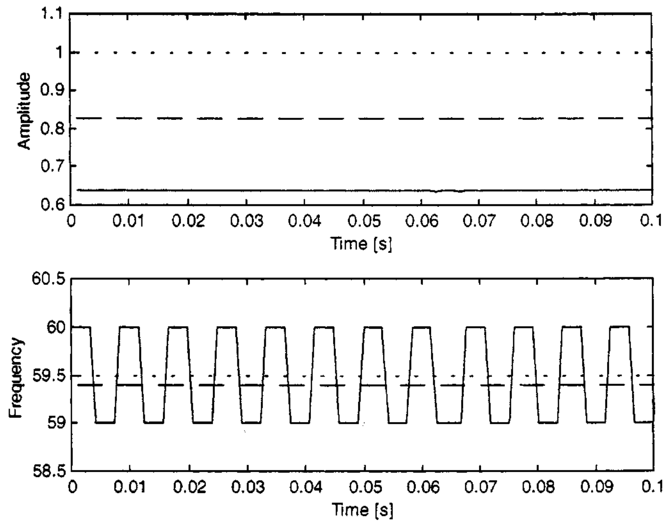

Fig. A. Estimated amplitude and frequency using DFT for the test signal of Fig. 1: $T_w = 1$ s (solid), $T_w = 3.33$ s (dashed) and $T_w = 10$ s (dotted).

3) What is the DFT based algorithm adopted for the frequency and phase detection used for comparison with the proposed algorithm? Is it simply the search of the component with the maximum amplitude or is it more elaborated?
4) The test on DFT shown in Fig. 1 was reproduced using Matlab (see Fig. A) assigning the parameter T_w different values, but no oscillations were observed (except for $T_w = 1$ s). The amplitude and the phase of the test signal with noninteger frequency are tracked correctly only at the higher frequency resolution of 0.1 Hz. Could the author explain briefly the behavior of the algorithm and the sensitivity of the algorithms to window length T_w?

Closure to Discussion of "A Precise Calculation of Power System Frequency and Phasor"

Jun-Zhe Yang and Chih-Wen Liu

We thank Dr. Marisotti for his interest in our paper.[1]

The first question: In DFT and proposed algorithm, it is not necessary to set the window $T_w = 10$ s for frequency resolution of 0.1 Hz. According to Sampling theorem, the range of estimated frequency is from 0 to half of the sampling frequency.

The second question: It is also not necessary for SDFT$_n$ to know the frequency of nonintegral harmonic in *priori*. If the fundamental frequency is much closer 60 Hz than nonintegral harmonic, then we can easily discriminate the frequency from fundamental and nonintegral harmonics. But SDFT$_n$ is not suitable for flicker, it only take one nonintegral harmonic into consideration. For flicker, we have another algorithm, which also belongs to SDFT family. We will present it in a later paper.

The third question: The DFT based algorithm, which we used to compare with SDFT, is from [11]. We didn't test other modified DFT methods because we didn't find a modified DFT method, which can take leakage error into consideration. Meanwhile, a basic DFT with recursive computing can serve as a fair comparison with respect to CPU time and accuracy.

The fourth question: We believe that you can observe oscillations from DFT when system frequency deviates from nominal frequency. The sensitivity of the algorithms to window length will be published in coming Summer Meeting.

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[1]J.-Z. Yang and C.-W. Liu, *IEEE Trans. Power Delivery*, vol. 15, no. 2, pp. 494–499, April 2000.

Discussion of "New Power Transformer Model for the Calculation of Electromagnetic Resonant Transient Phenomena Including Frequency-Dependent Losses"

Bryce L. Hesterman

Although the new model is presented in the context of a utility-grade transformer, I have applied it to a two-winding high-frequency transformer with a gapped ferrite core.

