IGSPICE SIMULATION OF INDUCTION MACHINES WITH SATURABLE INDUCTANCES

A. Keyhani, Member, IEEE, and H. Tsai, Student Member, IEEE
The Ohio State University, Department of Electrical Engineering, Columbus, Ohio 43210

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 follow: 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 characterizes 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 depends 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.

88 SM 611-6 A paper recommended and approved by the IEEE Rotating Machinery Committee of the IEEE Power Engineering Society for presentation at the IEEE/PES 1988 Summer Meeting, Portland, Oregon, July 24 - 29, 1988. Manuscript submitted August 4, 1987; made available for printing April 28, 1988.

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 ω of the rotating axes is not specified [3], [4]:

$$v_{qs} = r_s i_{qs} + p\lambda_{qs} + \omega\lambda_{ds} \tag{1}$$

$$v_{ds} = r_s i_{ds} + p\lambda_{ds} - \omega\lambda_{qs} \tag{2}$$

$$v_{os} = r_s i_{os} + p\lambda_{os} \tag{3}$$

$$v_{qr} = r_r i_{qr} + p\lambda_{qr} + (\omega - \omega_r)\lambda_{dr} \tag{4}$$

$$v_{dr} = r_r i_{dr} + p\lambda_{dr} - (\omega - \omega_r)\lambda_{qr} \tag{5}$$

$$v_{or} = r_r i_{or} + p\lambda_{or} \tag{6}$$

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

The voltages $v_{qs}, v_{ds}, v_{os}, v_{qr}, v_{dr}$, and v_{or} 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_{or}, i_{os}, v_{or}, and v_{os} 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

0885-8969/89/0300-0118$01.00

voltages can be expressed as:

$$v_{qs} = \left(\frac{2}{3}\right)^{1/2} V_{L-L} \cos \omega_e t \tag{7}$$

$$v_{ds} = -\left(\frac{2}{3}\right)^{1/2} V_{L-L} \sin \omega_e t \tag{8}$$

where V_{L-L} is the rated line-to-line rms source voltage and ω_e is the source electrical angular velocity.

The flux linkages $\lambda_{qs}, \lambda_{ds}, \lambda_{qr}$ and λ_{dr} of Eq. (1) through Eq. (4) are related to the currents by

$$\lambda_{qs} = (L_{lsa} + L_{lsi})i_{qs} + L_m(i_{qs} + i_{qr}) \tag{9}$$

$$\lambda_{ds} = (L_{lsa} + L_{lsi})i_{ds} + L_m(i_{ds} + i_{dr}) \tag{10}$$

$$\lambda_{qr} = (L_{lra} + L_{lri})i_{qr} + L_m(i_{qs} + i_{qr}) \tag{11}$$

$$\lambda_{dr} = (L_{lra} + L_{lri})i_{dr} + L_m(i_{ds} + i_{dr}) \tag{12}$$

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_{lsi} and L_{lri} correspond to the sum of the iron-depentdent saturated leakage inductances which represent the leakage flux of slot, Zig-Zag, belt and skew leakage for stator and rotor respectively. These different components of leakage flux are discussed in [1] - [11]. The terms L_{lsa} and L_{lra} 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_m, L_{lsi} and L_{lri} 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

$$\begin{aligned} v_{qs} &= r_s i_{qs} + L_{lsa} p i_{qs} + p\lambda^*_{qs} + \omega\lambda^*_{ds} \\ &\quad +\omega L_{lsa} i_{ds} \end{aligned} \tag{13}$$

$$\begin{aligned} v_{ds} &= r_s i_{ds} + L_{lsa} p i_{ds} + p\lambda^*_{ds} - \omega\lambda^*_{qs} \\ &= -\omega L_{lsa} i_{qs} \end{aligned} \tag{14}$$

$$\begin{aligned} v_{qr} &= r_r i_{qr} + L_{lra} p i_{qr} + p\lambda^*_{qr} + (\omega - \omega_r)\lambda^*_{dr} \\ &\quad -(\omega - \omega_r) L_{lra} i_{dr} \end{aligned} \tag{15}$$

$$\begin{aligned} v_{dr} &= r_r i_{dr} + L_{lra} p i_{dr} + p\lambda^*_{dr} - (\omega - \omega_r)\lambda^*_{qr} \\ &- \ (\omega - \omega_r) L_{lra} \end{aligned} \tag{16}$$

