Harmonic Domain Modelling of Transformer Core Nonlinearities Using the DIgSILENT PowerFactory Software

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Summary: This paper demonstrates the results of implementation and verification of an already existing algorithm that allows for calculating saturation characteristics of single-phase power transformers. The algorithm was described for the first time in 1993. Now this algorithm has been implemented using the DIgSILENT Programming Language (DPL) as an external script in the harmonic domain calculations of a power system analysis tool PowerFactory [10]. The algorithm is verified by harmonic measurements on a single-phase power transformer. A theoretical analysis of the core nonlinearities phenomena in single and three-phase transformers is also presented. This analysis leads to the conclusion that the method can be applied for modelling nonlinearities of three-phase autotransformers.

1. INTRODUCTION

In harmonic analysis studies, power transformers can be either modelled as linear power system components, or using more sophisticated models that take into account the iron core nonlinearity and its property of distorting currents resulting in an “injection” of harmonic currents into the network. If the transformer core is to be modelled as an impedance, it is sufficient to model the iron core as a shunt linear inductance and a resistance, connected in parallel. In order to model the nonlinear behaviour of the core, voltage dependence of an equivalent inductance and resistance shall be also included. Presented in this paper modelling of nonlinear cores of power transformers is a part of a larger project, where an existing computer model of the western grid of the Danish TSO Energinet.dk is to be adjusted for harmonic analysis purposes. Besides all linear power system components, special focus is attached to modelling the nonlinearities present on the transmission level: converters of HVDC links and power transformers/autotransformers.

In order to minimize the core size, power transformers are usually designed to operate just below the knee point. Therefore, when the voltage increases above a certain value, the transformer magnetising current becomes distorted; even at the nominal operating voltage the no-load current THD can be as high as 20% [4]. The harmonic contribution from power transformers can be most likely noticed in a grid with many transformers installed, during low load hours, when the load current is low and the voltage rises [4].

DIgSILENT PowerFactory enables to precisely model transformer core nonlinearity in the time domain (EMT) [8]. In the frequency domain it is not possible to do it so far. Since the frequency domain is more convenient for harmonic analysis than EMT, an attempt was made to model transformer core nonlinearity in the harmonic domain calculations of PowerFactory. Several ways of modelling the iron core nonlinearities exist, described for instance in [5, 9 or 2].

These methods require that the dynamic hysteresis loops of the material are known. This is not always possible, especially if the nonlinearities of older transformers are of interest. In this project the only data available for the autotransformers are typical test reports. Therefore, a different method has been found, where the core nonlinearities can be determined using a simpler set of input data [7]. For computation, the method requires only $V_{rms} - I_{rms}$ curves and the no-load losses at fundamental frequency. These data are present in the test reports. The method was implemented as an external script of DIgSILENT PowerFactory using the DIgSILENT Programming Language (DPL). The implementation of the method and its verification is illustrated. It is also shown that the method is applicable for modelling nonlinearities of power autotransformers.

2. THEORY

In an iron-core power transformer, the flux $\Phi$ and the flux density $B$ within the transformer core is proportional to the voltage $u_1$ applied across the primary winding, according to the Faraday’s law. In the sinusoidal case it can be written [1]:

$$u_1 = U_{1m} \sin \omega t = \frac{d\Phi}{dt}$$

(1)

$$\Phi = \int U_{1m} \sin \omega t \, dt = \frac{U_{1m}}{\omega} \cos \omega t = \Phi_m \cos \omega t$$

(2)

$$B = \frac{\Phi}{S} = \frac{\int U_{1m} \sin \omega t \, dt}{S}$$

(3)

where $S$ is the cross-section of the core.