The authors obtained the values $\boldsymbol{R}_f$, $\boldsymbol{L}_f$ and $\boldsymbol{M}$ by simultaneously fitting the expressions for $\boldsymbol{R}(\omega)$ and $\boldsymbol{L}(\omega)$ in (28) to measured impedance data, while using the second assumption of the paper, which is that all of the auxiliary circuits in each group k have the same self inductance and resistance values. I found that this is not sufficient information to extract a unique set of coefficients. There is an extra degree of freedom so that the values of the elements of, $\boldsymbol{R}_f$ $\boldsymbol{L}_f$ and $\boldsymbol{M}$ may be arbitrarily scaled by a factor a as shown below.

$$\boldsymbol{R}(\omega) \approx \boldsymbol{R}_b + \sum_{k=1}^{r} \frac{\omega^2 (a\boldsymbol{R}_{fk})}{(a\boldsymbol{R}_{fk})^2 + \omega^2 (a\boldsymbol{L}_{fk})^2} \left(\sqrt{a}\boldsymbol{M}_k\right)^2 \tag{29}$$

$$\boldsymbol{L}(\omega) \approx \boldsymbol{L}_b - \sum_{k=1}^{r} \frac{\omega^2 (a\boldsymbol{L}_{fk})}{(a\boldsymbol{R}_{fk})^2 + \omega^2 (a\boldsymbol{L}_{fk})^2} \left(\sqrt{a}\boldsymbol{M}_k\right)^2 \tag{30}$$

I came to this conclusion after considering why it was that my curve-fitting routine would produce different sets of coefficients depending on what my initial guess values were, and that many of these sets produced equally good fits.

After realizing that I had this extra degree of freedom, I decided that, instead of following the second assumption, I would set the value of the inductance of each auxiliary winding to be equal to the element of $\boldsymbol{L}_b$ that corresponds to the self inductance of the main winding to which the auxiliary winding is coupled. I believe that this should produce better numerical results than those produced using the authors' assumption when the inductance values of the main windings have values that are widely different in magnitude.

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What I am suggesting here corresponds in effect to a model which uses a lossless transformer having n main windings with self and mutual inductances equal to the respective values of $\boldsymbol{L}_b$. The lossless transformer also has n sets of r auxiliary windings, with each one having a 1 : 1 turns ratio with respect to the main winding with which it is coupled.

Although the authors' curve fitting method simultaneously utilized both the real and imaginary parts of the measured impedances, I found that I could obtain the values of $\boldsymbol{R}_f$, $\boldsymbol{L}_f$, and $\boldsymbol{M}$ simply by fitting the resistance measurements to the expression for $\boldsymbol{R}(\omega)$ in (28). The calculated values of the inductances matched the measured values closely, except at high frequencies, where winding capacitances affected the measurements. (At high frequencies, the measured inductances started to rise with increasing frequency, while the calculated inductances continued to fall.) The ability to determine the values of $\boldsymbol{R}_f$ and $\boldsymbol{M}$ from the resistance measurements alone when given arbitrarily selected values of $\boldsymbol{L}_f$ suggests that when the three component values of each auxiliary circuit are fit to measured data, the three component values are not linearly independent.

As I understand the paper,[1] only one auxiliary winding from each group k is coupled to a particular main winding. This makes the submatrices $\boldsymbol{M}_k$ diagonal. Having diagonal $\boldsymbol{M}_k$ matrices allows the model to correctly represent the winding impedances measured one at a time, but the model cannot represent the effects of the interactions among the windings on the winding resistances. I verified this by comparing the leakage impedance of the transformer measured at the primary terminals when the secondary terminals are shorted together with the leakage impedance that I computed using (6). I performed this comparison at each frequency that I had used in my curve-fitting routine.

