Fast Initialization for Transient Calculations with Non-Sinusoidal Steady State

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Abstract— This paper presents an analytical framework to represent non-linear and switched devices in efficient harmonic initialization procedures for the time domain simulation of electromagnetic transients. This modular approach, associated with the fully analytical modeling of the non-linear components, allows implementation of several different solution techniques, ranging from a simple fixed point (Gauss-Seidel) iteration method to a full Newton-Raphson procedure.

Keywords— Periodic steady state, Non-linear systems, Switched devices.

I. Introduction

The accurate determination of the initial condition for a time-domain electromagnetic transients study can be a difficult and time-consuming task. In general, this initial condition corresponds to a periodic steady state of the system, under pre-defined operating conditions such as loading and topology.

The greatest computational challenge comes from the presence of non-linear or switched components. In order to obtain the periodic steady state of such a system, several approaches based on time-domain and/or frequency-domain modeling of the system have been presented [1]-[7]. A classification of these methods is presented in Section II.

To solve this problem, it is sound practice to subdivide the system into a linear part (LP), consisting of all conventional linear components, and a nonlinear part (NLP) formed by the nonlinear and switched devices.

This division can lead to a modular approach to represent the NLP, since the current injections of each component can be analytically obtained from its terminal voltages. As seen from the LP, these devices are voltage-dependent harmonic current sources. These currents are then injected into the LP of the system to obtain an update to the voltages. This is essentially a fixed-point iteration process with fast convergence if the injected currents are not too large.

This modular approach is adopted in this paper. The modeling of a simple non-linear component and a switched device is presented in Section III.

The issues of convergence and robustness of the different methods are illustrated through two simple example systems, i.e. a non-linear reactor and a thyristor-controlled reactor (TCR). These results are presented in Section IV.

II. Classification of the Methods

A simple classification to sort out the methods for obtaining the periodic steady state of a non-linear power system, based on the nature of the solution, is given in the following. Table I gives an overview of this classification.

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Table I – Classification of the methods

- **— TIME DOMAIN METHODS**
- **— FREQUENCY DOMAIN METHODS**
- **— HYBRID METHODS**
  - **— Sequential**
    - *— Full current injection*
    - *— Non-linear Norton equivalent*
  - **— Simultaneous**
    - *— Linear Norton equivalent*
    - *— Complex decoupled Newton-Raphson*
    - *— Decoupled Newton-Raphson*
    - *— Newton-Raphson*

I. TIME DOMAIN METHODS

This is, perhaps, the simplest approach. One can simply let the time domain simulation run until steady state is reached. Its major drawback is the long simulation time required if less than a perfect initial guess is used, especially if the system has low damping.

II. FREQUENCY DOMAIN METHODS

This is the classical approach to solve linear systems, where superposition leads to decoupled frequencies. Modeling of non-linear components directly in frequency domain can be quite difficult. Worse than that, these models produce coupling between the frequencies, requiring the simultaneous solution for all frequencies.

III. HYBRID METHODS

The hybrid methods are usually based on a nodal representation of the LP of the system. These methods rely on the frequency domain solution of the LP and use time domain models (or frequency domain models based on time domain responses) to deal with the NLP. The nonlinear components are represented as current injections at their terminal nodes. These injected currents can be usually modeled as non-linear functions of the voltages.

For each frequency h, one should solve

$$Y_h v_h = i_h \qquad (1)$$

where Y_h is the nodal admittance matrix associated with the LP of the system, calculated at frequency $h\omega$, v_h is the vector of nodal voltages and i_h is the vector of nodal current injections from the non-linear components.

The periodic steady state of the system is obtained when (1) is satisfied for every frequency of interest. The hybrid methods can be classified according to the approach taken to solve the problem:

III.A. Sequential

In this approach, the LP and the NLP are solved alternately. The LP is solved frequency by frequency and only the NLP solution has to deal with the frequency coupling.

The harmonic current injections for each non-linear device in the system are calculated from the knowledge of the harmonic nodal voltages at the previous solution of the LP. If the direct calculation of these currents is not feasible, an iterative procedure may be required.