where " * " represents the flux leakage nonlinear effects which can be expressed as

$$\lambda^*_{qs} = L^*_{lsi} i_{qs} + L^*_m(i_{qs} + i_{qr}) \tag{17}$$

$$\lambda^*_{ds} = L^*_{lsi} i_{ds} + L^*_m(i_{ds} + i_{dr}) \tag{18}$$

$$\lambda^*_{qr} = L^*_{lri} i_{qr} + L^*_m(i_{qs} + i_{qr}) \tag{19}$$

$$\lambda^*_{dr} = L^*_{lri} i_{dr} + L^*_m(i_{ds} + i_{dr}) \tag{20}$$

In Eq. (17) through Eq. (20) L^*_{lsi}, L^*_{lri} and L^*_m 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

$$\lambda^*_{mq} = L^*_{mq} i_{mq} \tag{21}$$

$$\lambda^*_{lqsi} = L^*_{lqsi} i_{qs} \tag{22}$$

$$\lambda^*_{qs} = \lambda^*_{lqsi} + \lambda^*_{mq} \tag{23}$$

where $i_{mq} = i_{qs} + i_{qr}$.

Since L^*_{lqsi} and L^*_{mq} are time-varying inductances and they are functions of their exciting currents i_{qs} and i_{mq}, we need to use the chain rule of derivative for calculating $p\lambda^*_{qs}$ term of equation (13). Applying the chain rule to equation (23), we have

$$\frac{d}{dt}(\lambda^*_{qs}) = \frac{d(\lambda^*_{lqsi})}{d(i_{qs})} \cdot \frac{d(i_{qs})}{dt} + \frac{d(\lambda^*_{mq})}{(di_{mq})} \cdot \frac{d(i_{mq})}{dt} \tag{24}$$

let

$$\bar{L}_{lqsi} = \frac{d(\lambda^*_{lqsi})}{d(i_{qs})} \tag{25}$$

$$\bar{L}_{mq} = \frac{d(\lambda^*_{mq})}{d(i_{mq})} \tag{26}$$

Then, Eq. (24) can be written as

$$p\lambda^*_{qs} = \bar{L}_{lqsi} \cdot p i_{qs} + \bar{L}_{mq} p i_{mq} \tag{27}$$

The terms $\bar{L}_{lqsi}$ and $\bar{L}_{mq}$ 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 $\lambda^*_{ds}, \lambda^*_{qr}$ and λ^*_{dr}. 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:

$$v_{qs} = r_s i_{qs} + (L_{lqsa} + \bar{L}_{lqsi}) p i_{qs} + \bar{L}_{mq} p i_{mq} + v_1 \tag{28}$$

$$v_{ds} = r_s i_{ds} + (L_{ldsa} + \bar{L}_{ldsi}) p i_{ds} + \bar{L}_{md} p i_{md} - v_2 \tag{29}$$

$$v_{qr} = r_r i_{qr} + (L_{lqra} + \bar{L}_{lqri}) p i_{qr} + \bar{L}_{mq} p i_{mq} + v_3 \tag{30}$$

$$v_{dr} = r_r i_{dr} + (L_{ldra} + \bar{L}_{ldri}) p i_{dr} + \bar{L}_{md} p i_{md} - v_4 \tag{31}$$

and v_1, v_2, v_3, and v_4 can be written as

$$v_1 = (\lambda_{ldsa} + \lambda^*_{ldsi} + \lambda^*_{md}) \cdot \omega \tag{32}$$

$$v_2 = (\lambda_{lqsa} + \lambda^*_{ldsi} + \lambda^*_{mq}) \cdot \omega \tag{33}$$

$$v_3 = (\lambda_{ldra} + \lambda^*_{ldri} + \lambda^*_{md}) \cdot (\omega - \omega_r) \tag{34}$$

$$v_4 = (\lambda_{lqra} + \lambda^*_{lqri} + \lambda^*_{mq}) \cdot (\omega - \omega_r) \tag{35}$$