Before discussing the results of these comparisons, let me explain how the leakage impedances were calculated. The values of $\boldsymbol{u}_b$ were assigned values that correspond to having the primary winding driven with one volt, and the secondary winding shorted, so that (6) becomes:

$$\begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} \boldsymbol{Z}_b & j\omega\boldsymbol{M} \\ j\omega\boldsymbol{M}^T & \boldsymbol{Z}_f \end{bmatrix} \cdot \begin{bmatrix} i_{b1} \\ i_{b2} \\ \boldsymbol{i}_f \end{bmatrix}. \tag{31}$$

The primary current i_{b1} is computed by:

$$\begin{bmatrix} i_{b1} \\ i_{b2} \\ \boldsymbol{i}_f \end{bmatrix} = \begin{bmatrix} \boldsymbol{Z}_b & j\omega\boldsymbol{M} \\ j\omega\boldsymbol{M}^T & \boldsymbol{Z}_f \end{bmatrix}^{-1} \cdot \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix}. \tag{32}$$

The computed leakage resistance and the leakage inductance values measured at the primary terminals are found by:

$$R_{leak} = \mathrm{Re}\left[\frac{1}{i_{b1}}\right]. \tag{33}$$

$$L_{leak} = \mathrm{Im}\left[\frac{1}{\omega i_{b1}}\right]. \tag{34}$$

As I expected, the computed leakage resistance was approximately equal to the sum of the primary resistance plus the reflected secondary resistance. The measured leakage resistance, however, was much smaller. The reason for this is that the magnetic field that is produced in the winding space by the primary current is largely cancelled by the magnetic field produced by the secondary current. Consequently, the total power dissipated in both windings for a given primary current must be lower when the secondary is shorted than it is when the secondary is open. Another way of looking at this is to say that the mutual resistance terms of the transformer impedance matrix $\boldsymbol{Z}$ are positive [1]. It should be noted that there are winding arrangements in which some of the mutual resistance terms are negative [1]. My experience with high frequency transformers indicates that transformer models must account for the mutual resistance terms in the transformer impedance matrix to properly model winding losses. Before reading this paper, however, I did not see an easy way to model the effect of frequency-dependent mutual resistances using fixed circuit elements.

When the off-diagonal elements of the $\boldsymbol{M}_k$ submatrices are included in a model for two windings with two groups of auxiliary circuits, (28) becomes:

$$\begin{aligned} R(\omega)_\lambda \approx R_{b\lambda} &+ \sum_{k=1}^{r} \frac{\omega^2(R_{fp})}{(R_{fp})^2 + \omega^2(L_{fp})^2}\,(M_{k_{\lambda\lambda}})^2 \\ &+ \sum_{k=1}^{r} \frac{\omega^2(R_{fq})}{(R_{fq})^2 + \omega^2(L_{fq})^2}\,(M_{k_{12}})^2 \end{aligned} \tag{35}$$

$$\begin{aligned} L(\omega)_\lambda \approx L_{b\lambda} &- \sum_{k=1}^{r} \frac{\omega^2(L_{fp})}{(R_{fp})^2 + \omega^2(L_{fp})^2}(M_{k_{\lambda\lambda}})^2 \\ &- \sum_{k=1}^{r} \frac{\omega^2(L_{fq})}{(L_{fq})^2 + \omega^2(L_{fq})^2}\,(M_{k_{12}})^2 \end{aligned} \tag{36}$$

where

$\lambda \quad = 1, 2;$

$p \quad = r(\lambda - 1) + k$; and

$q \quad = r(2 - \lambda) + k.$

The $\boldsymbol{R}_f$ and $\boldsymbol{L}_f$ matrices are diagonal with four unique diagonal elements. The diagonal elements of the submatrices $\boldsymbol{M}_k$ are denoted as $M_{k_{\lambda\lambda}}$, and the off-diagonal elements are denoted as $M_{k_{12}}$.

I obtained the values for $\boldsymbol{R}_f$, $\boldsymbol{L}_f$, and $\boldsymbol{M}$ by using a curve-fitting routine that including (32), (33) and (35). As before, when the off-diagonal elements of the submatrices $\boldsymbol{M}_k$ were neglected, I did not need to have the expression for inductance, (36), included in the curve-fitting routine.

Extending this method to more than two windings would mean that each auxiliary winding would be coupled to each main winding, but the auxiliary windings would not be coupled to each other. I have not tried to implement this, but I presume that I could set up an appropriate curve-fitting routine to simultaneously determine the coefficients for all of the windings.

