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PARAMETER IDENTIFICATION OF AN INVERTER–FED INDUCTION MOTOR AT STANDSTILL WITH A CORRELATION METHOD

A. Bünte, H. Grotstollen

Universität Paderborn – Gesamthochschule –, Germany

Abstract. Two methods to identify the electrical parameters and the magnetization curve of a voltage source inverter–fed induction motor (IM) at standstill will be presented.
A special advantage of these methods is that the differences between the reference voltage values and the real phase voltage values caused by the unknown switching times of the PWM–inverter will be completely suppressed. Therefore a measurement of the phase voltage is not necessary and the parameters of the induction motor can be estimated with a high accuracy.
Thus, no additional hardware is necessary and therefore this identification method is a good solution for use in commercial self–commissioning drives.

Keywords. Induction Motor, Frequency Response, Inverter, Self–Commissioning, Magnetic hysteresis

1 INTRODUCTION

The increasing user's demands on electrical drives with inverter–fed induction machines led to complex control schemes like Field Orientated Control or Direct Self Control. This control schemes satisfy the demands on the dynamics but if the full performance of these schemes is required the electrical parameters of the induction machine (IM) have to be known. For an optimal design of the drive the knowledge of the saturation curve is necessary, too.
To apply the control scheme to an unknown IM the electrical parameters must be determined in a first step. The problem of the parameters determination is that they can't be measured with a simple measuring, with one exception: The stator resistance can be measured with a simple d.c. test. Furthermore the accuracy of parameters calculated from name–plate–data is not sufficient for the most applications [4].
These are the reasons why it is a problem if a customer wants to combine an inverter of one manufacturer with an IM of another manufacturer.
The commissioning by specially trained staff is a poor solution in view of the costs. An alternative is to fit out the inverter with the capability of self–commissioning. In this case the essential tasks are: "parameter identification", "design of the controller" and "supervision and tests of plausibility".
In this paper the parameter identification will be discussed.

In case of series production of the inverter the self–tuning should require only a minimum of additional hardware, to minimize the costs. The realisation of a drive with the capability of self–tuning is possible with on–line and off–line methods.
In the first way the parameters will be estimated and the regulation characteristics will be optimized steadily at run time.
In the second way the parameters will be identified in a first step. Then the controller will be designed and at last the modified software for the signal processor or the controller will be started. In this case only an adaptation to the inertia factor and perhaps to temperature variation is necessary at run time.
Therefore the stressing of the controller with the off–line method is less than the stressing with the on–line method. So, in view of the costs the off–line method should be preferred because a cheaper controller with a lower performance can be used.
Basic demands that should be met by a modern off–line parameter identification scheme are:

- The stimulation of the IM should be carried out with the inverter of the drive.
- The electrical parameters should be identified at standstill of the IM because a rotation at the commissioning phase is often undesirable.
- The standstill of the IM results to the next demand. The IM must not produce a torque because a locking of the rotor is undesirable as well.

Considering the costs, these demands are problematic. Apart from the calculation from name–plate–data all identification schemes need measured values of phase currents and phase voltage. The voltage at the most tests must be very small because at standstill only the small stator resistance restricts the current at steady state.
Usually a controlled drive is fit out with a current detection but not with a phase voltage detection because the measuring is difficult and costly in view of the inverter. At best only a detection of the d.c. link voltage can be presupposed for low cost drives.
That means, that the phase voltage must be estimated with the known value of the voltage source U_D and the reference values of the inverter transistors duty cycles $d^*_{a,b,c}$. If the real, phase current dependent switching behaviour of the transistors is taken into account the real phase voltage is not accurately known. Exactly in case of small phase voltage, the deviation is enormous and mustn't be neglected. So, the identification is possibly disturbed.
This is the reason why nearly all identification schemes need a sometimes very costly phase voltage detection [5],[9].
A possibility to avoid this detection is to correct the reference values of the duty cycles in dependency of the phase currents with an error curve [4], [8]. This curve must be determined for every new inverter because of the switching times variation of the transistors.