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,

$$\lambda = a_1 \arctan(a_2 i) + a_3 i \tag{36}$$

By using the actual data collected for λ and i, the coefficients a_1, a_2, and a_3 can be estimated. the incremental inductance is then

$$L_{inc} = \frac{d\lambda}{di} = \frac{a_1 \cdot a_2}{1 + a_2^2 i^2} + a_3 \tag{37}$$

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 $\lambda_{lsi}^*(\lambda_m^*)$ through linear terms, i.e.,

$$\lambda_{lsi}^*(i_n, X+dX) = \lambda_{lsi}^*(i_n, X) + \sum_{j=1}^{3} \left(\frac{\partial \lambda_{lsi}^*}{\partial x_j}\right)(dx)_j \qquad n = 1, \ldots, N \tag{38}$$

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

$$(F + I\theta)dX = h \tag{39}$$

where

$$F(3 \times 3) = J^T \cdot J$$

$$J(n \times 3) = \left(\frac{\partial \lambda_{lsi}^*(i_n)}{\partial x_j}\right), \qquad n = 1, \ldots, N; \qquad j = 1, 2, 3$$

$$h(3 \times 1) = \sum_{i=1}^{N} [\lambda_{lsi}^*(i_i, X + dX) - \lambda_{lsi}^*(i_i, X)] \frac{\partial \lambda_{lsi}^*(i_i, X)}{\partial x_j}$$

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

$$\Delta \lambda_{lsi}^* = \sum_{i=1}^{N} \left[\lambda_{lsi}^*(i_i) - \hat{\lambda}_{lsi}^*(i_i)\right]^2 \tag{40}$$

where λ_{lsi}^* is the measured flux leakage and $\hat{\lambda}_{lsi}^*$ is the estimated one, the value of θ 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 θ in Eq. (39) at the iteration step $r-1$. Initially, let $\theta^{(0)}$ to be a constant, say 10^{-2}. Compute $\Delta\lambda_{lsi}^*(\theta^{(r-1)})$ and $\Delta\lambda_{lsi}^*(\theta^{(r-1)}/v)$. Then the value of $\theta^{(r)}$ at iteration step r can be selected by the following procedure:

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

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.

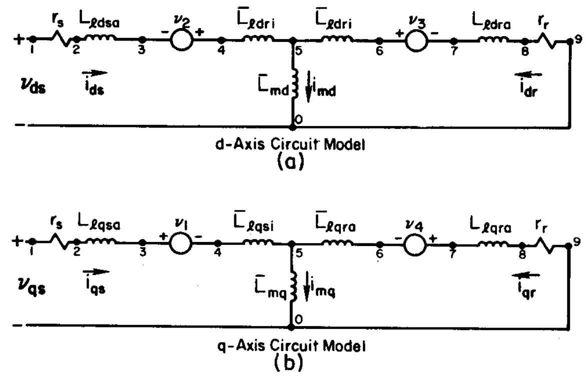


Figure 1: a) The d-axis and b) the q-axis equivalent circuits of a squirrel-cage machine; arbitrary reference frame.

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 $(3,4), (4,5), (5,0), (5,6)$ 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:

$$T_e = (3/2)(p/2)(\lambda^*_{ds} i_{qs} - \lambda^*_{qs} i_{ds}) \tag{41}$$

$$T_e - T_L = J(\frac{d\omega_{rm}}{dt}) \tag{42}$$

$$\omega_{rm} = \frac{2}{p}\omega_r \tag{43}$$

where p is the number of the poles, ω_{rm} is the rotor mechanical speed, and J is the rotor inertia.

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

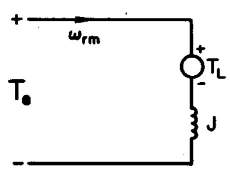


Figure 2: Circuit topology realizing the mechanical equation of the induction machine.

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 $\lambda^*_{lsi(lri)}$ and λ^*_m are translated from V_s based on Eq. (44) and Eq. (45), respectively.

