IGSPICE SIMULATION OF INDUCTION MACHINES
WITH SATURABLE INDUCTANCES

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Abstract. A new approach for dynamic analysis of induction machines with saturable inductances is presented. The machine dynamics are represented by a set of nonlinear time-varying differential equations. The machine parameters (i.e., saturable inductances) are modeled by closed form nonlinear functions. The equations which define the machine operation are modeled by $d$ and $q$ equivalent circuits. The equivalent circuits are simulated using IGSPICE software.

Key Words: IGSPICE, Induction Machines, Saturable Inductances.

Introduction

Power systems containing induction machines need to be studied in order to determine the interactions of power systems, drive systems and induction machines [1] through [11]. The problems normally studied are as follows: Will the motor start? If so, what is the magnitude of starting current? What is the duration of starting period? Will the motor start without exceeding the motor's thermal rating? What are the impacts of motor's high starting current on the power system load voltages? What is the magnitude of pulsating torque? How many times the motor can be started without degradation to windings' insulation due to high inrush currents? To study these problems, we need an accurate machine dynamic model for the entire operating range and then use the model to evaluate the transient response of starting current, torque, and motor speed.

The inductances which characterize the machine are normally considered constant in conventional models [3], [4], [5]. However, the inductances can vary widely depending on the state of the flux in different parts of the machine. The flux levels are in turn determined by the machine currents which depend on the operating modes [1]-[11]. This requires that the machine be represented by differential equations with time-varying and nonlinear parameters. The nonlinear time-varying parameters will characterize the saturated and unsaturated modes of the machine operation.

The unsaturated machine models have been routinely solved either on an analog or a digital computer [3], [4], [5]. Methods have been described in the literature for modeling saturation of air gap or magnetizing inductances [6]-[11]. In a recent paper, Lipo and Consoli [1], proposed a method for modeling the stator and rotor leakage reactance saturation. They experimentally measured the stator current and rotor speed of a machine during free acceleration from test and then simulated the machine model with their proposed saturated leakage reactances using an analog computer.

This paper presents a new analytical method for developing induction machine models where machine parameters (i.e., leakage and magnetizing inductances) may be saturated depending on the operating mode. The degree of inductance saturation is expressed by some nonlinear functions in terms of exciting currents. Furthermore, from saturated machine equations, equivalent circuit models are developed based on the approach proposed by Krause [3] and these circuits are simulated using IGSPICE software. The machine tested and simulated by Lipo and Consoli [1] is studied and the results are compared.

Induction Machine Equations Including Saturation Effects

The transient behavior of an induction machine is generally represented in an orthogonal coordinate system which is either rotating or stationary [3]. The machine performance can be expressed by the following equations, if we assume the angular speed $\omega$ of the rotating axes is not specified [3], [4]:

$$v_{qs} = r s i_{qs} + p \lambda_{qs} + \omega \lambda_{ds}$$  
$$v_{ds} = r s i_{ds} + p \lambda_{ds} - \omega \lambda_{qs}$$  
$$v_{qs} = r s i_{qs} + p \lambda_{qs}$$  
$$v_{qr} = r s i_{qr} + p \lambda_{qr} + (\omega - \omega_0) \lambda_{dr}$$  
$$v_{dr} = r s i_{dr} + p \lambda_{dr} - (\omega - \omega_0) \lambda_{qr}$$  
$$v_{qr} = r s i_{qr} + p \lambda_{qr}$$

(1) (2) (3) (4) (5) (6)

where $\lambda$ is the total flux-linkage of a particular winding, and $p$ is the operator $\frac{d}{dt}$.