In this paper an off–line method will be presented, which facilitates the identification of the electrical parameters with a minimum of sensory equipment, especially a phase voltage detection isn't necessary. A measuring of the magnetization curve will be presented as well.

2 THE MODEL OF THE INDUCTION MACHINE AT STANDSTILL

A model of the induction machine which considers the saturation was deduced in [2].
This model describes the IM in the stator reference frame. Skin effects, magnetic hysteresis and iron losses are not taken into account.
(1) shows this model:

$$\begin{bmatrix} u_{S\alpha} \\ u_{S\beta} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_S & 0 & 0 & 0 \\ 0 & R_S & 0 & 0 \\ 0 & \omega_{RS} L_h & R_R & \omega_{RS} L_R \\ -\omega_{RS} L_h & 0 & -\omega_{RS} L_R & R_R \end{bmatrix} \begin{bmatrix} i_{S\alpha} \\ i_{S\beta} \\ i_{R\alpha} \\ i_{R\beta} \end{bmatrix} +$$

$$\begin{bmatrix} L_S+L_{d\alpha\alpha} & L_{d\alpha\beta} & L_h+L_{d\alpha\alpha} & L_{d\alpha\beta} \\ L_{d\alpha\beta} & L_S+L_{d\beta\beta} & L_{d\alpha\beta} & L_h+L_{d\beta\beta} \\ L_h+L_{d\alpha\alpha} & L_{d\alpha\beta} & L_R+L_{d\alpha\alpha} & L_{d\alpha\beta} \\ L_{d\alpha\beta} & L_h+L_{d\beta\beta} & L_{d\alpha\beta} & L_R+L_{d\beta\beta} \end{bmatrix} \begin{bmatrix} i_{S\alpha} \\ i_{S\beta} \\ i_{R\alpha} \\ i_{R\beta} \end{bmatrix} \tag{1}$$

The inductance terms are defined as follows:

$$\begin{aligned} |i_\mu| &= \sqrt{(i_{S\alpha}+i_{R\alpha})^2+(i_{S\beta}+i_{R\beta})^2} \\ L_h &= L_h(|i_\mu|) \\ L_S &= L_h+L_{S\sigma} \qquad L_R = L_h+L_{R\sigma} \\ L_d &= |i_\mu|\frac{\mathrm{d}L_h(|i_\mu|)}{\mathrm{d}|i_\mu|} \qquad L_{d\alpha\beta} = L_d\sin\gamma\cos\gamma \\ L_{d\alpha\alpha} &= L_d\cos^2\gamma \qquad L_{d\beta\beta} = L_d\sin^2\gamma \end{aligned} \tag{2}$$

This model can be simplified if the IM is at standstill ($\omega_{RS} = 0$). The supposition that the IM is magnetized only in the α-axis is possible since the orientation of the reference frame on the a-phase is arbitrary. Then $u_{S\beta}$, $i_{S\beta}$ and $i_{R\beta}$ can be assumed to be zero as well.
This results in the model:

$$\begin{aligned} \begin{bmatrix} u_{S\alpha} \\ 0 \end{bmatrix} &= \begin{bmatrix} R_S & 0 \\ 0 & R_R \end{bmatrix}\begin{bmatrix} i_{S\alpha} \\ i_{R\alpha} \end{bmatrix} + \begin{bmatrix} L_S+L_d & L_h+L_d \\ L_h+L_d & L_R+L_d \end{bmatrix}\begin{bmatrix} \dot i_{S\alpha} \\ \dot i_{R\alpha} \end{bmatrix} \\ &= \begin{bmatrix} R_S & 0 \\ 0 & R_R \end{bmatrix}\begin{bmatrix} i_{S\alpha} \\ i_{R\alpha} \end{bmatrix} + \begin{bmatrix} L_D+L_{S\sigma} & L_D \\ L_D & L_D+L_{R\sigma} \end{bmatrix}\begin{bmatrix} \dot i_{S\alpha} \\ \dot i_{R\alpha} \end{bmatrix} \end{aligned} \tag{3}$$