Table 1: Locked-Rotor Test Data

Measured Vs (V, RMS)	Calculated $\lambda^*_{lsi} = \lambda^*_{lri}$	Measured Is (A, RMS)
0.00	0.00000	0.00
6.25	0.00677	1.88
15.00	0.01625	5.00
24.50	0.02653	10.00
32.50	0.03520	15.00
40.00	0.04332	20.00
47.50	0.05144	25.00
54.50	0.05902	30.00
60.75	0.06579	35.00
67.50	0.07310	40.00
72.50	0.07852	45.00
77.50	0.08393	50.00
82.50	0.08935	55.00
87.50	0.09476	60.00
91.25	0.09882	65.00
95.00	0.10289	70.00

Table 2: No-Load Test Data

Measured Vs (V, RMS)	Calculated λ^*_m	Measured Is (A, RMS)
0.00	0.00000	0.00
70.00	0.15162	2.50
136.25	0.29512	5.00
175.00	0.37905	6.75
187.50	0.40613	7.50
200.00	0.43320	8.75
215.00	0.46569	11.50
230.00	0.49818	15.00
245.00	0.53067	20.00
255.00	0.55233	25.00

$$\lambda^*_{lsi(lri)} = \frac{1}{\sqrt{6}}\frac{V_s}{\omega_b} \tag{44}$$

$$\lambda^*_m = \sqrt{\frac{2}{3}}\frac{V_s}{\omega_b} \tag{45}$$

In Eq. (44), λ^*_{lsi} and λ^*_{lri} are assumed equal, and ω_b 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 λ^*_{lsi} and λ^*_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 λ^*_{lsi} and λ^*_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 λ^*_m and λ^*_{lsi} are estimated. A sample computer output for the coefficients of λ^*_{lsi} is given by Table 3.

Table 3: Estimated Coefficients of λ^*_{lsi}.

```
NONLINEAR LEAST-SQUARES CURVE-FITTING PROGRAM

CARD    DEP.VAR.:    MIN Y=0.000E+00
                     MAX Y=1.029E-01
                   RANGE Y=1.029E-01

IND.VAR(I)    NAME     COEF.A(I)
    1          A1      2.76848E-02
    2          A2      4.79025E-02
    3          A3      6.74171E-04

NO. OF OBSERVATIONS                16
NO. OF COEFFICIENTS                 3
RESIDUAL ROOT MEAN SQUARE          0.00124994
RESIDUAL MEAN SQUARE               0.00000156
RESIDUAL SUM OF SQUARES            0.00002031
```

Notice in Table 3, the dependent variable is λ^*_{lsi} and the independent variables are the coefficients of Eq. (36). The RMS error between the estimated λ^*_{lsi} and the λ^*_{lsi} listed in Table 1 is given by Table 3 as 0.00125. From Table 3, the flux leakage$\lambda^*_{lsi(lri)}$ can be written as

$$\lambda^*_{lsi} = \lambda^*_{lri} = 0.02768\arctan(0.0479 i_{ex}) + 6.74 \times 10^{-4} i_{ex} \quad (46)$$

Similarly, λ^*_m can be obtained as

$$\lambda^*_m = 0.4095\arctan(0.1318 i_{ex}) \quad (47)$$

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

$$\bar{L}_{lsi} = \bar{L}_{lri} = \left(\frac{.00133}{1 + .0023 i^2_{ex}}\right) + 6.74 \times 10^{-4} \quad (48)$$

$$\bar{L}_m = \left(\frac{0.05396}{1 + 0.0174 i^2_{ex}}\right) \quad (49)$$

Note that the functional forms of $\bar{L}_{lqsi}, \bar{L}_{lqri}, \bar{L}_{ldsi}$ and $\bar{L}_{ldri}$ are the same as Eq. (48), except the exciting current i_{ex} will be i_{qs}, i_{qr}, i_{ds} and i_{dr} respectively. Similarly, $\bar{L}_{mq}$ and $\bar{L}_{md}$ terms are the same as Eq. (49) except the exciting current i_{ex} will be equal to i_{mq} and i_{md} 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.**

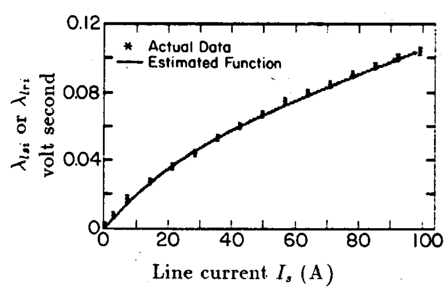


Figure 3: Locked-rotor saturation curve for tested 5-hp machine of reference [1] (I_s is in peak to peak value)