The harmonic currents for all non-linear devices and independent sources should be assembled to obtain the vectors i_h, to be used in the next solution of the LP. The following procedures can be devised:

III.A.1. Full current injection – The nodal current vectors are directly used in the solution of (1) to obtain new values for the voltages. This approach is known as fixed-point iteration or Gauss method. It requires, at each iteration, the computation of the terminal currents, for each device in the NLP. These currents are obtained from the nodal voltages calculated at the previous iteration.

The implementation of this method is very simple, but its convergence is greatly affected by the relative magnitude of the non-linear current injection.

III.A.2. Non-linear Norton equivalent – To improve convergence, the non-linear current injection of each non-linear device can be written as a sum of a linear term $Y_h^{dev} v_h^{dev}$ and a correction term $i_h'^{\,dev}$ as

$$i_h^{dev} = i_h'^{\,dev} + Y_h^{dev} v_h^{dev} \qquad (2)$$

where Y_h^{dev} is a diagonal admittance matrix built using approximations for the harmonic admittances of the non-linear device.

This idea leads to a Norton equivalent, as shown in Figure 1. The harmonic vector $i_h'^{\,dev}$ reflects the non-linearity of the device.

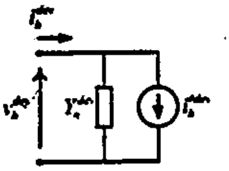

Fig. 1. Harmonic Norton equivalent

The harmonic admittance Y_h^{dev} can be included in Y_h and, therefore, part of the non-linear current injection is transferred to the LP. This can improve convergence if the modified non-linear current injection $i_h'^{\,dev}$ has smaller components than i_h^{dev}.

III.B. Simultaneous

The simultaneous solution of the LP and NLP requires the simultaneous solution of (1) for every frequency. This can be expressed as

$$Yv - i(v) = 0 \qquad (3)$$

where Y is a block diagonal matrix containing the Y_h matrices for all frequencies. The vectors v and i are built in a similar way, containing vectors v_h and i_h, respectively.

The coupling between frequencies is implicit to the non-linear currents and the harmonic components of these currents are, in general, function of all the harmonics of the voltage.

Equation (3) represents a set of non-linear algebraic equations and the methods available for the solution of this problem are similar to those applied to solve a conventional power flow problem. The solution of this problem can be divided in several categories:

III.B.1. Newton-Raphson – This is the only linearized method that can ultimately achieve quadratic convergence. It requires, on the other hand, the use of (3) in its augmented real form to allow linearization. If the Jacobian matrix associated with the real form of (3) can be determined (analytically or numerically), it will lead to a set of $2 \times n_h \times n_b$ linear equations due to the coupling between frequencies, where n_h is the number of harmonic frequencies considered and n_b is the number of nodes in the system.

III.B.2. Decoupled Newton-Raphson – The Jacobian matrix associated with the augmented real form of (3) has frequency coupling entries that could be disregarded to yield n_h sets of $2 \times n_b$ linear equations. This approach can lead to faster iterations, especially if parallel computing is applied, but will require more iterations to achieve convergence, when compared to the full Newton-Raphson method.

III.B.3. Complex decoupled Newton-Raphson – The need to use the augmented real form of (3) arose from the fact that the linearization of the non-linear models led to general 2×2 linear operators as shown in (4), which cannot be expressed in terms of complex numbers.

$$\begin{bmatrix} a & b \\ c & d \end{bmatrix} \begin{bmatrix} \Delta x' \\ \Delta x'' \end{bmatrix} \qquad (4)$$

To use complex numbers, these linear operators should be restricted to a specific structure:

$$\begin{bmatrix} a & -b \\ b & a \end{bmatrix} \begin{bmatrix} \Delta x' \\ \Delta x'' \end{bmatrix} \Rightarrow (a + jb)(\Delta x' + j\Delta x'') \qquad (5)$$

The solution of (3) using complex numbers requires the use of approximations of (4) in the form shown in (5). This leads to n_h sets of n_b complex linear equations to be solved at each iteration. The expected convergence will be largely influenced by this approximation and can be poorer than that achieved with Decoupled Newton-Raphson.

III.B.4. Linear Norton equivalent – Another simplification that can be used is the assumption of a fixed linearized Norton equivalent for the NLP, similar to that shown in Figure 1. In this case, the simplification comes from the use of a constant set of admittances, completely avoiding the need for sophisticated linearized models to be updated at each iteration.