The corresponding vector notation is:

$$\begin{aligned} U &= RI + L\dot I \\ \dot I &= -L^{-1}RI + L^{-1}U = f(I(t), U(t)) \end{aligned} \tag{4}$$

If $\mathrm{d}L_h/\mathrm{d}i_{S\alpha}$ and $\mathrm{d}L_h/\mathrm{d}i_{S\beta}$ exists this model can be linearized in the operating point $i_{S\alpha} = I_{S0}$, $i_{R\alpha} = I_{R0}$ and $u_{S\alpha} = U_{S0}$ [11].

$$\Delta\dot I = \left.\frac{\mathrm{d}f(I,U)}{\mathrm{d}I}\right|_{\substack{i_{S\alpha}=I_{S0}\\ i_{R\alpha}=I_{R0}\\ u_{S\alpha}=U_{S0}}}\Delta I + \left.\frac{\mathrm{d}f(I,U)}{\mathrm{d}U}\right|_{\substack{i_{S\alpha}=I_{S0}\\ i_{R\alpha}=I_{R0}\\ u_{S\alpha}=U_{S0}}}\Delta U \tag{5}$$

If the identification is carried out at a stationary operating point of this model it can be presumed that $I_{R0} = 0$ and $U_{S0} = R_S I_{S0}$. Then the linearization leads to the model (6), if $L_{S\sigma} = L_{R\sigma} = L_\sigma$ is valid supplementary.

$$\Delta\dot I = \begin{bmatrix} \dfrac{-R_S(L_{D0}+L_\sigma)}{2L_{D0}L_\sigma+L_\sigma^2} & \dfrac{R_RL_{D0}}{2L_{D0}L_\sigma+L_\sigma^2} \\ \dfrac{R_SL_{D0}}{2L_{D0}L_\sigma+L_\sigma^2} & \dfrac{-R_R(L_{D0}+L_\sigma)}{2L_{D0}L_\sigma+L_\sigma^2} \end{bmatrix}\Delta I + \begin{bmatrix} \dfrac{L_{D0}+L_\sigma}{2L_{D0}L_\sigma+L_\sigma^2} \\ \dfrac{-L_{D0}}{2L_{D0}L_\sigma+L_\sigma^2} \end{bmatrix}\Delta U \tag{6}$$

(7) is the corresponding transfer function.

$$Y(j\omega) = \frac{\Delta i_{S\alpha}(j\omega)}{\Delta u_{S\alpha}(j\omega)} = \frac{1+j\omega L_{D0}/R_R}{R_S+j\omega(1+R_S/R_R)(L_{D0}+L_\sigma)+(j\omega)^2(2L_{D0}L_\sigma+L_\sigma^2)/R_R} \tag{7}$$

3 THE IDENTIFICATION SCHEME

There are many identification schemes which allow the estimation of the parameters of the transfer function (7), for example the recursive least-square method [4],[8], model adjustment technique [1] and frequency response test [10],[12].
In this paper the frequency response test is discussed because this method is robust with regard to measuring disturbances if the transfer function is calculated with:

$$Y_i(j\omega_i) = \frac{\int_0^{nT_i} i_{S\alpha}(t)\,e^{-j\omega_i t}\,\mathrm{d}t}{\int_0^{nT_i} u_{S\alpha}(t)\,e^{-j\omega_i t}\,\mathrm{d}t} \tag{8}$$