The resultant magnetizing current $I_m$ needed to produce this sinusoidal flux is related to the magnetic field strength...
The magnetic flux density $B$ is related to the magnetic field strength $H$ by the magnetic permeability of the steel used in the transformer core $\mu$:

$$B = \mu H$$ \hspace{1cm} (5)

Equation 5 shows that if the magnetic permeability $\mu$ is linear and $B$ is sinusoidal, the magnetic field strength $H$ (and what follows, the magnetizing current $I_m$) will be sinusoidal. If the transformer operates in saturation, $\mu$ is no longer linear and the magnetic field strength $H$ is nonsinusoidal, because the product of $H$ with the magnetic permeability $\mu$ must still produce the sinusoidal flux density $B$, see Figure 1. If the magnetizing current $I_m$ is allowed to flow freely (low impedance path $Z$ for all harmonics), after decomposing it to Fourier components, it will contain the fundamental frequency and all the odd harmonics: $3^{rd}$, $5^{th}$, $7^{th}$, $9^{th}$ etc [1].

3. THE ALGORITHM AND ITS IMPLEMENTATION

In order to determine the magnetizing current harmonics, a special algorithm has been implemented [7]. Initially this technique was proposed in [3] for estimating the magnetizing inductance $L_m$. Reference [7] extended this algorithm with computation of the nonlinear resistance $R_{Fe}$ (see Figure 2a). The main advantage of this algorithm is that both the equivalent resistance $R_{Fe}$ and inductance $L_m$ are calculated directly from the conventional transformer no-load test data: $V_{rms} - I_{rms}$ curve and the no-load losses at fundamental frequency. The nonlinearity (piecewise linear $u$-$i$ characteristic) accounts for two effects: linear eddy current loss and nonlinear hysteresis loss. This resistance is found from the measured total no-load losses. Calculated in this way the resistive part of the total, measured no-load current can be subtracted geometrically. This allows for a calculation of the current flowing through the nonlinear inductance (piecewise linear $\lambda$-$i$ characteristic). Next, the resulting, real and imaginary harmonic components are geometrically added and modelled as a current source. During harmonic load flow calculations, this current source will inject harmonic currents into the investigated network (Figure 2b).

The algorithm requires that the following assumptions and simplifications are made:

- winding impedances are ignored,
- supplying voltage and resulting flux are sinusoidal,
- the piecewise resistance and piecewise inductance are both symmetric.

The algorithm has been implemented in a dedicated simulation software DIgSILENT PowerFactory (PF). PowerFactory allows for creating automatic simulation tasks and build miscellaneous models of power system components, as the program possesses its own programming language DPL (DIgSILENT Programming Language). The syntax of this language is similar to C++. This language was used to implement the described algorithm. The basic idea was that the whole algorithm was programmed as several DLP subroutines and a main DPL controlling script. The structure of the algorithm implemented in PF is shown in Figure 3, and individual components of the algorithm are described in the following sections. Special emphasis is put

![Fig. 1. Typical relations between main flux $\Phi$, supplying voltage $U$ and no-load (magnetizing) current $I$ (without hysteresis effects) [6]](image1)

![Fig. 2. Equivalent diagram of a power transformer: $R_{IN}, L_{IN}$ – winding resistances, $L_{IN}, L_{LV}$ – Leakage inductances; a) $R_{Fe}$ – magnetizing resistance – hysteresis losses and eddy current losses in the core, $L_m$ – magnetizing inductance; b) In the harmonic domain, nonlinearities are modelled as current source; $I_m$ – magnetizing current](image2)
on the implementation details, and calculated, ready-to-use equations.

3.1. Routine Main

The main script Main (see Fig. 3), when executed recalculates all input data stored in the Data matrix \((U_m, I_m)\) to the peak values and stores them in the matrix of results ‘Results’. Letter \(m\) denotes the number of available measurements stored in Data. Next, the script automatically connects a current source to the HV transformer bus. Then it initiates conventional load-flow calculations. As a result of the load-flow, the three-phase voltages at the transformer bus are obtained. Next, the script Main executes in turn all the subroutines, beginning with subroutine volt_comp.

