximity losses, HF-Litz wire is used for the primary and secondary winding to ensure an approximately homogenous current density across the cross section of the turn. The high frequency losses per length in a Litz wire bundle are calculated by [14]

$$P'_{\text{Bundle}} = N_S \left( F_{St} \frac{l^2}{N^2_S} + G_{St} \frac{l^2}{\pi r_a^2} \int |H|^2 dA \right)$$

(3)

$$= N_S \left( F_{St} \frac{l^2}{N^2_S} + G_{St} H_{avg} \right)$$

(4)

where $F_S$ and $G_S$ are the skin respectively the proximity effect factors, $N_S$ is the number of strands per bundle, $I$ is the total peak current of the winding and $H$ is the peak magnetic field of each bundle. Figure 5d shows the norm of the total magnetic field $|H|$ inside the core window. For calculating the high frequency losses, this field can be calculated by the superposition of the internal field $H_{int}$ (Figure 5a) and the external field $H_{ext}$ (Figure 5b) inside the turns. Considering additionally the field in the space outside the turns $H_S$ (Figure 5c), also the leakage inductance can be calculated. The internal Field $H_{int}$ is a radial field inside each Litz wire, caused by the current in each wire and it is not effected by any other field, see Figure 5a. The external field $H_{ext}$ inside a considered wire is caused by the currents respectively the fields of all other wires but not by the considered wire itself and the field $H_S$ in the space outside the turns is caused by the currents of all conductors. All field components can be calculated directly with the help of the mirror current method [1]. For the external field components, equations (15) and (16), from [1] can be used inside the core window and for the region outside the core window, these equations can be simplified for the x-component to

$$H_{ext,x} = \sum_{j=1}^{N} \frac{l_j}{2\pi} \left( \frac{y - y_j}{(x - x_j)^2 + (y - y_j)^2} + \frac{y - y_j}{(x + x_j)^2 + (y - y_j)^2} \right)$$

(5)

$$H_{ext,y} = \sum_{j=1}^{N} \frac{l_j}{2\pi} \left( \frac{x - x_j}{(x - x_j)^2 + (y - y_j)^2} + \frac{x + x_j}{(x + x_j)^2 + (y - y_j)^2} \right).$$

(6)

These equations can be also applied for $H_S$. For the radial internal field component $H_{int}$ inside the core window as well as outside the core window

$$H_{int,x} = \frac{l_j}{2\pi} \left( \frac{y - y_j}{r_a^2} \right)$$

$$H_{int,y} = \frac{l_j}{2\pi} \left( \frac{x - x_j}{r_a^2} \right)$$

(7)

are employed.

The total magnetic field $|H|$ is the norm of the superposition of the three magnetic field component vectors

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

(8)

with

$$H_x = H_{int,x} + H_{ext,x} + H_{S,x}$$

$$H_y = H_{int,y} + H_{ext,y} + H_{S,y}.$$
The high frequency losses then are given by

\[ P_{W_{\text{HF}}} = N_S l_m \left( \frac{N F_{\text{St}} l^2}{N_S^2} + \sum_{n=1}^{N} G_{\text{St}} H_{\text{avg},n} \right) \text{D}, \]  

(10)

with

\[ H_{\text{avg},n} = \frac{1}{\pi r_a^2} \int |H|^2 dA_n, \]  

(11)

where \( N \) is the number of turns and \( l_m \) is the mean winding length of the considered winding. \( H_{\text{avg},n} \) is the mean averaged field, calculated for each turn cross section \( A_n \) separately.

### 2.3 TRANSFORMER PARASITICS CALCULATION BASED ON CONDUCTOR ARRAYS

In order to design the components of the resonant tank of a SPRC bm correctly, the parasitics of the transformer have to be estimated accurately. The image charge method [1] matches well with the 2D-FEM based analysis when calculating the transformer parasitics, but the analysis of large geometries results in a high computational effort for the image charge method. This effort originates from a high number of calculation points, which must be derived to approximate the surface integral with the desired high accuracy. Thus, for larger geometries the application of the image charge method for calculating the transformer parasitics is not beneficial in comparison to a 2D-FEM based analysis. In order to have a fast calculation of the parasitics multi-conductor arrays can be used. In this case, all capacitances and inductances between the conductors are derived by solving the corresponding field quantity integrals directly with the given geometric distances of the conductors (see Figure 6). The core acts as an electric mirror due to its electrical conductivity, so that the conductors can be mirrored as described in [1]. This analysis is conducted in the two dimensional space. The calculation method of the stray capacitance \( C_d \) via multi-conductor arrays including image charges has been presented in [15] and therefore is only shortly repeated in the following.