$u_{S\alpha}(t)$ is a periodic function with the period T_i and the corresponding fundamental frequency $\omega_i = 2\pi/T_i$.
The small delicateness of this method is well-founded by the orthogonal relations of the trigonometric functions. Apart from disturbances with the fundamental frequency and drifts, all disturbances can be completely suppressed if a lengthy measuring time is chosen.
If the IM is excited sinoidal with $u_{s\alpha}(t) = U_{S0} + \Delta U_{S0}\sin\omega_i t$ equation (8) can be simplified because the denominator is known:

$$\begin{aligned} \Rightarrow Y_i(j\omega_i) &= \frac{2}{\Delta U_{s0}nT_i}\int_0^{nT_i} i_{S\alpha}(t)\,(\sin\omega_i t + j\cos\omega_i t)\,\mathrm{d}t \\ &\approx \frac{2T}{\Delta U_{s0}nT_i}\sum_{k=1}^{(nT_i)/T} i_{S\alpha}(kT)\,(\sin\omega_i kT + j\cos\omega_i kT) \end{aligned} \tag{9}$$

(9) provides the value of the transfer function for only one frequency. Repeated measuring at different frequencies result in a non-parametric process model which is unusable in the view of self commissioning. Therefore in a second step the non-parametric process model will be converted in a parametric process model. This is possible with a least-square algorithm which estimates the parameters of the function:

$$\tilde Y(j\omega) = \frac{1+j\omega b_1}{a_0+j\omega a_1+(j\omega)^2 a_2} \tag{10}$$

The parameter b_1, a_0, a_1 and a_2 will be chosen with the goal to minimize the error function:

$$\begin{aligned} E_1 = &\sum_i\left(\Re\{Y_i\} - \frac{a_2}{a_0}\omega_i^2\Re\{Y_i\} - \frac{a_1}{a_0}\omega_i\Im\{Y_i\} - \frac{1}{a_0}\right)^2 + \\ &\sum_i\left(\Im\{Y_i\} - \frac{a_2}{a_0}\omega_i^2\Im\{Y_i\} + \frac{a_1}{a_0}\omega_i\Re\{Y_i\} - \frac{b_1}{a_0}\omega_i\right)^2 \end{aligned} \tag{11}$$

(11) permits the estimation of the parameters with linear methods. A nonlinear optimization isn't necessary [10]. With the knowledge of b_1, a_0, a_1 and a_2, the calculation of the electrical parameters is no problem.
The available parameter U_{S0} can be used to adjust the operating point concerning the saturation. In this way the curve $L_D(i_\mu)$ can be completely measured. With that, the saturation curve $L_h(i_\mu) = 1/i_\mu\int_0^{i_\mu} L_D\,\mathrm{d}i_\mu$ is known.
In case of $U_{S0} = 0$ this method is identical with the Standstill Frequency Response Test (SSFR) [10],[12].

4 THE INFLUENCE OF THE INVERTER ON THE FREQUENCY RESPONSE TEST

As already mentioned, it is a problem if the stimulation of the IM is carried out with the inverter and no phase voltage detection is available. Fig. 1 shows the schematic structure of a voltage source inverter and Fig. 2 shows the corresponding phase voltage and the driver signals of one phase.
The difference between the reference duty cycle d_a^* and $d_a = T_a/T_S$ is the error $\Delta d_a = -(T_{on} - T_{off})/T_S$ and this results in an error $u_{ESa} = U_D \Delta d_a$ of the phase voltage. T_{off} depends on the phase current (see Fig. 3) and can differ from transistor to transistor. Thus, a different error curve is effective for every inverter and this curve must be determined if the error voltage should be compensated like in [4],[8].