3.2. Subroutine volt_comp

This subroutine compares the calculated terminal voltages with the voltages stored in Data matrix and returns the row number \(f\) of the closest voltage value. If the bus voltage is higher than the highest voltage in the input data, the subroutine returns the max number of rows \(f = m\). This means that the distorted magnetizing current will be constant for higher voltage values. This shall not be problematic during normal operation, since the voltage in no-load tests is usually varied up to 1.1 Un. When Subroutine volt_comp is accomplished, main script executes the next subroutine iR.

3.3. Subroutine iR

This subroutine calculates the first linear and next piecewise linear elements of the nonlinear resistance curve \(R\). It is done in the following way. For the first segment, i.e. up to the first measured voltage, the peak current and the first linear segment of the resistance curve is calculated directly from input data:

\[
P_1 = V_{rms_1} I_{rms_1}
\]

The power measured at higher voltage values is a product of the sinusoidal voltage and nonsinusoidal current \(I_m\) (see Fig. 4). It can be seen that if linear resistance determined by \((8)\) is used, the resultant current would be sinusoidal, regardless of the amplitude of the applied voltage. The peak value of this current would be much lower than in reality. Therefore, in order to assess the real current and the next segments of the piecewise linear curve, the power definition must be used:

\[
P = \frac{\pi}{2} \int_0^\pi v(\alpha) i(\alpha) d\alpha
\]

It is sufficient to calculate the current shape for \(1/4\) of a period because the voltage is assumed sinusoidal. Therefore, the resultant current will contain only odd harmonic components and therefore, will be symmetrical.

The applied voltage can be written as \(v(\alpha) = V_k \sin(\alpha)\). The total no-load power loss:

\[
P_k = \frac{2}{\pi} \int_0^{\pi/2} (V_k \sin(\alpha) \frac{V_k \sin(\alpha)}{R_1} d\alpha + \int_{\pi/2}^\pi (V_k \sin(\alpha) \frac{I_k + V_k \sin(\alpha) - V_1}{R_2} d\alpha + \int_\pi^{3\pi/2} (V_k \sin(\alpha) \frac{I_{k-1} + V_k \sin(\alpha) - V_{k-1}}{R_k} d\alpha
\]

Fig. 3. Detailed structure of the implemented algorithm
The angles \(a_1, a_2, \ldots, a_{k-1}\) are calculated from \(a_j = \arcsin(V_j/V_k), j = 1, 2, \ldots, k-1\) and stored at each recursive step in the matrix of phase angles \(\text{Angles}\). The computation using (10) must be done recursively — segment by segment. The part of (10) that is underlined with a single line shall be recomputed at each recursive step, because the voltage \(V_k\) must be updated. Part of (10) underlined with double line must be updated and computed recursively — segment by segment. The part of (10) underlined with a single line shall be recomputed at each recursive step, because the voltage \(V_k\) must be updated.

4. Sinusoidal voltage applied to the piecewise linear \(v\)-\(IR\) curve and the resultant, approximated nonlinear current

\[
P_k = a_k + \frac{b_k}{R_k} \Rightarrow R_k = \frac{b_k}{P_k - a_k}
\]

(11)

Components \(a_k\) and \(b_k\) were computed and are given below in a ready-to-use fashion:

\[
a_k = \frac{2}{\pi} \left[ \frac{V_1}{R_k} \frac{1}{2} \left( a_1 - \frac{\cos a_1 \sin a_1}{2} \right) + \frac{V_k}{R_k} \int_{r(k-1)}^{r(k)} \left( \cos \alpha(k-1) - \cos \alpha(k) \right) \right]
\]

(12)

\[
b_k = \frac{2}{\pi} \left[ \frac{V_1}{R_k} \frac{1}{2} \left( a_1 - \alpha(k-1) + \alpha_1 - \alpha(k) \right) - \frac{V_k}{R_k} \int_{r(k-1)}^{r(k)} \left( \cos \alpha(k-1) - \cos \alpha(k) \right) \right]
\]