III. Non-Linear Devices

The modeling of the non-linear devices is crucial to the successful application of the modular approach.

The basic idea is to obtain the currents (outputs) as a function of the voltages (inputs). To achieve this goal, the voltages are considered known and are expressed in terms of their Fourier coefficients as

$$v(t) = 2\sum_{h=1}^{n_h} v_h' \cos(h\omega t) - v_h'' \sin(h\omega t) \qquad (6)$$

The steady state time domain characteristics of the non-linear device is used to determine analytic expressions for the currents as a function of the voltages. These time domain functions are periodic and can be expressed as a Fourier series. The Fourier coefficients of the currents are then analytically obtained as definite integrals over one period.

This section presents two examples of such modeling: a non-linear reactor and a thyristor-controlled reactor (TCR).

A. Non-Linear Reactor

Figure 2 shows a non-linear shunt reactor. The current is given by

$$i(t) = K\left[\phi(t) + \phi(t)^3\right] \quad (7)$$

where ϕ is the magnetic flux and K is a constant. The terminal voltage is related to the flux:

$$v(t) = \frac{d\phi(t)}{dt} \quad (8)$$

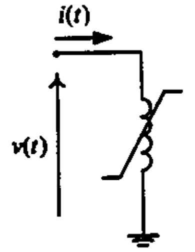

Fig. 2. Non-linear shunt reactor

Assuming that the terminal voltage is given as a Fourier series as in (6), the flux can be written as

$$\varphi(t) = 2\sum_{h=1}^{n_h} \frac{v'_h}{h\omega}\sin(h\omega t) + \frac{v''_h}{h\omega}\cos(h\omega t) \quad (9)$$

Substitution of (9) in (7) leads to the expression of the current, in the time domain, as a function of the applied voltage. This expression can be directly written as a Fourier series, converting the powers and the multiplications of trigonometric functions to their harmonic equivalent expressions. In other words, the Fourier coefficients for the current can be directly determined and they are, in general, functions of the harmonic components of the voltage. This can be expressed as

$$i_h = i'_h(v'_t, v''_t) + j i''_h(v'_t, v''_t) \quad (10)$$

where v'_t and v''_t are the vectors with the real (cos) and imaginary (sin) parts of the harmonic coefficients of the terminal voltage, respectively. These harmonic components of the non-linear reactor current will be injected into the network at the terminal node of the device. If the non-linear reactor is connected to node k, the harmonic contribution of the current, at frequency $h\omega$, should be added to position k of the vector i_h in (1).

The augmented real form of (3) can be written as

$$\begin{bmatrix} G & -B \\ B & G \end{bmatrix}\begin{bmatrix} v' \\ v'' \end{bmatrix} - \begin{bmatrix} i'(v', v'') \\ i''(v', v'') \end{bmatrix} = 0 \quad (11)$$

where $Y = G + jB$. The vectors v' (v'') and i' (i'') contain the real parts (imaginary parts) of the harmonic components of all nodal voltages and current injections, respectively.

The linearization of (11) can be written as

$$\left\{\begin{bmatrix} G & -B \\ B & G \end{bmatrix} - \begin{bmatrix} J_1 & J_2 \\ J_3 & J_4 \end{bmatrix}\right\}\begin{bmatrix} \Delta v' \\ \Delta v'' \end{bmatrix} = -r \quad (12)$$

where the matrices

$$J_1 = \frac{\partial i'}{\partial v'} \quad J_2 = \frac{\partial i'}{\partial v''} \quad J_3 = \frac{\partial i''}{\partial v'} \quad J_4 = \frac{\partial i''}{\partial v''} \quad (13)$$

can be easily obtained from the analytical expressions of the currents. For the non-linear reactor, these matrices have non-zero elements at the positions associated with the terminal nodes (rows and columns). The vector r is the residue associated with (11).

B. Thyristor-Controlled Reactor

Figure 3 presents the basic structure of the TCR considered in this paper. The basic steady state operation of this device is shown in Figure 4 [8].