The consequence of the error voltage for the response measuring is:
If the phase current is sinoidal with $i_{Sa} = I_{S0} + \Delta I_{S0} \sin\omega t$, u_{ESa} consists of a d.c. component, a term with fundamental frequency and harmonic terms. The d.c. component and the harmonics will be completely suppressed by using (9). A further examination shows that there is no phase shift between the phase current and the phase voltage error fundamental oscillation. These considerations are also true for space vector quantities and so the consequence for the measured frequency response is:

$$\frac{1}{Y_i(j\omega_i)} = \overbrace{\frac{U^*_{S\alpha}(j\omega_i)}{i_{S\alpha}(j\omega_i)}}^{\text{measurable}} = \frac{U_{S\alpha}(j\omega_i) - U_{ES\alpha}(j\omega_i)}{i_{S\alpha}(j\omega_i)} = \underbrace{\frac{U_{S\alpha}(j\omega_i)}{i_{S\alpha}(j\omega_i)}}_{\text{desired function}} + k_e \qquad (12)$$

The reciprocal of the desired transfer function differs from the measured function in a real constant but the imaginary parts are identical.
The idea of estimating the coefficients of (10) only with the imaginary parts of $1/Y_i$ suggests itself because this method is independent of errors caused by the inverter.
This consideration can be extended to any transfer function and so any inverter-fed linear system can be determined in this way with a high accuracy but maybe not all parameters can be estimated only with the imaginary part.
In this case the stator resistance can't be estimated, because the imaginary part is independent of this parameter. The real part of $1/Y_I$ can be written in a form $R_S + f(j\omega)$, where $f(j\omega)$ is independent of R_S. That means, that R_S can't be separated from the constant k_e if the real and imaginary part are used to identify the parameters, but the estimation of the other parameter is not affected.
This is the reason why the estimation is carried out here with both, real and imaginary part like in (11).

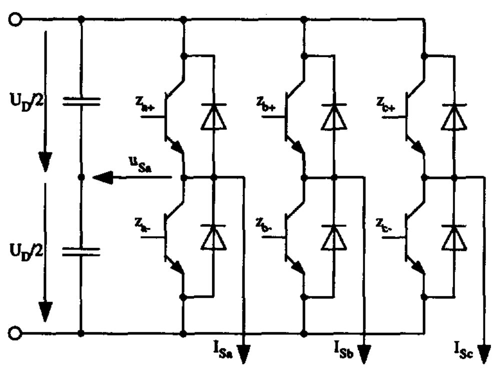

Figure 1: Voltage source PWM–inverter

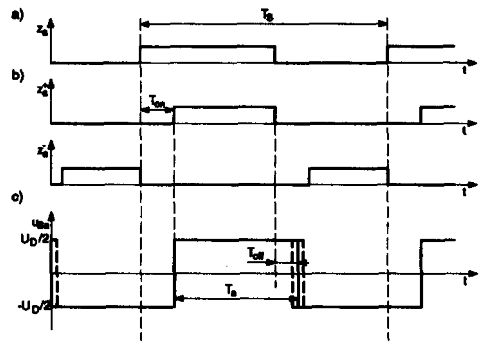

Figure 2: a) Ideal driver signal of phase a
b) Driver signals of phase a
c) Phase voltage, ($i_{Sa}(t) > 0$)

The sum of the effects of R_S and the inverter can be measured with a simple d.c. test with a negligible error.
But if only R_S has to be estimated without a phase voltage detection other schemes couldn't be more accurately. Possibly, if other schemes are used the additional effects of the harmonic terms of the error voltage have to be take in account.

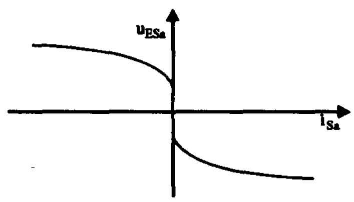

Figure 3: Voltage error curve

5 PROVING OF THE IDENTIFICATION SCHEME

The described identification scheme was proven with a simulation and at a real drive. The parameters of the chosen drive are listed in Table 1. The parameters of the cold machine were determined with a no–load test, a locked–rotor test and a d.c. test. These parameters were used in the simulation, too.