(13)

Next, the subroutine \(iR\) must calculate the current \(I_{r2}\) from:

\[
I_{r2} = I_{r1} + \frac{V_k - V_{k-1}}{R_k}
\]

(14)

When \(R_2\) and \(I_{r2}\) is known, the calculations can be started again, now for \(k = 3\), to determine \(R_3\). This procedure is repeated for each segment (each voltage in the \(Data\) matrix) except for the first one, i.e. for \(k \geq 2\). Resistance \(R\) and current peak \(I_{rk}\) values are stored in the matrix \(Results\).

3.4. Subroutine \(iL\)

The \(\lambda\)-\(iL\) curve is calculated by the Subroutine \(iL\). Peak flux values \(\lambda_k\) are related to the peak voltage values \(V_k\) by the expression [7]:

\[
\lambda_k = \frac{V_k}{\alpha}
\]

(15)

where \(I_{r\text{rmsk}}\) is the total measured rms no-load current, \(I_{r\text{rmsk}}\) is the determined in previous section rms resistive current component and \(I_{r\text{rmsk}}\) is the inductive rms current component to be computed. For the first segment in the \(\lambda\)-\(iL\) curve subroutine \(iL\) calculates the assumed linear inductance using equations:

\[
I_{iL} = I_{r\text{rmsk}} \sqrt{2}
\]

(16)

\[
\lambda_1 = \frac{V_1}{\alpha}
\]

(17)

\[
L = \frac{V_1}{I_{iL}} = \frac{V_1}{\alpha I_{iL}}
\]

(18)

For segments \(k \geq 2\), assuming \(\lambda_k(\alpha) = \lambda_k \sin(\alpha)\), rms definition for the current will be:

\[
I_{r\text{rmsk}}^2 = \frac{2}{\pi} \int_{\alpha_{k-2}}^{\alpha_k} \left( I_{iL} + \frac{\lambda_k \sin \alpha - \lambda_{k-1}}{L_k} \right) d\alpha + \ldots + \int_{\alpha_{k-2}}^{\alpha_k} \frac{\lambda_{k-1} \sin \alpha - \lambda_k}{L_k} d\alpha
\]

(19)

This equation is analogous to (10), i.e. it must be computed segment by segment. This equation can be rewritten in the form:

\[
a_{ilk} \left( \frac{1}{L_k} \right)^2 + b_{ilk} \left( \frac{1}{L_k} \right) + (c_{il} + x_0) = 0
\]

(20)

For this quadratic equation it is assumed [7] that \(a_{ilk} > 0\), \(b_{ilk} > 0\) and \((c_{il} + x_0) < 0\). Therefore, the linear segments of the inductance \(L_k\) are calculated from:

\[
L_k = \frac{2a_{ilk}}{-b_{ilk} + \sqrt{b_{ilk}^2 - 4a_{ilk}(c_{il} + x_0)}}
\]

(21)

Components \(a_{ilk}, b_{ilk}, c_{il}, x_0\) were computed and are given below in a ready-to-use fashion:
\[ x_0 = \frac{\lambda_k^2}{L_k} \left[ \frac{\cos \alpha (k-1) \sin \alpha (k-1) - \cos \alpha (k) \sin \alpha (k)}{2} \right] + \] 
\[ a_k = \frac{-\lambda_k^2 \cdot (x_1 + \lambda_k^2 \cos \alpha)}{2 \lambda_k \lambda_k (\cos \alpha - \cos \alpha)} \] 
\[ b_k = \frac{-2 \lambda_k \lambda_k (\cos \alpha - \cos \alpha)}{2 \lambda_k \lambda_k (\cos \alpha - \cos \alpha)} \] 
\[ c_k = \left[ - I_{k}^2 \right] (\sin \alpha - \sin \alpha) \] 

and the peak current \( I_k \) must be also computed at each segment from:

\[ i_k = i_{k-1} + \frac{1}{L_k} (\lambda_k - \lambda_{k-1}) \]  