This extension of the authors' model has the potential to predict the frequency-dependant losses of transformers more accurately, and I intend to further investigate this idea.

Another item worth mentioning is that I was able to model the transformer from do to 200 kHz with just two groups of auxiliary circuits. I found that adding a third group produced no benefits over this frequency range, but that the third group was useful when the frequency range was extended. There should be some method for determining how many auxiliary groups are necessary to extend the ac resistance curve a certain range above the inflection point where skin effect becomes predominant over the proximity effect. Do the authors have any suggestions?

I have found that there is one potential difficulty that can occur with the model presented by the authors, and also with the modifications that I have suggested. For a given set of self inductances, there is a space of allowed values for the mutual inductances in a system of coupled windings that will ensure that the system is stable. To determine the

[1]E. E. Mombello and K. Möller, *IEEE Trans. Power Delivery*, vol. 15, no. 1, pp. 167–174, January 2000.

bounds of this mutual inductance space, we can construct an inductance matrix for the system of the main and auxiliary windings as follows:

$$\boldsymbol{L}_{sys} = \begin{bmatrix} \boldsymbol{L}_b & \boldsymbol{M}_1 & \boldsymbol{M}_2 & \cdots & \boldsymbol{M}_r \\ \boldsymbol{M}_1 & \boldsymbol{L}_{f1} & 0 & \cdots & 0 \\ \boldsymbol{M}_2 & 0 & \boldsymbol{L}_{f2} & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ \boldsymbol{M}_r & 0 & 0 & \cdots & \boldsymbol{L}_{fr} \end{bmatrix}. \tag{37}$$

The inductance matrix $\boldsymbol{L}_{sys}$ must be positive definite to ensure stability of the system [2]. There are several tests for checking the positive definiteness of a matrix [3]. One test that is easy to implement with general-purpose mathematical computation programs is to compute the eigenvalues of $\boldsymbol{L}_{sys}$. A matrix is positive definite if and only if all of its eigenvalues are positive.

I found that when implementing both the authors' model and my variations, some sets of initial conditions for my curve-fitting routine led to sets of coefficient matrices that produced fits that were fairly close, but not as good could as the best fits. Some of these mediocre coefficient sets produced non positive-definite system inductance matrices. Consequently, I recommend checking the results of a curve-fit for positive definiteness of $\boldsymbol{L}_{sys}$ before using the model. Having a non positive-definite model may cause time-domain simulations to have convergence problems for no apparent reason, and frequency-domain simulations may produce winding impedances with negative inductance values.

I have not yet determined a method for finding good initial guess values for $\boldsymbol{R}_f$, $\boldsymbol{L}_f$, and $\boldsymbol{M}$ to use in my curve-fitting routines, and I would welcome any suggestions from the authors.

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Closure to Discussion of "New Power Transformer Model for the Calculation of Electromagnetic Resonant Transient Phenomena Including Frequency-Dependent Losses"

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The authors wish to thank Mr. Hesterman for his interest and comments on our paper.[1]

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[1] E. E. Mombello and K. Möller, *IEEE Trans. Power Delivery*, vol. 15, no. 1, pp. 167–174, January 2000.

Before discussing Mr. Hesterman comments, it should be first noted that in our model each main inductive branch is magnetically coupled with all auxiliary circuits, as stated in our paper (page 2, second column, last bullet), and not only to a particular one. This means that all $\boldsymbol{M_k}$ matrices are full matrices. This was considered in this manner so that the model can properly account for the winding losses.

General equation (27) can be written for the elements of $\boldsymbol{R}_{eq}$ and $\boldsymbol{L}_{eq}$ as:

$$R_{eq_{ij}} = R_{bij} + \sum_{k=l}^{k=r} \frac{R_{fk}\omega^2}{R_{fk}^2 + L_{fk}^2\omega^2} M_{kij}^2 \tag{38}$$

$$L_{eq_{ij}} = L_{bij} - \sum_{k=l}^{k=r} \frac{L_{fk}\omega^2}{R_{fk}^2 + L_{fk}^2\omega^2} M_{kij}^2 \tag{39}$$

It should be noted that in the discusser's equations (29) and (30), $\boldsymbol{R}(\omega)$, $\boldsymbol{R_b}$ and $\boldsymbol{M_k}$ are matrices and R_{fk} and L_{fk} are scalars.