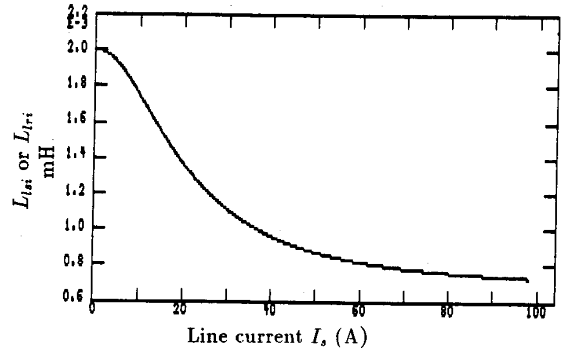


Figure 4: Incremental Inductance $\bar{L}_{lsi}$ or $\bar{L}_{lri}$ (I_s is in peak to peak value)

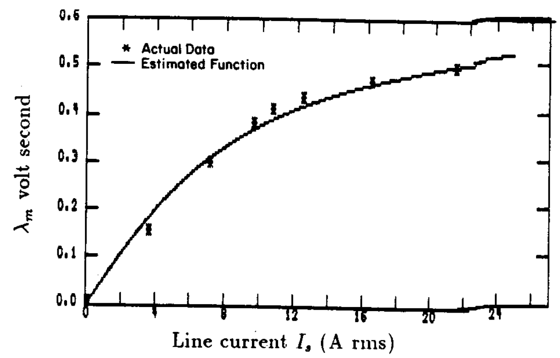


Figure 5: No-load saturation curve for tested 5-hp machine of reference

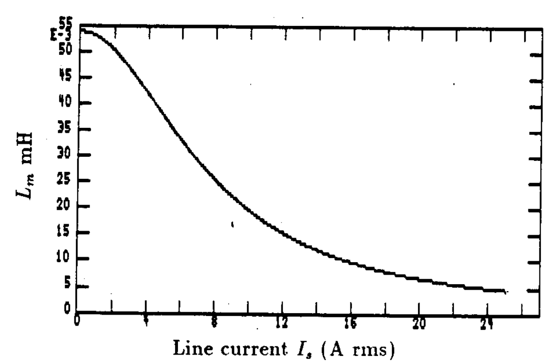
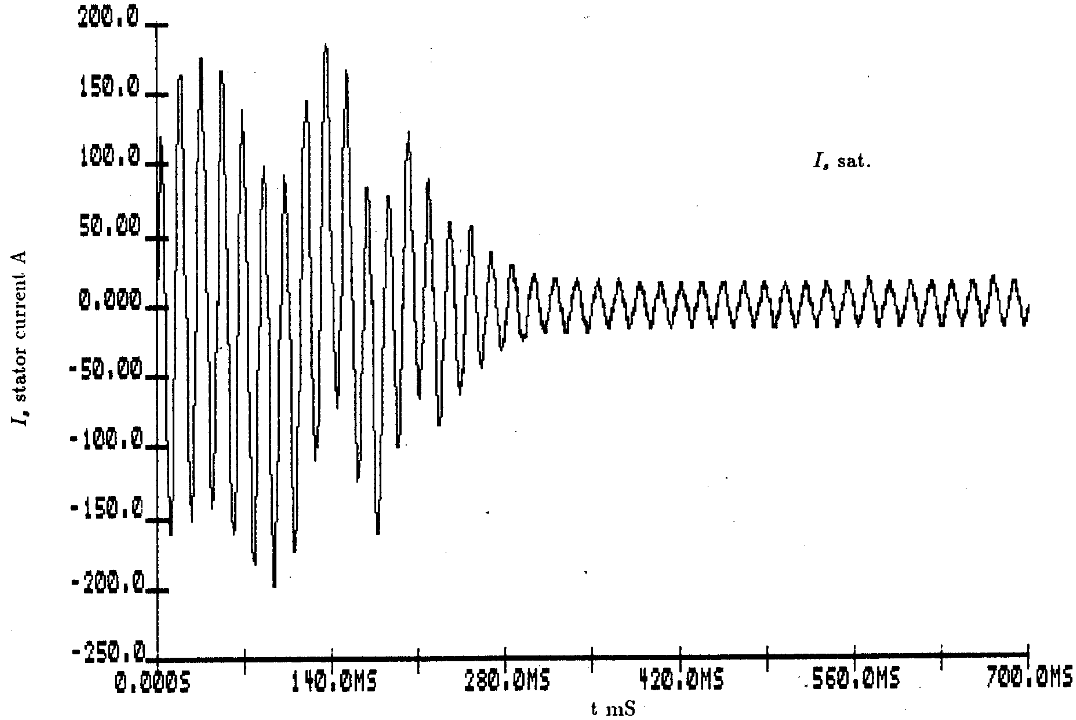