The voltages $v_{qs}, v_{ds}, v_{qs}, v_{qr}, v_{dr}$, and $v_{qr}$ are the applied stator and rotor voltages. The equations of transformation relating the $d-q-o$ stator and rotor voltages to the actual applied $abc$ voltages are given in [3]. Since only 3-wire balanced voltage systems without a neutral return are considered, the zero quantities $i_{qs}, i_{ds}, i_{qr}$, and $v_{qs}$ are nonexistent. Therefore the Eq. (3) and Eq. (6) will not be included in this analysis. Furthermore, we will concentrate our analysis on single excited machines, consequently, the machine rotor windings are short and the rotor voltages, $v_{qr}$, and $v_{dr}$ are identically zero. Assuming that the machine is fed from a balanced voltage set and the reference frame is fixed in the stator [3], then the applied stator...
voltage can be expressed as:

\[ v_{se} = \left(\frac{2}{3}\right)^{1/2} V_{L-L} \cos \omega t \]  

(7)

\[ v_{ds} = -\left(\frac{2}{3}\right)^{1/2} V_{L-L} \sin \omega t \]  

(8)

where \( V_{L-L} \) is the rated line-to-line rms source voltage and \( \omega \) is the source electrical angular velocity.

The flux linkages \( \lambda_{se}, \lambda_{ds}, \lambda_{sp} \) and \( \lambda_{dp} \) of Eq. (1) through Eq. (4) are related to the currents by

\[ \lambda_{se} = (L_{se} + L_{te})i_{se} + L_m(i_{te} + i_{pr}) \]  

(9)

\[ \lambda_{ds} = (L_{ds} + L_{ds})i_{ds} + L_m(i_{dse} + i_{dr}) \]  

(10)

\[ \lambda_{sp} = (L_{sp} + L_{tp})i_{sp} + L_m(i_{te} + i_{pr}) \]  

(11)

\[ \lambda_{dp} = (L_{dp} + L_{tp})i_{dp} + L_m(i_{dse} + i_{dr}) \]  

(12)

In the above equations, we have used the modeling procedure proposed by Lipo [1]. In this approach, the total stator and rotor leakage inductances are separated into air-dependent and iron-dependent portions. The terms \( L_{se} \) and \( L_{ds} \) correspond to the sum of the iron-dependent saturated leakage inductances which represent the leakage flux of slot, yoke, back and skew leakage for stator and rotor respectively. These different components of leakage flux are discussed in [1] - [11]. The terms \( L_{se} \) and \( L_{ds} \) correspond to the air-dependent end-winding leakage inductances and they are assumed to be constant [1]. The term \( L_m \) is the magnetizing inductance and is also assumed to be saturable. When the saturation occurs in the machine, the inductances, \( L_{se}, L_{ds} \) and \( L_{sp} \), become nonlinear and the value of these inductances are determined from their exciting currents. Using Eq. (9) through Eq. (12) in Eq. (1) through Eq. (5), then we will have

\[ v_{se} = r_{se}i_{se} + L_{se}i_{se} + p\lambda_{se}' + \omega\lambda_{ds}' \]  

(13)

\[ v_{ds} = r_{ds}i_{ds} + L_{ds}i_{ds} + p\lambda_{ds}' - \omega\lambda_{se}' \]  

(14)

\[ v_{sp} = r_{sp}i_{sp} + L_{sp}i_{sp} + p\lambda_{sp}' + (\omega - \omega_r)\lambda_{se}' \]  

(15)

\[ v_{dp} = r_{dp}i_{dp} + L_{dp}i_{dp} + p\lambda_{dp}' - (\omega - \omega_r)\lambda_{ds}' \]  

(16)

where \( \lambda_{se}' \) represents the flux leakage nonlinear effects which can be expressed as

\[ \lambda_{se}' = L_{se}i_{se} + L_m'(i_{te} + i_{pr}) \]  

(17)

\[ \lambda_{ds}' = L_{ds}i_{ds} + L_m'(i_{dse} + i_{dr}) \]  

(18)

\[ \lambda_{sp}' = L_{sp}i_{sp} + L_m'(i_{te} + i_{pr}) \]  

(19)

\[ \lambda_{dp}' = L_{dp}i_{dp} + L_m'(i_{dse} + i_{dr}) \]  

(20)

In Eq. (17) through Eq. (20) \( L_{se}', L_{ds}', L_m' \) and \( L_{sp}', L_{dp}' \) are nonlinear inductances and they are functions of their respective exciting currents. To bring this nonlinear behavior into focus, we will define the following terms for \( q \)-axis

\[ \lambda_{se}^* = L_m'i_{se} \]  

(21)

\[ \lambda_{ds}^* = L_{ds}'i_{ds} \]  

(22)

\[ \lambda_{sp}^* = \lambda_{sp}' + \lambda_{se}^* \]  

(23)

where \( i_{ds}^* = i_{ds} + i_{dr} \).