In response to the first comment about the available degrees of freedom, the discusser is correct. This topic is treated extensively in [1], but there was no enough space in the paper for including all details.

The optimization process was divided in two stages. The first stage comprises the determination of the circuit parameters R_{bij}, L_{bij}, R_{fk} and L_{fk} of one of the elements of matrix $\boldsymbol{Z_{eq}}$ (Z_{eq11} was chosen for this purpose).

In the first stage, the optimization is to be done considering that:

$$N_{k_{11}} = 1 \qquad k = 1, \cdots, r \tag{40}$$

where

$$\boldsymbol{N_k} = \boldsymbol{M_k^2} \tag{41}$$

The parameters to be optimized in this stage are not really R_{fk} and L_{fk}, but

$$kr_k = \frac{1}{R_{fk}} \qquad \tau_k = \frac{L_{fk}}{R_{fk}} \tag{42}$$

Once the values of kr_k and τ_k have been determined, they are to be considered definitive values and valid for all other elements of the matrix. This can be assumed, due to the fact that all impedances $Z_{ij}(\omega)$ have very similar frequency-dependencies. The values of R_{b11} and L_{b11} are only valid for the element Z_{eq11}. The determination of the parameters for Z_{eq11} is then complete. It is not convenient to determine R_{fk} and L_{fk} from kr_k and τ_k at this stage as it will be seen below. The values assumed for N_{k11} from (40) are provisional (normalized) ones, and they will be modified at the end of the second stage.

The second stage comprises the determination of all elements (excluding the elements with subindices 1, 1) of the matrices $\boldsymbol{L_b}$, $\boldsymbol{R_b}$ and $\boldsymbol{N_k}$. In this stage the values kr_k and τ_k are considered as constants. Equations (38) and (39) are now optimized for R_{bij}, L_{bij} and N_{kij} $(k = 1, \cdots, r)$ and for each element separately (i.e., for each set of values ij).

It should be noted that there is not a unique solution for the problem. As the discusser states, there are only two independent variables in equation (28) in order to define the circuit parameters, and they are the magnitude $(kr_k \boldsymbol{M_k^2})$ and the parameter τ_k, as one possible combination.

Once the complete optimization has been done, the circuit parameters should be calculated so as the matrix $\boldsymbol{L_{sys}}$ becomes positive definite. We have the possibility to vary the factor a in equations (29) and

(30) in order to change the values of the matrices $\boldsymbol{M}_k$ and $\boldsymbol{L}_{fk}$. Using a as reduction factor, it can be seen that the values of the elements of both sets of matrices will be smaller, due to the multiplication by $\sqrt{a}$ and a respectively. If matrix $\boldsymbol{L}_{sys}$ is not positive definite for a certain set of values, using the reduction factor it will become positive definite, assuming that matrices $\boldsymbol{L}_{fk}$ and $\boldsymbol{L}_b$ are positive definite matrices. When $\boldsymbol{L}_{sys}$ meets this condition, the definitive values of $\boldsymbol{M_k}$, R_{fk} and L_{fk} can be finally calculated.

The discusser states that, *instead of following the second assumption, he sets the value of the inductance of each auxiliary winding to be equal to the element of $\boldsymbol{L}_b$ that corresponds to the self inductance of the main winding to which the auxiliary winding is coupled.* Actually, each auxiliary winding is coupled with all main windings. If I understand correctly, the discusser's assumption should mean that the elements of a given diagonal submatrix $\boldsymbol{G}_{fk}$ (or $\boldsymbol{B}_{fk}$) are no longer equal to each other, which implies that equations (24) and (25) are also no longer valid and, for example, the product $\boldsymbol{M}\boldsymbol{G}_f\boldsymbol{M}^{\boldsymbol{T}}$ would not have a simple form as in (24). The second assumption is necessary to keep the validity of Eq. (28).