Figure 6: Incremental Inductance $\bar{L}_m$



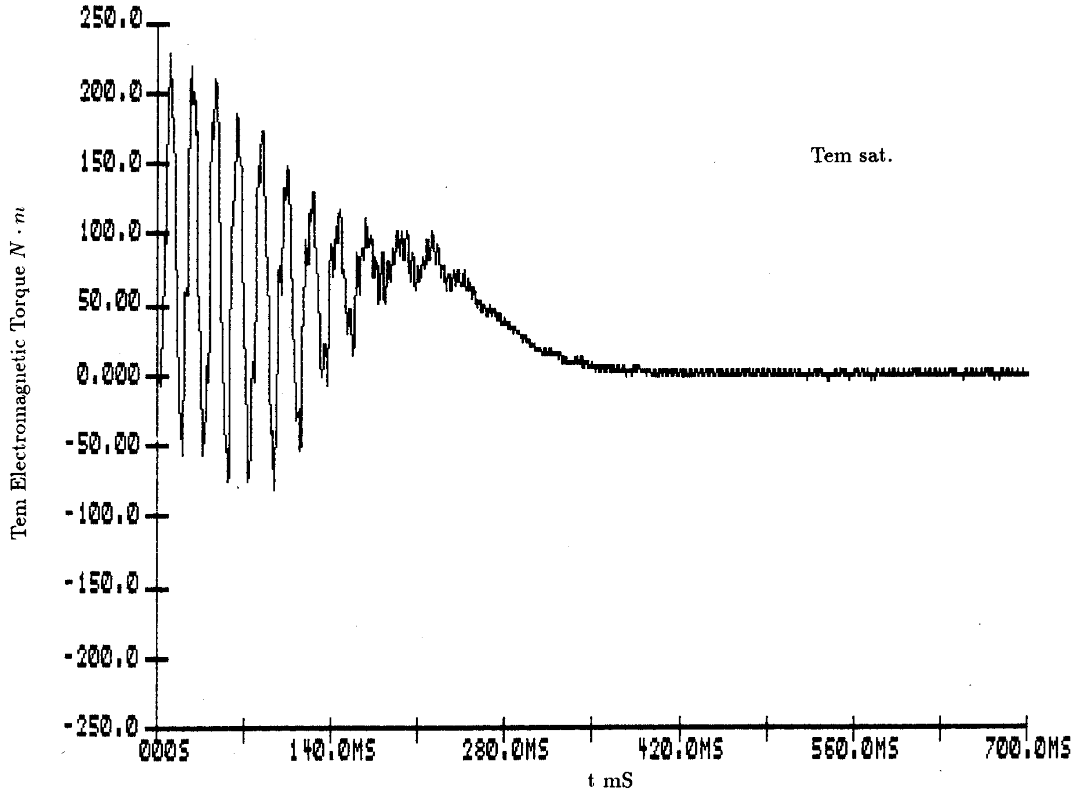
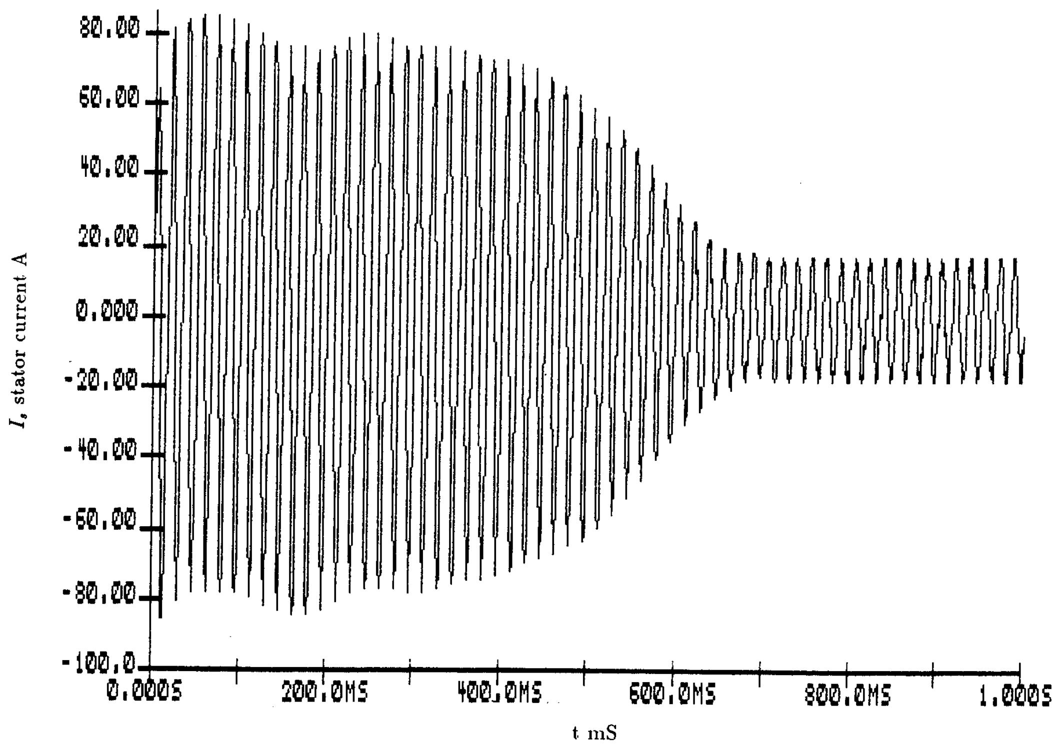