Since \( L_{se} \) and \( L_{ds} \) are time-varying inductances and they are functions of their exciting currents \( i_{se} \) and \( i_{ds} \), we need to use the chain rule of derivative for calculating \( p\lambda_{se}^* \) term of equation (13). Applying the chain rule to equation (23), we have

\[ \frac{d}{dt}(\lambda_{se}^*) = \frac{d(i_{se})}{dt} + \frac{d(i_{ds})}{dt} \cdot \frac{d(i_{ds})}{dt} \]  

(24)

let

\[ L_{se} = \frac{d(i_{se})}{dt} \]  

(25)

\[ L_{ds} = \frac{d(i_{ds})}{dt} \]  

(26)

Then, Eq. (24) can be written as

\[ p\lambda_{se}^* = L_{sei} + p\lambda_{se} + L_{ds}i_{ds} \]  

(27)

The terms \( L_{sei} \) and \( L_{ds} \) are called the incremental inductances and they are the derivative of the hysteresis loop at a given exciting current. The above procedure can also be carried out for the terms \( \lambda_{ds}', \lambda_{sp}' \) and \( \lambda_{dp}' \). Using the above incremental inductance concept, the saturated machine inductances can be calculated and the machine dynamics in saturated and nonsaturated regions can be expressed by the following equations:

\[ v_{se} = r_{se}i_{se} + (L_{se}i_{se} + L_{sei}i_{se}) + L_m'i_{se} + v_1 \]  

(28)

\[ v_{ds} = r_{ds}i_{ds} + (L_{ds}i_{ds} + L_{ds}i_{ds}) + L_m'i_{ds} + v_2 \]  

(29)

\[ v_{sp} = r_{sp}i_{sp} + (L_{sp}i_{sp} + L_{sp}i_{sp}) + L_m'i_{sp} + v_3 \]  

(30)

\[ v_{dp} = r_{dp}i_{dp} + (L_{dp}i_{dp} + L_{dp}i_{dp}) + L_m'i_{dp} + v_4 \]  

(31)

and \( v_1, v_2, v_3 \), and \( v_4 \) can be written as

\[ v_1 = (\lambda_{se} + \lambda_{se} + \lambda_{se}) \cdot \omega \]  

(32)

\[ v_2 = (\lambda_{se} + \lambda_{se} + \lambda_{se}) \cdot \omega \]  

(33)

\[ v_3 = (\lambda_{se} + \lambda_{se} + \lambda_{se}) \cdot (\omega - \omega_r) \]  

(34)

\[ v_4 = (\lambda_{sp} + \lambda_{sp} + \lambda_{sp}) \cdot (\omega - \omega_r) \]  

(35)
Saturated Flux Models

In order to simulate the induction machine dynamics, the saturable flux leakage and magnetizing flux have to be modeled into some functional forms. The flux saturation can be reasonably modeled by a normal magnetizing curve rather than the hysteresis loop. The functional form used in this study for modeling the flux saturation effect is given by the following equation,

\[ \lambda = a_1 \arctan(\alpha_2 i) + a_3 i \]  
(30)

By using the actual data collected for \( \lambda \) and \( i \), the coefficients \( a_1, a_2, \) and \( a_3 \) can be estimated. The incremental inductance is then

\[ L_{inc} = \frac{d\lambda}{di} = \frac{a_1 \cdot a_2}{1 + a_2^2} + a_3 \]  
(37)

The techniques to identify the constants in Eq. (36) are based on an nonlinear least squares estimation algorithm developed by Marquardt [12]; this can be done by first writing the Taylor series of \( \lambda_m^*(\lambda_m) \) through linear terms, i.e.,

\[ \lambda_m^*(i_n, X+dX) = \lambda_m^*(i_n, X) + \sum_{j=1}^{3} \left( \frac{\partial \lambda_m^*(i_n, X)}{\partial x_j} \right) (dx)_j \quad n = 1, \ldots, N \]  
(38)

where

\[ X = [a_1, a_2, a_3]^T = [x_1, x_2, x_3]^T \]