### 4.5. Subroutine iRILrms

This subroutine calculates the \( rms \) value of the distorted resistive current component \( I_{k,rms} \) and later, the \( rms \) value distorted inductive component \( I_{k,rms} \) at each recursive step. The \( rms \) value of the resistive current \( I_{k,rms} \) is found from the \( rms \) definition:

\[ I_{rms}^2 = \frac{\pi}{2} I_0^2 (\alpha) d\alpha \]  

\[ I_{rms}^2 = \frac{\pi}{2} \int_0^\alpha \left[ \left( \frac{V_s \sin \alpha}{R_1} \right)^2 + \right] d\alpha + \] 
\[ \int \left( \frac{I_n + V_s \sin \alpha - V_k}{R_2} \right)^2 d\alpha + \ldots + \] 
\[ \frac{2}{R_k} \int \left( I_{k,n} + V_s \sin \alpha - V_{k-1} \right)^2 d\alpha \] 

Equation 29 is given below in a ready-to-use fashion:

\[ I_{rms}^2 = \frac{\pi}{2} \int_0^\alpha \left[ \left( \frac{V_s \sin \alpha}{R_1} \right)^2 + \right] d\alpha + \] 
\[ \int \left( \frac{I_n + V_s \sin \alpha - V_k}{R_2} \right)^2 d\alpha + \ldots + \] 
\[ \frac{2}{R_k} \int \left( I_{k,n} + V_s \sin \alpha - V_{k-1} \right)^2 d\alpha \] 

If sinusoidal voltage is applied to the parallel connection of linear inductance and resistance, the resistive current will be in phase with the applied voltage and inductive current will be shifted by 90°. In case of nonlinear resistance and inductance, the resistive and reactive currents will be distorted, i.e. they will contain harmonics but still the adequate harmonic components will be orthogonal with respect to each other [7]. Based on this assumption, the subroutine computes the inductive part of the current subtracting geometrically the computed above resistive \( rms \) component form the total measured \( rms \) no-load current.

Determined in angle domain inductive and resistive current components are stored in separate vectors of the length 256 points each. Next, a DFT function available in DPL is applied to the vectors, and the resultant harmonic magnitudes and phase angles are geometrically added and assigned to a current source. This current source shall inject the calculated harmonic currents into the investigated network during harmonic load flow studies.

### 4. VERIFICATION

The implemented algorithm has been verified first with the example data provided in the original paper [7]. The results were identical as in [7], which means that the algorithm has been implemented correctly. This verification process is presented in detail in [10]. In this paper an additional verification has been performed, against own, real-world harmonic measurements performed on a single-phase transformer. The no-load test was performed. The voltage has been changed in the range form 0.1 p.u. up to 1.3 p.u. and the following values have been measured: sinusoidal supplying voltage \( U_{rms} \), no-load current \( I_{rms} \), harmonic distortion of the no-load current, harmonics 3rd, 5th, 7th, 9th, 11th, 13th, no-load losses \( P_0 \). Next, the measured no-load current and no-load losses were used as input data for the implemented algorithm. Determined piecewise linear resistance \( R \) and inductance \( L \) are shown in Figure 5. The resultant inductive and resistive current components for saturated core (\( U = 1.3 \) p.u.) are shown in Figure 6. Figure 7 shows the content of no-load current harmonics versus exciting voltage. It can
be seen that the variation of the no-load current harmonics versus exciting voltage measured and determined using the algorithm is comparable, especially for a heavily saturated core. However, higher order harmonics show better agreement with the measured values than lower order harmonics.