Figure 7: IGSPICE simulation results for free acceleration of test machine using saturated magnetizing and leakage inductances



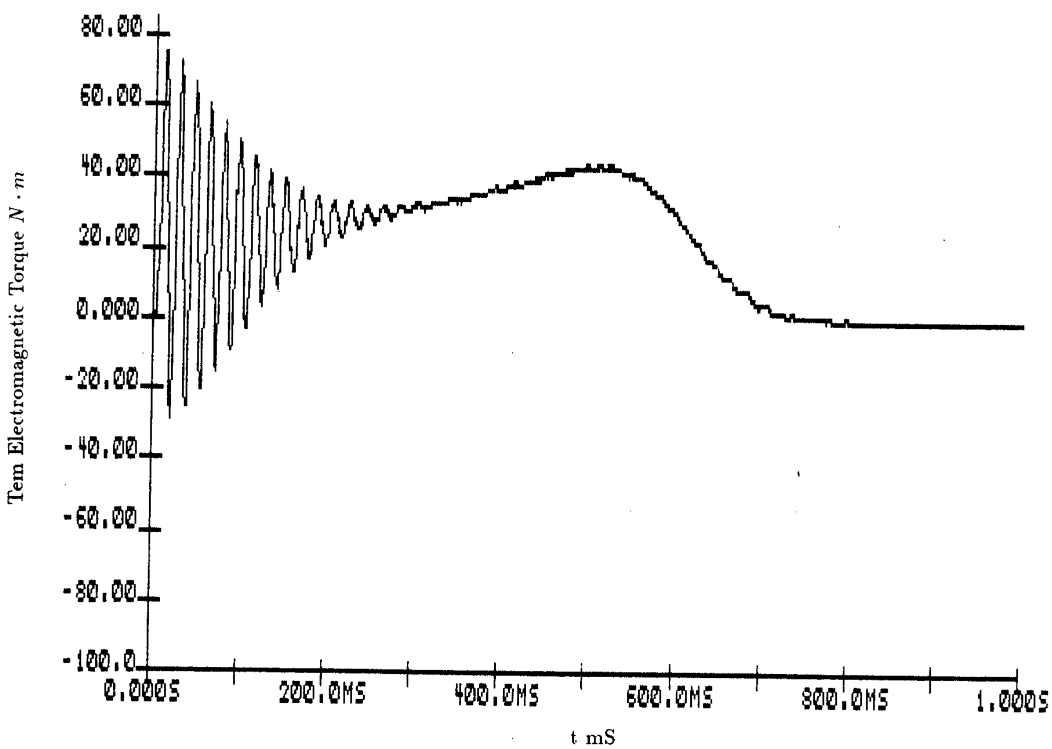


Figure 8: IGSPICE simulation results for free acceleration of test machine using unsaturated (linear) magnetizing and leakage inductances

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 modules 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.