#### 2.3.1 STRAY CAPACITANCE CALCULATION BASED ON CONDUCTOR ARRAYS

The potential coefficient \( p_{kl} \) between conductor \( k \) and \( l \), considering image charges due to grounded surfaces, can be described by

\[ p_{kl} = \frac{1}{2 \pi \varepsilon_0 r_c} \ln \left( \frac{d_{kl}}{r_c} \right) + \sum_{m=mn_1}^{N/2} \ln \left( \frac{d_{k,mpl}}{d_{k,mnl}} \right) \]  

(12)

where \( d_{kl} \) is the distance from conductor \( k \) to conductor \( l \), \( r_c \) is the conductor’s radius, \( d_{k,mpl} \) is the distance of conductor \( k \) to all positive image charges and \( d_{k,mnl} \) to all negative image charges of conductor \( l \). With this equation the potential coefficient matrix can be derived, which can be transformed to the capacitance coefficient matrix by a matrix inversion. This inverted matrix contains the inter-conductor capacitances, which can be used to derive the equivalent stray capacitance of the system by summation of the energy stored in each of the inter-conductor capacitances.

Next, the calculation of the leakage inductance based on multi-conductor arrays is presented.

#### 2.3.2 LEAKAGE INDUCTANCE CALCULATION BASED ON CONDUCTOR ARRAYS

Assuming a multi-conductor system the magnetic flux per unit length \( \Phi'_{B} \) is obtained by

\[ \left[ \Phi'_{B} \right] = \left[ L' \right] \left[ I \right]. \]  

(13)

Since the analysis is limited to the two-dimensional space, the interconnection of the conductors does not matter [16]. The external inductance \( L'_{kl,\text{ex}} \) between conductor \( k \) and conductor \( l \), assuming both conductors with equal radius \( r_c \), can be described by

\[ L'_{kl,\text{ex}} = \frac{\mu_r}{2 \pi} \ln \left( \frac{d_{kl}}{r_c} \right), \quad k \neq l, \]  

(14)

where \( d_{kl} \) is the distance between the two conductors. The internal inductance \( L'_{kk} \) caused by the conductor itself can be described by

\[ L'_{kk} = \frac{\mu_r}{8 \pi}. \]  

(15)

Since the magnetic core has a very high relative permeability \( \mu_r \), it can be considered as magnetic mirror [17]. In order to take this effect into account, the conductors can be mirrored.
analogously with the image charge method. In contrast to the image charges in case of the stray capacitance $C_d$, the direction of the current flow remains equal for the mirrored conductors. Considering the image conductors, the total self inductance $L_{k,k,box}$ inside the core window can be described by

$$L'_{k,k,box} = \frac{\mu_r}{2\pi} \left( \frac{1}{4} + \sum_{m \neq k}^{N} \ln \left( \frac{d_{k,m} \cdot r_c}{r_c} \right) \right)$$

(16)

where $d_{k,m}$ is the distance of conductor $k$ to all its image conductors. The external inductance in this case is

$$L'_{k,ex} = \frac{\mu_r}{2\pi} \left( \ln \left( \frac{d_{k,l}}{r_c} \right) + \sum_{mc}^{N} \ln \left( \frac{d_{k,mc} \cdot r_c}{r_c} \right) \right), k \neq l$$

(17)

where $d_{k,l}$ is the distance of conductor $k$ to conductor $l$, $d_{k,mc}$ the distance of conductor $k$ to all image conductors of conductor $l$. If a homogeneous current distribution over all conductors is assumed, the total inductance per unit length of the geometry can be obtained by

$$L'_\sigma = \sum_{k} \sum_{l} L'_{kl}$$

(18)

Both quantities, the capacitance per unit length $C'_\sigma$ and the leakage inductance per unit length $L'_\sigma$, must be multiplied with its associated length, to obtain the desired quantities $C_d$ and $L_\sigma$.