5. THREE-PHASE TRANSFORMERS

5.1. Three-limb three-phase transformer

Now, a case where three, single-phase transformers are combined, to create a three-limb transformer but with all three windings supplied with the same sinusoidal (zero-sequence) voltage $U_1$ is considered, see Figure 8a. The resultant fluxes will have the same direction in all three legs, and therefore they will be pushed-out of the ferromagnetic core and run in the air, oil, tank walls, etc. The magnetic permeability of the air or oil $\mu_{\text{air}}$ is much lower than the permeability of the ferromagnetic core $\mu$, and therefore the resultant zero-sequence flux $\Phi_{\text{air}}$ will be much lower than the positive/negative sequence flux $\Phi$. This in turn means that the zero-sequence equivalent core inductance $L_{m}'$ is much smaller than the positive/negative sequence inductance $L_m$, where the voltages and resultant fluxes, shifted with respect to each other by 120 deg run entirely in the magnetic core, see Figure 8b. Moreover, the zero-sequence inductance $L_{m}'$ probably may be assumed linear, because the flux practically depends on the linear air/oil permeability, and the flux will be reduced because of the higher equivalent reluctance, which will cause the core to be less saturated. For the same reason the zero-sequence core losses may be ignored.

Therefore, justified would be the assumption that for three-leg transformers, the magnetizing current caused by the zero-sequence, sinusoidal voltage is also sinusoidal. Besides, it can be concluded that even if the magnitudes of supplying voltages are the same in both cases 8a and 8b, the magnetizing current in case 8a will be larger than in case 8b. If the flux $\Phi_{\text{air}}$ would run entirely in the air, the resultant equivalent inductance would have similar value like the leakage inductance of the primary winding $L_{HV}$. In reality the fluxes still enter the core and therefore the inductance $L_{m}'$ is several times higher than the leakage inductance of the windings $L_{HV} < L_{m}' << L_m$ [6]. In case 8b, the three-phase sinusoidal...
voltage will create a three-phase sinusoidal flux that will run entirely in the core. The consequence is that the flux will pass only through the material that is characterized by a high and nonlinear value of permeability $\mu$, and the magnetizing current will be nonlinear in the same way like in the case of a single-phase transformer (ignoring core asymmetries that introduce coupling between sequences).

5.2. Effect of winding connections

The analysis presented above was made with the assumption that the primary winding is connected in grounded star, so the impedance for all harmonics is $Z$. In that case the magnetizing current will contain all odd harmonics including that of the zero-sequence, which are the 3rd, the 9th, 15th, etc. If the grounding of the starpoint is removed, the impedance $Z$ for all zero-sequence harmonics would become infinitely large, which means that the triplen current harmonics will be blocked, see Figure 9. Since (5) must be always true, the blocked triplen current harmonics will cause the magnetic flux and the voltage at the winding terminals to be distorted. This is a very undesirable phenomenon. The zero-sequence flux components will be pushed out of the core (this will result in the lower magnitude of the flux, due to the higher reluctance) and this flux will in turn induce back zero-sequence voltage in the primary winding.

Therefore, a power transformer connected in ungrounded Wye (with no delta connected winding) if exposed to saturation will produce zero-sequence harmonic voltages. This voltage distortion can be modelled as a voltage source connected across the transformer terminals, see Figure 9.

In this figure, the equivalent diagrams for the fundamental and higher frequencies of the positive, negative and zero-sequence harmonics are shown, for two transformer winding arrangements: $Yy$ and $Yd$. When looking at the equivalent diagram, the advantage of including a winding connected in delta becomes apparent. If the starpoint of the primary winding is isolated, winding connected in delta will provide the only path for zero-sequence currents generated by the core. Therefore, the presence of a delta winding will prevent the transformer from distorting the supplying voltage with zero-sequence harmonics. A power system can usually cope with relatively high levels of current distortion. Bearing in mind that the measure of power quality in a power system is actually the voltage quality [4], it can be emphasized that the distorted current does not have the direct influence on the power quality – it must be first coupled by the system impedance $Z$, and then, by the created harmonic voltage drops.
it will affect the power quality. All the harmonics for which the diagrams are valid are created within the transformer, due to the saturation. It shall be also stated that the 4 or 5-legged transformers will behave principally like single-phase transformers – the flux will always have a return path through the core. There can be small differences noticed for each phase due to the core asymmetries.