Acknowledgments

The authors wish to express their gratitude and appreciation to the chairman of the Department, Professor H.C. Ko, for his continuous support and assistance. Financial support by the NSF Grant ECS-8303330 and the General Electric Research Foundation are gratefully acknowledged.

References

[1] T.A Lipo and A. Consoli, "Modeling and Simulation of Induction Motors with Saturable Leakage Reactances" *IEEE Trans. Ind. Appl.*, vol. IA-20, No.1, pp.180–189, Jan./Feb. 1984.

[2] H. C. Stanley, "Analysis of symmetrical induction machinery," *AIEE Trans.*, vol, 57, pp. 751–757, 1938.

[3] P. C. Krause and C. H. Thomas, "Simulation of symmetrical induction machinery," *IEEE Trans. Power App. Syst.*, vol. PAS-84, pp. 1038–1053, Nov. 1965.

[4] L. F. Wiederholt, A. F. Fath, and H. J. Wertz, "Motor transient analysis on a small digital computer," *IEEE Trans. Power App. Syst.*, vol. PAS-86, pp.819–824, July 1967.

[5] E. Klingshirn and H. E. Jordan, "Simulation of polyphase induction machines with deep rotor bars," *IEEE Trans. Power App. Syst.*, vol. PAS-89, pp. 1038–1043, July/Aug. 1970.

[6] F. P. DeMello and G. W. Walsh, "Reclosing transients in induction motors with terminal capacitors," *Trans. AIEE (Power App. Syst.)*, vol. 80, pp. 1206–1213, Feb. 1961.

[7] P. D. Agarwal and P. L. Alger, "Saturation factors for leakagereactance of induction machines," *AIEE Trans., Pt. III (Power App. Syst.)*, vol. 79, pp.1037–1042, 1960(Feb. 1961 sect.).

[8] G. Angst, "Saturation factors for leakage reactance of induction motors with skewed rotors," *IEEE Trans. Power App. Syst.*, vol.PAS-82, pp. 716–725, Oct. 1963.

[9] G. Grellet and L. Mariaux, "Saturation factors for leakage reactances of cage induction motors," in *IEEE Ind. Appl. Soc. Ann. Meeting Conf. Rec.*, pp. 1066–1070, 1976.

[10] *IEEE Standard Test Procedure for Polyphase Induction Motors and Generators,* IEEE standard 112-1978, May 12, 1978.

[11] V. B. Honsinger, "Measurement of end-winding leakage reactance," *AIEE Trans. Part IIIA. Power App. Syst.*, vol. 78, pp.426–431, Aug. 1959.

[12] Donald W. Marquardt,"An Algorithm for Least-Square Estimation of Nonlinear Parameters," *J. Soc. Ind. Appl Math*, vol. 11, no. 2, pp. 431-441, June 1963.

[13] *IGSPICE Manual*, A.B. Associates, Inc., P.O. Box 82215, Tampa, FL.

[14] C. Daniel and F.S. Wood, *Fitting Equations to Data*, John Wiley & Sons, New York, 1980.

Ali Keyhani (M '67) was born in Tehran, Iran on December 27, 1942. He received the Ph.D. degree from Purdue University, West Lafayette, Indiana in 1975. From 1967 to 1969, he worked for Hewlett-Packard Co. on the computer-aided design of electronic transformers. From 1970 to 1973, he worked for Columbus and Southern Ohio Electric Co. on computer applications for power system engineering problems. In 1975, he joined TRW controls and worked on the development of computer programs for energy control centers. Since 1980, he has been an Associate Professor of Electrical Engineering at The Ohio State University. Dr. Keyhani's research interest is in control of power systems, environmental systems, motion control and modeling, parameter estimation, failure detection of electrical machines, transformers and drive systems.

Hong Tsai (S '87) was born in Wuhan, PRC, on July 16, 1962. He received the B.S. degree from The Ohio State University, Columbus, Ohio, in 1986. Currently he is working toward his M.S. degree in electrical engineering at The Ohio State University.