A comparison between FEM calculations, the image charge method and the conductor array method with respect to measurements is given in Table 4. The error is smaller than 2 % in the case of the leakage inductance for all methods in relation to the conducted measurement. In the case of the stray capacitance, it can be noticed that FEM I calculations, image charge and the conductor array method result in similar values, but the error to measurement is in the range of 25 %. This can be explained by the fact that only a single permittivity value for the geometry is assumed. The error is reduced below 2 % in FEM II if all permittivities of the different insulation materials are considered. Based on the image charge method and the FEM evaluations an insulation design procedure is introduced in the following Section.

### 2.4 INSULATION DESIGN PROCEDURE

In the following, the insulation design procedure (areas high-lighted in gray in Fig.2) which can be divided into an analytical maximum electrical field evaluation and a FEM supported post insulation field conform design check are described more in detail.

#### 2.4.1 EVALUATION OF THE MAXIMUM ELECTRICAL FIELD

Due to the complexity of the transformer insulation structure, it is not computationally efficient to use a comprehensive analytical model of the transformer including all details as e.g. bobbins, winding fastenings and oil gap barriers (see Figure 7) in the optimization procedure. Instead an analytical maximum electrical field calculation is used, which is based on the image charge method [1] and which allows a fast basic insulation design check considering the maximal electrical field that has to be below a certain constraint value (see Figure 2). This method considers a single insulation material permittivity and is more than 7 times faster than FEM, because only a few points along the surface of the turn with the highest potential are evaluated to estimate the highest $E_{max}$ value.

#### 2.4.2 FEM SUPPORTED FIELD CONFORM DESIGN

In the following, the FEM supported field conform post design procedure is discussed. The main insulation material is the transformer oil MIDEI7131 [18] with a relative permittivity of 3.2. All other insulation materials are chosen with respect to the transformer oil such that they have a similar permittivity (see Table 5) to avoid local field

| Table 4. Comparison between FEM, image method, conductor array method and the measured parasitics |
|-----------------|-----------------|-----------------|-----------------|-----------------|
| FEM I | FEM II | Image Cond. array | Measurement |
| $C_d$ (pC) | $C'_\sigma$ (pC) | $L_\sigma$ (µH) | $L'_\sigma$ (µH) |
| 16.7 | 22.1 | 16.4 | 17.2 | 21.7 | 0.84 | 0.84 | 0.84 | 0.83 |
| Error (%) | -23 | +1.8 | -24.4 | -20.7 | - | +1.2 | +1.2 | +1.2 | - |

Figure 7. (a) Transformer prototype (built by AMPEGON AG). (b) Top view of the transformer, which has no mountings between primary and secondary inside the transformer, (c) core window with oil gap barrier and (d) last turn mounted inside the field shape ring.
enhancements at the boundary layer of different materials and maximum electrical strength. Figure 7a shows the transformer prototype, which has no fastenings between the primary and the secondary bobbin inside the transformer in order to avoid creepage paths (see Figure 7b). The bobbins are fastened outside of the core window at the top and the bottom of the transformer, resulting in a longer creepage distance between the windings (see Figures 8a and 8b). The maximum field strength occurs at the field shape ring inside the core window. Hence, triple points [19] between the field shape rings and the secondary bobbin inside the core window must be avoided (see Figure 7c). Thus, all field shape ring fastenings are located in a region of weak E-field outside the core window. Hence, triple points [19] between the field shape rings and the secondary bobbin inside the core window must be avoided (see Figure 7c). Thus, all field shape ring fastenings are located in a region of weak E-field outside the core window as can be seen in Figure 7d. The primary bobbin is completely sintered of PA2200 material in a 3D printing process. This process allows complex designs but the resulting components are not void free [24], so this material is used only in non critical electrical field areas. The secondary bobbin is milled out of a single solid POM block to minimize voids and component intersections (see Figure 7b). Additionally, silk wrapped Litz-wire is used instead of foil so that no air bubbles are trapped beneath the foil.