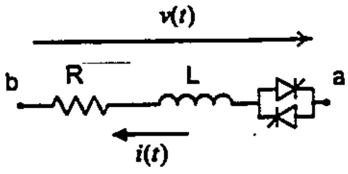

Fig. 3. Thyristor-controlled reactor

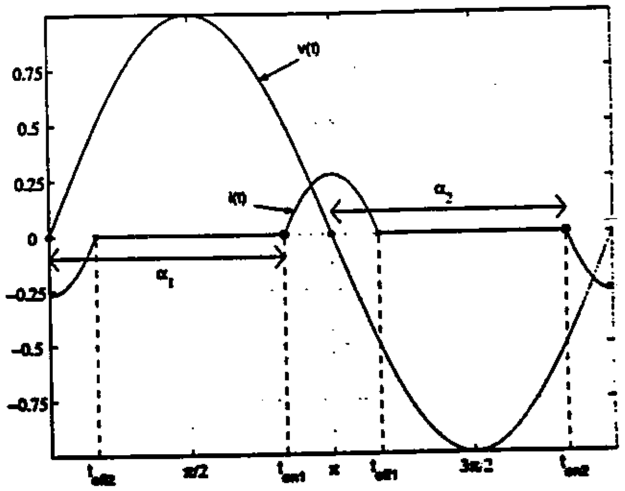

Fig. 4. TCR time domain characteristics

The TCR has non-continuous current conduction and the turn-on instants t_{on1} and t_{on2} are defined by the firing signal, obtained here as a delay α with respect to the zero crossing of the applied voltage:

$$t_{on} = t_0 + \alpha \quad (14)$$

The turn-off instants t_{off1} and t_{off2} are defined by the zero-crossing of the current. During conduction, the following differential equation applies:

$$L\frac{di}{dt} + Ri = v \quad (15)$$

The general solution of (15), considering the voltage expressed as in (6), is given by

$$i(t) = Ke^{-\frac{R}{L}(t-t_{on})} + \\ +2\sum_{h=1}^{n_h} \frac{v'_h}{Z_h}\cos(h\omega t - \varphi_h) - \frac{v''_h}{Z_h}\sin(h\omega t - \varphi_h) \quad (16)$$

where

$$Z_h = \sqrt{R^2 + (h\omega L)^2} \quad (17)$$

and

$$\varphi_h = \tan^{-1}\left(\frac{h\omega L}{R}\right) \quad (18)$$

The constant K can be determined from the initial condition ($i(t_{on}) = 0$):

$$K = -2\sum_{h=1}^{n_h} \frac{v'_h}{Z_h}\cos(h\omega t_{on} - \varphi_h) - \frac{v''_h}{Z_h}\sin(h\omega t_{on} - \varphi_h) \quad (19)$$

Equations (16)–(19) are valid at both conduction periods $t_{on1} \le t < t_{off1}$ and $t_{on2} \le t < t_{off2}$. It is easy to obtain the harmonic components of the TCR current from (16), both numerically, using an FFT algorithm, or analytically:

$$i_\ell = i'_\ell + j i''_\ell \qquad (20)$$

where

$$i'_\ell = \frac{\omega}{2\pi}\int_{t_{on1}}^{t_{off1}} i(t)\cos(\ell\omega t)\,dt + \frac{\omega}{2\pi}\int_{t_{on2}}^{t_{off2}} i(t)\cos(\ell\omega t)\,dt \qquad (21)$$

and

$$i''_\ell = \frac{\omega}{2\pi}\int_{t_{on1}}^{t_{off1}} i(t)\sin(\ell\omega t)\,dt + \frac{\omega}{2\pi}\int_{t_{on2}}^{t_{off2}} i(t)\sin(\ell\omega t)\,dt \qquad (22)$$

The harmonic components of the TCR current can be expressed as functions of the harmonic components of voltage and also of the turn-on and turn-off instants:

$$\begin{bmatrix} i'_{tcr} \\ i''_{tcr} \end{bmatrix} = \begin{bmatrix} f_1(v', v'', t_{on}, t_{off}) \\ f_2(v', v'', t_{on}, t_{off}) \end{bmatrix} \qquad (23)$$

where $t_{on} = [\ t_{on1}\ \ t_{on2}\ ]^T$ and $t_{off} = [\ t_{off1}\ \ t_{off2}\ ]^T$.

