$$S_{re} = \begin{bmatrix} 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 2 & 0 & 2 & -2 & 0 & 4 & -4 & 0 \\ 1 & -2 & -1 & 1 & 4 & -6 & 2 & 0 \\ 1 & 2 & -1 & 1 & 0 & -2 & 6 & -4 \\ 1 & -1 & -4 & 0 & -2 & 2 & 0 & 0 \\ 0 & -1 & 3 & 1 & -2 & 0 & 2 & 0 \\ 0 & 1 & -1 & -3 & 0 & -2 & 0 & 2 \\ 1 & 1 & 0 & 4 & 0 & 0 & -2 & 2 \end{bmatrix}$$

$$S_{im} = \begin{bmatrix} 0.5039 & 0.2500 & 0.1250 & 0.1250 & 0.0625 & 0.0625 & 0.0625 & 0.0625 \\ -0.2500 & 0.0039 & -0.1250 & 0.1250 & -0.0625 & -0.0625 & 0.0625 & 0.0625 \\ -0.0625 & 0.0625 & 0.0039 & 0 & -0.0625 & 0.0625 & 0 & 0 \\ -0.0625 & -0.0625 & 0 & 0.0039 & 0 & 0 & -0.0625 & 0.0625 \\ -0.0156 & 0.0156 & 0.0313 & 0 & 0.0039 & 0 & 0 & 0 \\ -0.0156 & 0.0156 & -0.0313 & 0 & 0 & 0.0039 & 0 & 0 \\ -0.0156 & -0.0156 & 0 & 0.0313 & 0 & 0 & 0.0039 & 0 \\ -0.0156 & -0.0156 & 0 & -0.0313 & 0 & 0 & 0 & 0.0039 \end{bmatrix}$$

where:

$$S_{re} = \begin{bmatrix} \beta_{1,1} & \beta_{1,2} & \cdots & \beta_{1,m} \\ \beta_{2,1} & \beta_{2,2} & \cdots & \beta_{2,m} \\ \vdots & \vdots & \ddots & \vdots \\ \beta_{m,1} & \beta_{m,2} & \cdots & \beta_{m,m} \end{bmatrix},$$

$$I = \begin{bmatrix} 1 & & & \\ & 1 & & \\ & & \ddots & \\ & & & 1 \end{bmatrix},$$

$$b = \begin{bmatrix} d_{11} - d_{21} \\ d_{12} - d_{22} \\ \vdots \\ d_{1m} - d_{2m} \end{bmatrix}, \quad \text{and} \quad A = \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_m \end{bmatrix}.$$

Note that the above equation is a real matrix equation. Its solution provides the unknown coefficients $[a]$ of equation (1), and therefore the solution $x(t)$.

Discussion on "Analytical Approach to Electric Circuits Containing Saturating Ferromagnetic Coils"

N. Janssens and Th. Van Craenenbroeck

In the past, numerous publications made use of the harmonic balance method to study the periodic evolutions of nonlinear circuits driven by sinusoidal voltage sources. A key point of the paper[1] of X.-F. Wang, X.-L. Wang and J.-H. Wang is to consider separately the linear and the nonlinear parts of the system, based on the fact that the different harmonics behave independently in the linear part and are coupled by the nonlinear elements. Another way to express this is to represent the linear part by its Thévenin equivalent for each frequency considered [1], [2]. These Thévenin equivalents are 1-port circuits in case of a unique nonlinear element or n-port circuits if there are n nonlinear elements. So, the size of the algebraic system to solve is no more related to the order of the linear part of the system (number of reactive elements) but only to the number of nonlinear elements and the number of harmonic components considered.

It is possible to obtain directly the domain of existence and stability of the various types of solutions in some parameters space by adding an equation expressing that the determinant of a Jacobian is equal to zero. This allows to find loci of bifurcations of the saddle-node type or of the period doubling type [1], [3], [4].

It is also worth mentioning that not all solutions are described by a set of odd harmonics or subharmonics. Periodic regimes with even components may exist [3], [5]. Quasiperiodic oscillations may also appear, for instance in three phase circuits. These situations can be handled by the harmonic balance method by making reasonable assumptions [4].

If a good correspondence between computed and measured values is desired, care must be taken regarding the modeling. A first important item is the representation of the magnetic characteristic. From our experience, in general, a representation of the flux—current relation by polynomials of limited order (e.g., 7) does not give an accurate model on the whole range, especially when the saturation knee is sharp. For instance, in the paper, the first order coefficient a_1 (linear part) is negative. We found it better to use a parabolic spline in order to reproduce the magnetization characteristic precisely. It was proven necessary to have this quality of precision to obtain quantitative results in accordance to field tests.