An inappropriate design of the insulation causes partial discharges as well as sliding discharges which can harm the insulation of the transformer permanently and lead to arcs between the windings or the core. Oil gap barriers between primary and secondary winding as well as between the secondary winding and the core are used to counter the decreasing electrical strength of long oil gaps due to the volume and the area effect [19]. Therefore, for long life times a proper insulation design is necessary and a detailed analysis of the electrical field distribution along long oil paths (P1-P6, see Figures 8a and 8b) and critical creepage paths (P7, see Figure 8b) was carried out with the help of the Weidmann design curve method [25]. There, the ratio of oil design curves ($E_d(z)$) which are derived from homogenous electrical breakdown tests [26] and the averaged cumulated electrical field strength $E_{avg}$ along certain path lengths ($z$) is calculated, resulting in safety factor curves $q$ [19].

\[
q = \frac{E_d(z)}{E_{avg}(z)}
\]

$E(z')$ is the electrical field at point $z'$. The factor $q$ has to be multiplied by 0.7 if used for creepage paths [19]. The used design curve is derived from lightning impulse tests and downscaled by a design insulation level (DIL) factor of 2 to 50 Hz conditions in [29], which is in the same frequency range as the pulse repetition rate of 14 Hz of the output pulse voltage of 115 kV. For a valid design, the $q$-values of all evaluated paths have to be above 1 (see Figure 9). With this method, insulation designs with homogenous as well as with strongly inhomogenous field distributions can be investigated. Finally, applying this method results in an electrical field conform design, which means that the equipotential lines have mostly tangential components along the surface of insulation boundaries, e.g. oil gap barriers (see Figure 8). Hence, the insulator is stressed mostly by the normal component of the electrical field and has its maximum electrical strength.
3 MEASUREMENT RESULTS

For verifying the design, partial discharge and output pulse measurements results are presented in the following.

3.1 PARTIAL DISCHARGE MEASUREMENTS

For long life times, it is not sufficient to know if the transformer withstands a certain voltage level without any breakthroughs. It is also of high importance to know if the transformer is suffering from partial discharges. Such discharges can harm the insulation permanently during normal operation and may lead to serious failures. Therefore, in this section the results of comprehensive partial discharge tests are presented.

The insulation of a single SPRC bm transformer has to withstand an operating voltage of 14.4 kV. Due to the series connection of the basic modules (see Figure 1) the required isolation voltage is increasing by 14.4 kV per SPRC bm. Hence, the last transformer in the series connection has to isolate the full output voltage of 115 kV (= 8 x 14.4 kV). Figure 10 shows the partial discharge measurement setup. The transformer (DUT) is placed inside the oil tank where its primary is shorted and grounded via the metal plate. The secondary is also shorted and connected to the high voltage electrode inside the double toroid. This double toroid is used to compensate the different material intersections at the end of the field shaping pipe, which could lead to additional external partial discharges. The whole setup is placed in a Faraday cage and has a ground noise level of about 300 pC. For optimal test conditions, the oil has been processed through a filtration system resulting in a relative moisture level lower than 6 %.

To reduce possible partial discharges to a minimum, it is important to remove the air out of the DUT. Therefore, the oil tank has been filled under a pressure of 200 mbar below atmospheric pressure. Further reduction of air bubbles has been achieved by rotating the DUT within the oil. All measurements have been carried out at a room temperature of 23.5 °C. The partial discharges $Q$ are recorded with an Omicron MPD600 measurement system [27] and are evaluated according to the IEC 60270:2000 standard [28] leading to $Q_{IEC}$.