TABLE 1: Parameters of the induction machine

Name–plate–data	Measured electrical parameters
Power: 3 kW	$R_S = 0.22\ \Omega$
Rated voltage: 220 V	$R_R = 0.231\ \Omega$
Rated current: 15 A	$L_\sigma = 1.204$ mH
Speed: 1500 rpm	$L_h = \frac{68.4\text{mH}}{e^{\lvert i_\mu \rvert/16.5}} - \frac{41.5\text{mH}}{e^{\lvert i_\mu \rvert/0.75}} + 4.8\text{mH}$
Frequency: 50 Hz	$J = 0.0124\ \text{kg m}^2$

5.1 Simulation

The employed simulator was presented in [6]. It allows the simulation of a drive with inverter. The current depending switching times and the discontinuous current mode are taken into account. The model of the

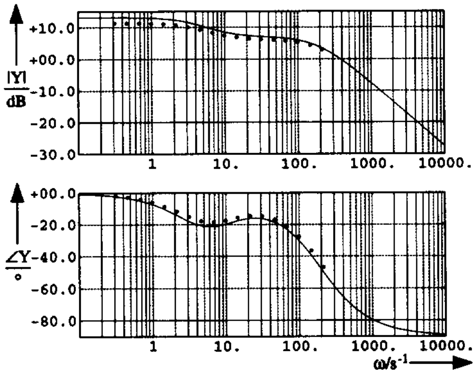

Figure 4: Amplitude and phase of $Y(j\omega)$ and $Y_i(j\omega_i)$

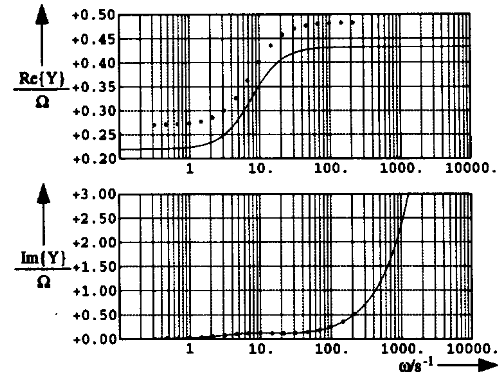

Figure 5: Real and imaginary part of $Y(j\omega)$ and $Y_i(j\omega_i)$

IM is identical with (1) and so no skin effects, iron losses and hysteresis effects are modelled.
With the help of the simulation the following problems were examined:

Choice of the voltage amplitude: There must be compromised about the voltage amplitude. The linearization of the model demands a small amplitude but the quantization of the measured current is more uncritical at large amplitudes.
Good results can be achieved if the magnitude of the voltage is chosen that the current amplitude is about 4.5 A.
Compared with the rated current this is an enormous value, and so the quantization should be no handicap.

Choice of the frequency range: The measured frequency range has an influence on the quality of the estimation. If the measured values are error free a wide range leads to a good estimation. But low frequencies increase the measuring time enormously and high frequencies are questionable in view of skin effects at the real drive.
With the simulation a frequency range about $0.05\ \text{Hz} \leq f \leq 25\ \text{Hz}$ could be determined which allows a good estimation of the parameters. In this range at 18 frequencies the response was measured but a lower number is sufficient.

Influence of the inverter: Actually the constant k_e of (12) isn't constant when the current amplitude changes from frequency to frequency. An effect on the estimation couldn't be found out.
Furthermore the switching behaviour of the inverter seems to be uncritical.

Choice of the Measuring time: Dependent on the period T_i, the measuring time and the waiting period with respect to the transient phenomenon must be varied. At low frequencies a measuring time about two periods is sufficient, but at higher frequencies a measuring time about a higher number of periods is necessary.
Totally, the measuring time is about 5 minutes.