The vector \( X \) in Eq. (38) represents the constants in Eq. (36) which need to be estimated, and \( i_n \) is the exciting current, \( dX \) is the small correction vector to \( X \). From Eq. (38), the following matrix equation can be obtained as

\[ (F + I\theta) dX = h \]  
(39)

where

\[ F(3 \times 3) = J^T \cdot J \]

\[ J(n \times 3) = \left( \frac{\partial \lambda_m^*(i_n, X)}{\partial x_j} \right), \quad n = 1, \ldots, N; \quad j = 1, 2, 3 \]

\[ h(3 \times 1) = \sum_{i=1}^{N} \left[ \lambda_m^*(i_n, X+dX) - \lambda_m^*(i_n, X) \right] \frac{\partial \lambda_m^*(i_n, X)}{\partial x_j} \]

In the above equations, \( N \) is the total number of experimental data points. Assuming

\[ \Delta \lambda_m^*(i) = \sum_{i=1}^{N} \left[ \lambda_m^*(i) - \hat{\lambda}_m^*(i) \right]^2 \]  
(40)

where \( \lambda_m^* \) is the measured flux leakage and \( \hat{\lambda}_m^* \) is the estimated one, the value of \( \theta \) in Eq. (39) is chosen based on the following strategy:

Let an arbitrary constant \( v \) to be greater than one, and let \( \theta^{(r-1)} \) represent the value of \( \theta \) in Eq. (39) at the iteration step \( r = 1 \). Initially, let \( \theta^{(0)} \) to be a constant, say 10^{-2}. Compute \( \Delta \lambda_m^*(\theta^{(r-1)}) \) and \( \Delta \lambda_m^*(\theta^{(r-1)}/v) \). Then the value of \( \theta^{(r)} \) at iteration step \( r \) can be selected by the following procedure:

1. If \( \Delta \lambda_m^*(\theta^{(r-1)}/v) \leq \Delta \lambda_m^*(\theta^{(r-1)}) \), let \( \theta^{(r)} = \theta^{(r-1)} \).
2. If \( \Delta \lambda_m^*(\theta^{(r-1)}/v) > \Delta \lambda_m^*(\theta^{(r-1)}) \) and \( \Delta \lambda_m^*(\theta^{(r-1)}) \leq \Delta \lambda_m^*(\theta^{(r)}) \), let \( \theta^{(r)} = \theta^{(r-1)}/v \).
3. If \( \lambda_m^*(\theta^{(r-1)}/v) > \lambda_m^*(\theta^{(r-1)}) \) and \( \lambda_m^*(\theta^{(r-1)}) > \lambda_m^*(\theta^{(r)}) \), increase \( \theta^{(r)} \) by successive multiplication of \( v \) until for some smallest \( v \), \( \Delta \lambda_m^*(\theta^{(r-1)}/v) \leq \Delta \lambda_m^*(\theta^{(r)}) \). Then let \( \theta^{(r)} = \theta^{(r-1)} \).

The above estimating process should be continued by updating the unknown vector \( X \) as

\[ X^{(r+1)} = X^{(r)} + dX^{(r)} \]

until the error between calculated points and corresponding test points can not be reduced further.

Induction Machine Circuit Models for Use in IGSPICE

The equations (28) - (35) can be represented by the following d-axis and q-axis equivalent circuits.

![Diagram](image_url)

Figure 1: a) The d-axis and b) the q-axis equivalent circuits of a squirrel-cage machine; arbitrary reference frame.

The IGSPICE is an excellent computer aided analysis package which facilitate fast dynamic simulation of circuit models [13]. To use the IGSPICE software, one needs to specify the type of a circuit element connected between two nodes. Many types of models exist for simulating active and passive circuit elements. However, the modeling of nonlinear elements are generally left for the user to develop. In the induction machine circuit models, the leakage and magnetizing inductances and the voltage sources \( v_1, v_2, v_3 \) and \( v_4 \) are current dependent voltage sources which are nonlinear functions of circuit inductances and their exciting currents. The circuit elements (see Fig. 1) connected between nodes \( (3, 4), (4, 5), (5, 0), (5, 0) \) and \( (6, 7) \) can not be modeled using standard IGSPICE modules.