These currents are added to the vectors i' and i'' in (11) at the positions associated with the TCR terminal node. The linearization of (11) has now extra terms, as compared with (12):

$$\left\{\begin{bmatrix} G & -B \\ B & G \end{bmatrix} - \begin{bmatrix} J_a & J_b \\ J_c & J_d \end{bmatrix}\right\}\begin{bmatrix} \Delta v' \\ \Delta v'' \end{bmatrix} - \begin{bmatrix} J_e \\ J_f \end{bmatrix}\Delta t_{on} - \begin{bmatrix} J_m \\ J_n \end{bmatrix}\Delta t_{off} = -r \qquad (24)$$

The matrices J_a to J_n are obtained by taking the proper partial derivatives of (23). It can be proven that *incremental changes in t_{off} have no effect on the harmonic components of the current*. This is not valid, though, for changes in t_{on}. As shown in (14), the turn-on instant is determined by the zero crossing of a certain reference voltage at t_0 and by the delay α.

The delay α is usually the output of a closed-loop control system, controlling voltages and/or currents in the system. The zero crossing of the reference voltage changes due to variations in the voltage harmonic contents. The sensitivity of t_0 to these changes is shown in the Appendix.

IV. Results

A. Example System 1

Figure 5 depicts a very simple system useful to illustrate some of the issues associated with the different solution methods presented in Section II. The magnitude of the non-linear current was adjusted using different values for the constant K in (7).

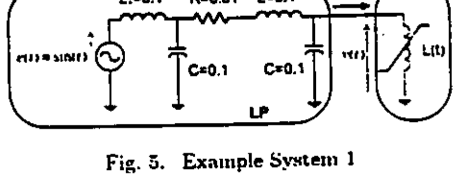

Fig. 5. Example System 1

Figure 6 shows the steady state voltages and currents. at the terminal of the non-linear reactor (NLR). for $K_1 = 0.1$. $K_2 = 0.5$ and $K_3 = 0.9$.

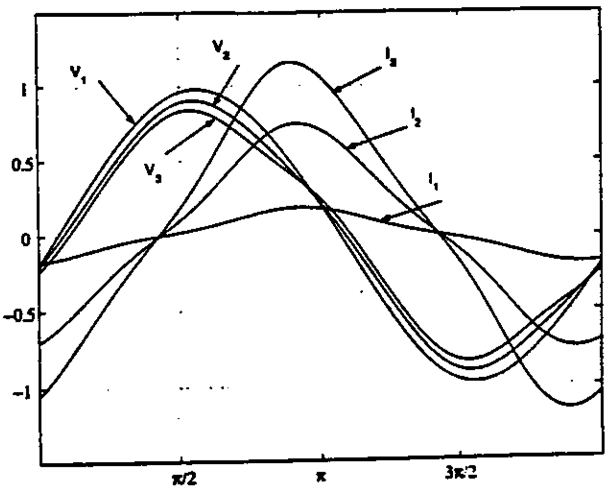

Fig. 6. Example System 1 – voltages and currents

The case with $K_3 = 0.9$ shows a peak voltage of 0.8 at the terminal of the NLR, corresponding roughly to a voltage drop of 20% across the LP of the system. This can be considered an extreme case and the full injection method (III.A.1, in the classification of Section II) and the non-linear Norton equivalent (III.A.2) required more than 30 iterations to satisfy the same convergence criterion that the Full Newton method (III.B.1) achieved in 5 iterations. The decoupled Newton method (III.B.2) required 16 iterations.

Figure 7 presents the convergence pattern of these four methods, for the intermediate case ($K_2 = 0.5$). The quadratic convergence of the full Newton method is evident in this figure. The full injection method and the non-linear Norton equivalent show almost identical behaviour. This denotes that the approximation used for the harmonic admittances in (2) is not good enough to result in better convergence. In this case, the linear part of the NLR was represented by a constant inductance $L = 1/K$. The decoupled Newton method shows quite good performance in this example, which can be viewed as an indication that the frequency coupling produced by the non-linear reactor is not very strong.