Regarding discusser's equations (35) and (36), they can not be a derivation from (28), since all the variables excepting R_{fk} and L_{fk} should have a double subindex (i.e., ij). For the case of a model with two main winding sections, (35) and (36) will give only two resistances and two inductances, while (28) will give four.

Finally, it should be noted that after the second stage of optimization, the elements of matrix $\boldsymbol{N_k}$ are obtained. The matrix $\boldsymbol{M_k}$ must be determined from $\boldsymbol{N_k}$. Consequently, matrix $\boldsymbol{N_k}$ must be positive definite, otherwise complex elements are calculated for matrix $\boldsymbol{M_k}$. Fortunately, $\boldsymbol{N_k}$ is either positive definite or almost positive definite. In the second case only very small corrections of some elements produce the matrix to be positive definite [1].

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Discussion of "A Current Transformer Model Based on the Jiles–Atherton Theory of Ferromagnetic Hysteresis"

W. C. Kotheimer

The authors of this paper[1] have identified an apparent limitation of the Jiles–Atherton algorithm using the Langevin function when used to model the core of a Ct subjected to high fault currents in the presence of a significant amount of remanence. Figs. 1 and 2 show a comparison of the B-H loops produced by test and by simulation using the JA algorithm. In Fig. 2 the region of early saturation of the B-H loop isidentified as not well modeled by the JA algorithm, whereas the linear and saturated regions are said to be reasonably accurate. Similarly, Figs. 6 and 7 show the B-H loops for test and for simulation using the JA algorithm with an improved anhysteretic function.

Close inspection of Figs. 2 and 7 shows that the maximum magnetic field strength for the test B-H loops is about 630 At/m whereas the maximum for the simulated B-H loops is off scale and must be well over 800 At/m. A better comparison would be to use B-H loops having the same maximum magnetic field strength excursions. Have the authors made this comparison?

In the maximum saturated region of the B-H loops shown in Fig. 2 there is a noticeable difference in the slopes of the test and simulated cases. Since this slope is a measure of the incremental permeability of the Ct core, it seems likely that it also would influence the behavior of the Ct when operating in the saturated region. In Fig. 2 the indicated permeability for the test case is about 139 whereas for the simulation it is about 17. This may be another reason why the test Ct performed better when in saturation. Have the authors considered this?

The foregoing comments in no way diminish the significant advancement the authors have made by introducing the improved anhysteretic function for the JA algorithm.

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[1]U. D. Annakkage *et al.*, *IEEE Trans. Power Delivery*, vol. 15, no. 1, pp. 57–61, January 2000.

Closure to Discussion of "A Current Transformer Model Based on the Jiles–Atherton Theory of Ferromagnetic Hysteresis"

U. D. Annakkage, P. G. McLaren, E. Dirks, R. P. Jayasinghe, and A. D. Parker

Both observations made by Mr. Kotheimer are quite correct. With regard to the maximum field strength used in the comparison of the measured and simulated B-H loops the authors should have used the same maximum field strength. The two results were recorded in different places at different times and this point was missed. It is relatively simple to alter the simulation result to have the same field strength as the practical test and this will be done for future reference. Since the new anhysteretic function gave good comparisons for the CT secondary currents the authors were content to leave the B-H loop comparison in an incomplete state at the time the work was carried out. The priority at the time was to move on to multi-CT applications in order to equip the contractors real time digital simulator with algorithms to cover all the typical CT interconnections in differential current protection schemes [1]. A Ph.D. research project is now looking at refinements to the anhysteretic curve and other relevant effects to improve the B-H loop comparisons.

In the heavy saturation case, Fig. 4 of the paper, the B-H trajectory moves rapidly through the shoulder area and any mismatches in this region have little effect on the secondary current. Once into the heavily saturated region the difference in incremental permeability pointed out

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