To take advantage of the IGSPICE software, we used fortran subroutines to model the incremental inductances as defined by Eq. (37) and the voltage dependent current sources as defined by equations (3) through (36). Then, these subroutines were linked into the IGSPICE software for simulation of the electrical transient response of induction machines.
So far, we did not consider the electromagnetic torque equation and mechanical equation relating the total inertia of the machine to the electromagnetic torque. These equations are:

\[ T_e = \frac{3}{2}(p/2)(\lambda_m^* \dot{i}_t - \lambda_{im}^* \dot{i}_o) \]  
\[ (41) \]

\[ T_e - T_L = J\frac{d\omega_m}{dt} \]  
\[ (42) \]

\[ \omega_{rm} = \frac{2}{p} \omega_r \]  
\[ (43) \]

where \( p \) is the number of the poles, \( \omega_{rm} \) is the rotor mechanical speed, and \( J \) is the rotor inertia.

Equation (42) can be realized by the following circuit topology. The equivalent circuits of Figures 1 and 2 can be used to simulate the electrical and mechanical transient response of induction machines.

![Circuit topology](image)

**Figure 2: Circuit topology realizing the mechanical equation of the induction machine.**

**Simulation Results**

The machine which was simulated on an analog computer and its response was measured experimentally [1] is studied in this paper. The machine is a three-phase three-wire 230V squirrel-cage machine rated at 5 hp. Table 1 and 2 give the measured line voltages with respect to the exciting currents for the locked-rotor test and the no-load test respectively. The \( \lambda_{im}^{*(lo)} \) and \( \lambda_m^* \) are translated from \( V_s \) based on Eq. (44) and Eq. (45), respectively.

**Table 1: Locked-Rotor Test Data**

<table>
<thead>
<tr>
<th>Measured</th>
<th>Calculated</th>
<th>Measured</th>
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</thead>
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<td>0.00</td>
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<td>24.50</td>
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<td>32.50</td>
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</tr>
<tr>
<td>40.00</td>
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<td>20.00</td>
</tr>
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<td>47.50</td>
<td>0.05144</td>
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<td>54.50</td>
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<td>87.50</td>
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<tr>
<td>95.00</td>
<td>0.10289</td>
<td>70.00</td>
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</table>

**Table 2: No-Load Test Data**

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<tr>
<th>Measured</th>
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<th>Measured</th>
</tr>
</thead>
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<td>( \lambda_m^* )</td>
<td>( \lambda_m^* )</td>
<td>( \lambda_m^* )</td>
</tr>
<tr>
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</tr>
<tr>
<td>255.00</td>
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</tr>
</tbody>
</table>

\[ \lambda_{in(lo)}^* = \frac{1}{\sqrt{6}} \frac{V_s}{\omega_0} \]  
\[ (44) \]

\[ \lambda_m^* = \frac{\sqrt{2}}{3} \frac{V_s}{\omega_0} \]  
\[ (45) \]

In Eq. (44), \( \lambda_{im}^* \) and \( \lambda_m^* \) are assumed equal, and \( \omega_0 \) is taken to be 377 rad/sec for both Eq. (44) and Eq. (45).

From Table 1 and 2, the coefficients of Eq. (36) which is used to model the flux leakages \( \lambda_{im}^* \) and \( \lambda_m^* \) can be estimated using a computer program which employs the previously discussed algorithm [12], [14]. To do so, a set of initial estimated coefficients must be provided. This can be done by first calculating three different slopes along the saturation curves of \( \lambda_{im}^* \) and \( \lambda_m^* \). Then the initial estimates can be calculated using the three slopes, namely the increment inductances, and the corresponding exciting currents from Eq. (37).

With the initial estimates and the data provided by Table 1 and 2, the coefficients of Eq. (36) for modelling \( \lambda_{im}^* \) and \( \lambda_m^* \) are estimated. A sample computer output for the coefficients of \( \lambda_{im}^* \) is given by Table 3.