From a practical point of view, the algebraic approximation used in (26) is limited to systems with a relatively low number of harmonics and to models that can be described by a polynomial of a limited order. The calculation of the coefficients at the right-hand side in (26) becomes a tedious task if these conditions are not fulfilled. An alternative and more general approach is to evaluate the nonlinear characteristic in the time-domain, combined with a Fourier transform to obtain the harmonic components of the corresponding variable [1], [3], [4].

Another important modeling aspect is the representation of the core losses. Fig. 5 of the paper shows an evolution of B_1 with relative extremes. This shows the presence of secondary hysteresis cycles. A con-

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N. Janssens is with CPTE, Belgium.
Th. Van Craenenbroeck is with K.U. Leuven, Belgium.
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[1]X.-F. Wang *et al.*, *IEEE Trans. on Power Delivery*, vol. 15, no. 2, pp. 697–702, April 2000.

losses in such circumstances. It is possible to handle this aspect by introducing, at each iteration of the Newton–Raphson computation, an equivalent conductance that is function of all the extremes of the flux periodic evolution. The eddy current losses are a function of the rms value of the voltage across the magnetizing branch and must be considered separately [3], [5].

Let us also mention that the model described in the paper on Fig. 1 does not consider a series resistance on the primary side. This aspect introduces additional terms in the equations, e.g., in (16), related in a complex manner to the state variables. The introduction of these resistances in the Thévenin equivalents is straightforward and does not burden significantly the computation.

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Discussion of "Measured Transformer Derating and Comparison with Harmonic Loss Factor (F_{HL}) Approach"

T. Batan

An innovative procedure for the measurement of the losses is presented in [1]. The derating measurements of transformers due to harmonics of the paper[1] at hand are based on [1] taking into account frequency dependent skin and proximity effects in the windings and the core. The measurement of iron-core and copper losses of single-phase transformers is important in particular for transformers feeding nonlinear loads. Reference [1] devises a new digital data-acquisition method for the separate measurement of iron-core and copper losses of transformers under any full or partial-load condition. The accuracy requirements of the instruments (voltage, current sensors, volt and current meters) employed are addressed. A 25 kVA, 7200/240 V single-phase transformer has been loaded with either three diode-bridge rectifiers and/or a thyristor-bridge rectifier at varying total harmonic voltage and current distortions such that at maximum permissible apparent power the losses are identical to the rated losses. The maximum errors of losses are acceptably low if voltage (dividers) and current (shunts) sensors with 2% and digital instruments with 0.1% errors are used. The derating under nonlinear loads is measured as a function of dominant 3rd and 5th current harmonics. The influence of voltage distortion is studied. The total harmonic distortion of the current (THD_i) was varied greatly from 0 to 90% and the resulting voltage THD_v values were below 6% for the given power system impedance.

The doctoral thesis [4] illustrates how the eddy currents within transformers can be calculated from field plots. The major conclusion of the above-mentioned thesis is that the eddy-current distribution within transformer windings is indirectly dependent upon the saturation of the transformer core [1, Fig. 2]. The authors have measured apparent and real powers. Have the authors also measured reactive and distortion powers?

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T. Batan is with the University of Colorado, Campus Box 425, Boulder, CO 80309.
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Discussion of "Measured Transformer Derating and Comparison with Harmonic Loss Factor (F_{HL}) Approach"

D. J. Roesler

In the above paper,[1] the authors show that the apparent power derating calculation of a transformer can be either based on the K-factor as described in reference 1 or on the harmonic loss factor F_{HL} provided the eddy-current loss coefficient P_{EC-R} is known. From this paper one can conclude that the F_{HL} approach generates a somewhat smaller reduction of the apparent power rating (as is depicted in Fig. 2) than the K-factor approach (as illustrated in Fig. 4 of reference [1]). Is this inequality always true? How large is this inequality?

For commonly occurring THD_i values of about 20% the apparent power derating (Fig. 2) and the reduction of real power capability (Fig. 3) are quite small, while for THD_i values of 90% the latter is quite large. It appears, therefore, that it might be appropriate to define transformer derating in terms of the reduction in real power. Can the authors comment on this question? How could the method of this paper be applied to very large transformers?

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D. J. Roesler was with the Department of Energy, Washington, DC. He is now at 6311 Fort Hunt Road, Alexandria, VA 22307.
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