For a valid insulation design, the DUT has to pass the following test procedure. First, the nominal operation voltage (115 kV/$\sqrt{2}$ = 82 kV RMS, frequency $f$ = 50 Hz) is applied as test voltage $V_{Test}$ to the DUT for 60 min. No breakthrough should occur and the partial discharge level $Q_{IEC}$ should be below 2 pC. Afterwards, $V_{Test}$ is increased stepwise up to a voltage of 110 kV RMS (136 %) with a test duration $T_{Test}$ of 5 min for each step. To pass the test, no breakthrough must occur. The DUT passed the first test with a $Q_{IEC}$ value far lower than 2 pC, as can be seen in Figure 11. Figure 12 shows a typical phase resolved discharge pattern which has been
recorded during the green time interval in Figure 11b. Most of the discharges occur at or near the positive or the negative half wave maximum of the test sinus. This could be interpreted as a combination of corona and void or surface discharges with single ended contact to one electrode according to [19]. Also the second test is passed successfully as depicted in Figure 13.

In the next step, $V_{\text{Test}}$ is increased stepwise (+9 %) from +100 % to +136 % (see Figure 13a). There, also no breakthrough occurred (see Figure 13b). The $Q_{\text{IEC}}$ level still remains below 2 pC for the first two voltage steps. The peaks in $A'$, $A''$ and $A'''$ are caused by the variable ratio transformer which is the controlled primary main supply of the HV test transformer.

3.2 PULSE MEASUREMENTS

Figure 14a shows the measured pulsed output voltage $V_{01,\text{meas}}(t)$ of a single SPRC bm operated under nominal conditions, with a pulse length of 3.5 ms and a repetition rate of 14 Hz. The module is open loop controlled and no droop compensation is active. The green curve is the mean average voltage $V_{01,\text{avg}}(t)$ of the measured output voltage and is shown to determine the rise and the fall time. Both times in Figures 14b and 14c are well below the given limits in Table 1. No breakthroughs occurred and all temperatures of the oil-immersed transformer did not exceed the temperature limits during the test duration of 60min. Table 6 summarizes the optimized parameters of the transformer.

4 CONCLUSION

In this paper, a design procedure of a 14.4 kV nominal output voltage, 100 kHz transformer with an isolation voltage of 115 kV is presented. The procedure contains a generalized magnetic model for the leakage and an electrical model for the parasitic capacitance estimation which are proven by measurements and FEM calculations. The error is lower than 2 % in the case of the leakage inductance whereas in the case of the stray capacitance the error is in the range of 25 % if only the permittivity of air is used for the calculations. The error is decreased below 2 % if all permittivities of the different insulation materials are considered.

In addition, a comprehensive insulation design procedure which is also part of the transformer optimization is derived in detail. It consists of a fast analytical maximum electrical field evaluation used during the automatic optimization of the transformer and a field conform post processing insulation design check. The resulting insulation system is verified by partial discharge
measurements. First, a 60 min long test at nominal voltage is performed and afterwards the test voltage has been increased from 100 % to 136 % in 9 % voltage steps. Each voltage step is applied for 5 min. Both tests are passed with no breakthroughs and the partial discharge level is lower than 2 pC at nominal voltage. For validating the design additionally, pulse measurements of a SPRC bm, with a pulse length of 3.5 ms and a repetition rate of 14 Hz are presented. No breakthroughs occurred and all temperatures of the oil-immersed transformer did not exceed the temperature limits during the test duration of 60 min.

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Jürgen Biela (S’04–M’06) received the Diploma (Hons.) degree from Friedrich-Alexander-Universität, Erlangen-Nürnberg, Germany, and the Ph.D. degree from the Swiss Federal Institute of Technology (ETH Zurich), Switzerland, in 1999 and 2006, respectively. In 2002, he joined the Power Electronic Systems Laboratory (PES), ETH Zurich, for working toward the Ph.D. degree, focusing on optimized electromagnetically integrated resonant converters. From 2006 to 2007, he was a Post-Doctoral Fellow with PES and a Guest Researcher with the Tokyo Institute of Technology, Tokyo, Japan. From 2007 to 2010, he was a Senior Research Associate with PES. Since 2010, he has been an Associate Professor in high-power electronic systems with ETH Zurich. His current research interests include the design, modeling, and optimization of PFC, dc–dc and multilevel converters with emphasis on passive components, and the design of pulsed-power systems and power electronic systems for future energy distribution.