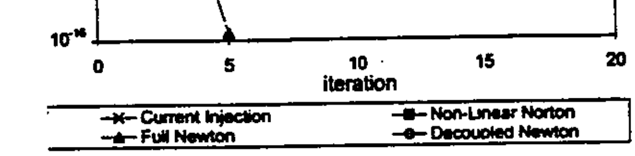

Fig. 7. Convergence pattern – Example System 1 – $K_2 = 0.5$

Figure 8 shows the behaviour of a system variable if the time domain initialization (Method 1. in Section II) is attempted. This graph gives the value of the mismatch of the non-linear current at the beginning of every cycle during a time domain simulation. It took more than 100 cycles for the transients to damp out and for the system to get close to its steady state ($\varepsilon < 10^{-3}$). Although the initial condition for this simulation was very far away from the steady state, the convergence to the steady state is linear and after 200 cycles the mismatch is still around 10^{-5}.

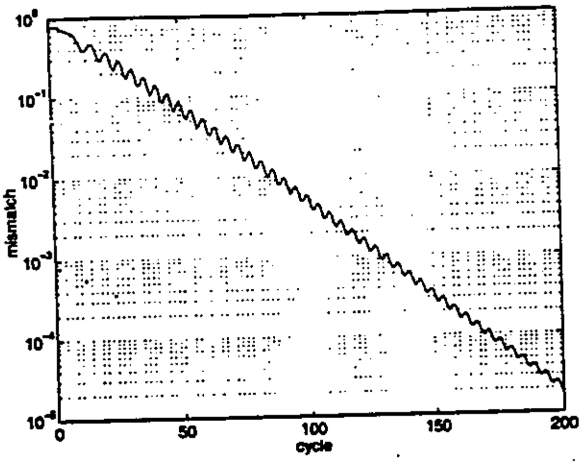

Fig. 8. Time domain initialization – $K_2 = 0.5$

B. Example System 2

Figure 9 presents the second example, in which the NLP now contains a TCR branch. A resistive load was also included in the LP.

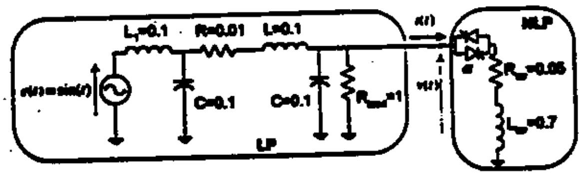

Fig. 9. Example System 2

Figure 10 shows the steady state voltages and currents, at the terminal of the TCR, for $\alpha_1 = 155°$, $\alpha_2 = 135°$, $\alpha_3 = 115°$ and $\alpha_4 = 95°$. These results are associated with fixed firing delays, which is equivalent to disregard the firing control system.

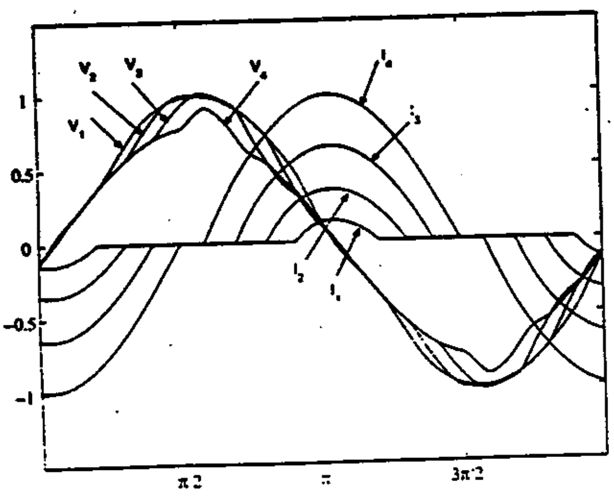

Fig. 10. Example System 2 – voltages and currents

Figure 11 presents the convergence pattern for the current injection (III.A.1), non-linear Norton equivalent (III.A.2), full Newton (III.B.1) and decoupled Newton methods, when applied to this example system with $\alpha_3 = 115°$. The Newton method converges quadratically, requiring few iterations. The decoupled Newton method shows a behaviour similar to that achieved with the full current injection method. This shows that the frequency coupling produced by the TCR model is quite strong and should be taken into consideration for faster convergence. In this example the use of the non-linear Norton Equivalent leads to much better performance than the current injection. For the TCR the harmonic admittances were obtained considering a constant equivalent inductance given by [2]