5.3. Restrictions when using the algorithm for three-phase transformers

The described in this paper algorithm works for single-phase transformers. However, under certain circumstances it can be used also for modelling the nonlinearities of three-phase power transformers. From the theoretical analysis arises that the following restrictions regarding the no-load tests shall apply:

— The three-phase transformer must be supplied through a star connected winding YN,
— The starpoint of this primary winding shall be grounded (YN) because otherwise all triplen harmonics will be blocked,
— The voltage source that is used for supplying the transformer must be connected in grounded star YN.
— Delta connected winding shall not be used as primary because the triplen harmonics will circulate in the winding and will not be present in the measured current,
— Also none of the secondary windings shall be connected in delta. This can be deducted from the equivalent diagram shown on Figure 9. The zero-sequence harmonic currents originating in the core will split and flow into the supplying winding Yn, any other winding connected in grounded star yn terminated with a load impedance connected in YN, or any winding connected in delta.

It does not matter if the transformer core is 3, 4 or 5-legged, because the supplying voltage and the resulting flux in the core is sinusoidal and purely positive sequence. No zero-sequence flux will appear due to the core nonlinearity because the supplying winding is connected in grounded star.

5.4. The autotransformer

The main difference between a regular transformer and an autotransformer is that the autotransformer has the primary and secondary windings connected in the way that the voltage drop on the primary winding rises the potential of the secondary winding, see Figure 10.

Therefore, the secondary voltage to ground is the sum of the voltage drops on the primary and secondary coils. By raising the voltage, the power that can be transferred through the transformer is higher. However, this operation removes the galvanic insulation between the windings. There is also a difference for the dc component. The dc current is not transferred through a regular transformer while in a autotransformer it will split and the major part will close through the resistance of the primary winding and the remaining part can flow into the secondary side, depending on the dc (zero-sequence) resistance of the loads. The no-load testing procedure performed on an autotransformer does not differ from the procedure applied when testing a regular transformer. If the transformer is supplied with nominal voltage, the resultant flux density in the core has also its nominal value, whichever winding is used as the primary. The core nonlinearity of power autotransformers can be modelled using the described algorithm if they comply with the restrictions described for regular three-phase power transformers. The autotransformers installed in the Energinet.dk power system are YNyn connected and some of them have an additional tertiary winding. However, this winding is normally not used and therefore, it is connected in open delta. Consequently it will not affect the flow of harmonic currents because no current can circulate and they cannot produce opposing flux. Consequently it can be concluded that if three-phase power autotransformers comply with the restrictions described in section V.C, the algorithm can be used for modelling their core nonlinearities.

6. SUMMARY AND CONCLUSIONS

— It is possible to implement the algorithm in the PowerFactory simulation software,
— The algorithm has been verified first against the original data provided in [7] and the agreement was perfect, which proves that the algorithm has been correctly implemented,
— The comparison of the results with own measurements on a small single phase transformer has shown that
higher order harmonics show better agreement with the measured values than lower order harmonics,
— A difference in the measured and calculated 3rd harmonic
content in the no-load current, for low exiting voltage
(0.8 – 0.9 p.u.) has been noticed but the reason for this
has not been revealed,
— Based on the performed theoretical analysis it can be
stated that the algorithm initially valid for single-phase
transformers can be used for modelling nonlinearities of
three-phase autotransformers, however verification with
measurements would be desirable,
— The algorithm allows assessing the levels of harmonic
distortion due to the aggregated effect of power
autotransformers on the harmonic levels in the Danish
transmission grid.

7. FUTURE WORK

That the most important future task shall be to verify
the implemented algorithm with precise measurements on full-size power autotransformers.

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