**Table 3: Estimated Coefficients of \( \lambda_{im}^* \)**

**NONLINEAR LEAST-SQUARES CURVE-FITTING PROGRAM**

<table>
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<th>CARD</th>
<th>DEP. VAR.</th>
<th>MIN: Y=0.0000400</th>
<th>MAX: Y=1.0258-01</th>
<th>RANGE: Y=1.0258-01</th>
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</thead>
<tbody>
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<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>A2</td>
<td>4.790258E-02</td>
<td></td>
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</tr>
<tr>
<td>3</td>
<td>A3</td>
<td>6.741713E-04</td>
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<tr>
<td>NO. OF OBSERVATIONS</td>
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<tr>
<td>NO. OF COEFFICIENTS</td>
<td>3</td>
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<td>RESIDUAL ROOT MEAN SQUARE</td>
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<td>RESIDUAL SUM OF SQUARES</td>
<td>0.00020317</td>
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</tbody>
</table>

Notice in Table 3, the dependent variable is \( \lambda_{im}^* \) and the independent variables are the coefficients of Eq. (36). The RMS error between the estimated \( \lambda_{im}^* \) and the \( \lambda_{im}^* \) listed in Table 1 is given by Table 3 as 0.00125. From Table 3, the flux leakage \( \lambda_{in(lo)}^* \) can be written as
\[
\lambda_{\text{int}} = \lambda_{\text{mr}} \\
= 0.02768 \arctan(0.0479i_{\text{m}}) + 6.74 \times 10^{-4} \quad (46)
\]

Similarly, \( \lambda_{\text{mr}} \) can be obtained as

\[
\lambda_{\text{mr}} = 0.4095 \arctan(0.1318i_{\text{m}}) 
\]

Then from Eq. (37) the incremental inductances are given by

\[
L_{\text{int}} = L_{\text{mr}} = \left( \frac{0.00133}{1 + 0.0023i_{\text{m}}^2} \right) + 6.74 \times 10^{-4} 
\]

\[
L_{\text{mr}} = \left( \frac{0.05306}{1 + 0.0174i_{\text{m}}^2} \right) 
\]

Note that the functional forms of \( L_{\text{int}}, L_{\text{mr}}, L_{\text{lda}}, \) and \( L_{\text{ldm}} \) are the same as Eq. (48), except the exciting current \( i_{\text{ex}} \) will be \( i_{\text{ex}}, i_{\text{ex}}, i_{\text{ex}}, i_{\text{ex}}, \) and \( i_{\text{ex}}, \) respectively. Similarly, \( L_{\text{mr}} \) and \( L_{\text{md}} \) terms are the same as Eq. (49) except the exciting current \( i_{\text{ex}} \) will be equal to \( i_{\text{ex}} \) and \( i_{\text{ex}} \) respectively. Figures (3), (4), (5), and (6) give the plots of leakage flux, incremental leakage inductance, magnetizing flux, and incremental magnetizing inductance respectively. Figures (7) and (8) show the stator current and the electromagnetic torque when the machine is accelerated at no-load and the saturated parameters and unsaturated parameters are used. The results indicate that the unsaturated model give a substantial error when compared with tests (see Ref. [1]). Therefore, we need to recognize that the model's parameters are time varying and they are a function of exciting current, and in order to obtain the correct simulation results, the saturation effect during the acceleration period can not be neglected.
Figure 7: IGSPICE simulation results for free acceleration of test machine using saturated magnetizing and leakage inductances.
Figure 8: IGSPICE simulation results for free acceleration of test machine using unsaturated (linear) magnetizing and leakage inductances.
Conclusion

The paper has presented a new approach for dynamic simulation of induction machines where the machine inductances are nonlinear functions of exciting currents. An induction machine which was tested in reference [1] was modeled and simulated using IGSPICE software. The IGSPICE is a very easy to use and very economical digital computer analysis package. This is in contrast with the analog computer approach which requires many hours of preparation. In addition, since IGSPICE has many circuits models of active and passive devices, the modeling of drive systems, and power system elements can also be simulated. We believe that IGSPICE will find widespread acceptance with those engineers engaged in the analysis, design and control of industrial drive systems.

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References


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