$$L_{eq} = \frac{\pi L}{\sigma - \sin\sigma} \tag{25}$$

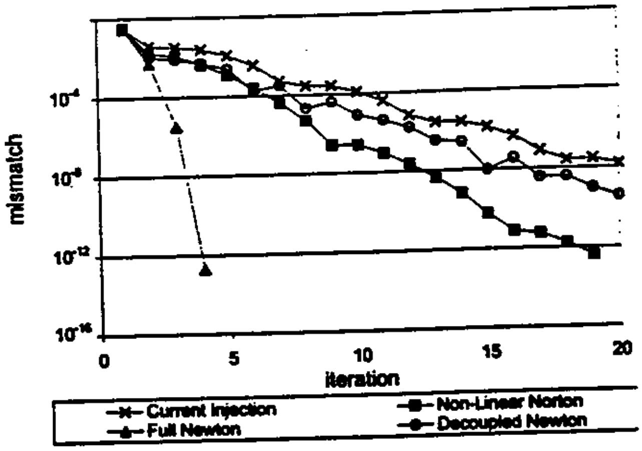

Fig. 11. Convergence pattern – Example System 2 – $\alpha = 115°$

V. Conclusions

The paper presented an analytical framework for the modeling of non-linear and switched devices that allowed the modular implementation of different methods to obtain the periodic steady state of the system. The approach was exemplified by the modeling of two devices: a non-linear reactor and a thyristor-controlled reactor.

Two simple example systems were used to illustrate the convergence and robustness characteristics of these methods. The analytical models allowed the implementation of a full Newton-Raphson method and fast (quadratic) and robust convergence was obtained in both studied cases. The results should stimulate the development of new analytical models and the extension of this approach to large systems. The sparsity of the Jacobian matrices associated with the non-linear devices can be exploited to yield very efficient implementations.

Acknowledgments

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APPENDIX – SENSITIVITY OF t_{on}

Given a voltage, expressed in its harmonic components as (6), its zero crossing occurs at $v(t_0) = 0$. The total derivative of the expression above. calculated at $t = t_0$, is given by

$$2\sum_{h=1}^{n_h} \cos(h\omega t_0)\,\Delta v'_h - 2\sum_{h=1}^{n_h} \sin(h\omega t_0)\,\Delta v''_h -$$
$$-2\sum_{h=1}^{n_h} h\omega\,[v'_h \sin(h\omega t_0) + v''_h \cos(h\omega t_0)]\Delta t_0 = 0$$

The expression of Δt_0 can be written, in vector form, as

$$\Delta t_0 = \frac{1}{S}\,[J'\Delta v' - J''\Delta v'']$$

where J' and J'' are row vectors. $\Delta v'$ and $\Delta v''$ are vectors containing all the harmonic components $\Delta v'_h$ and $\Delta v''_h$. The scalar S is the derivative of $v(t)$ calculated at t_0.

BIOGRAPHIES

Leonardo T. G. Lima received his B.Sc., M.Sc. and D.Sc. degrees from Federal University in Rio de Janeiro, Brazil, in 1986, 1991 and 1999 respectively. He is an associate professor at the Electrical Engineering Department of Universidade Federal Fluminense. in Niterói. Rio de Janeiro, Brazil and is currently working at the Department of Electrical and Computer Engineering of the University of Toronto, on a post-doctoral leave of absence. His interests are power systems dynamics and control. harmonics and electromagnetic transient analysis.

Adam Semlyen was born in 1923 in Rumania where he obtained a Dipl. Ing. degree and his Ph.D. He started his career there with an electric power utility and held academic positions at the Polytechnic Institute of Timisoara. In 1969 he joined the University of Toronto where he is a professor in the Department of Electrical and Computer Engineering, Emeritus since 1988. His research interests include steady state and dynamic analysis as well as computation of electromagnetic transients in power systems.

M. R. Iravani received his B. Sc. degree in electrical engineering in 1976 from Tehran Polytechnique University and started his career as a consulting engineer. He received his M.Sc. and Ph.D. degrees also in electrical engineering from the University of Manitoba. Canada in 1981 and 1985 respectively. Presently he is a professor at the University of Toronto. His research interests include power electronics and power system dynamic and control.