Fig. 4 shows a frequency response measured with the simulation and the curves, calculated with the parameters of Table 1.
The differences are almost caused by switching times of the inverter transistors. On Fig. 5 it can be seen that the errors occur almost in the real part.
With these measured values an estimation of the electrical parameters is possible with a high accuracy. The relative error of the leakage inductance is less than 0.1% for all tested offset currents. In case of the rotor resistance the error is less than 0.5%. The maximum error of the differential inductance is about 2%. The stator resistance can't be separated from the constant k_e and so the error is about 30%.
If the values of L_{D0} measured at different offset currents are used to calculate the magnetization curve $L_h(i_\mu)$ the curve in Fig. 6 is achieved.
The quality of these results is very good and for a self-tuning more than sufficient because the error of R_S is uncritical. If R_S is needed for the controller design or a flux model, it is better to take account both, R_S and the influence of the inverter, than to consider only R_S. The sum is represented well by the estimated value.

5.2 Experimental results

The practical examinations were carried out with a drive in accordance with Fig. 7. The DSP56001 of the NeXT station is used to perform the control, the modulation and the correlation (Equ. (9)). The M68040 is used for the least-square algorithm. The advantage of this drive for the examinations are a powerful interface between the two processors and a comfortable graphical user interface. More detailed informations could be found in [3].
In case of series production this conception is not acceptable in view of the costs, but the workstation can be replaced by a PC which is only necessary at the commissioning phase. By means of the serial interface the PC communicates with the controller which is placed in the cabinet of the inverter. The use of the PC causes no additional costs since usually during the commissioning phase a PC is present anyway. The data transfer between PC and controller is no problem because only a few values have to be transferred.

Examinations. The results of the examinations are as follows:

- Stator resistance: The relative error of R_S is now about 80%. The reason is an additional line and extrinsic resistance which is great compared to the measuring values of Table 1.

- Rotor resistance and leakage inductance: The relative errors of these quantities are about 1%...5% dependent on the offset current. So, the locked–rotor test and the frequency response test deliver nearly the same results and no method should be preferred in view of the accuracy because the results of the locked–rotor test are maybe as well incorrect as the results of frequency response test.

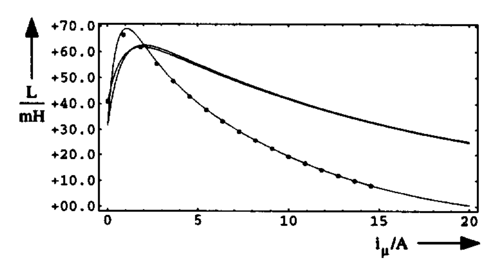

Figure 6: Curves of $L_h(i_\mu)$ and $L_D(i_\mu)$ from the no–load test and the frequency response test

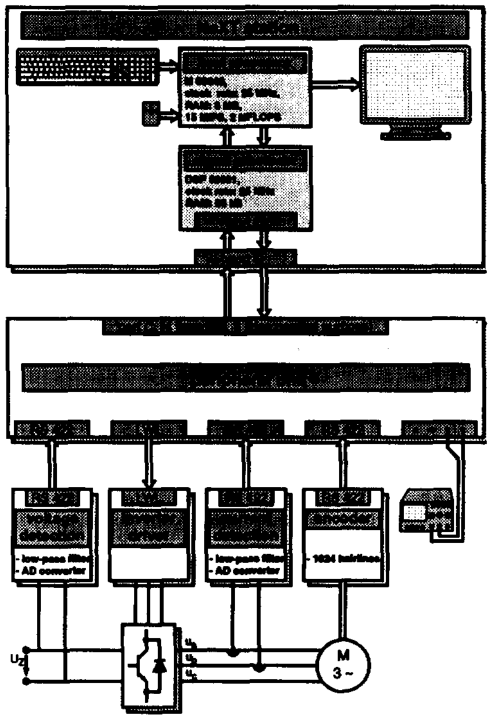

Figure 7: Diagram of the experimental setup

- Differential inductance: The estimated value of L_{D0} is always lower than the measured values of the no-load test. At small offset currents the error is about 45% and at offset currents comparable the rated current the error is about 10%.
  The calculated curve of $L_h(i_\mu)$ contains errors about 40%, too, since the error of L_{D0} at small currents is more critical than the errors at a larger I_{S0}.
  The accuracy of this parameter is consequently not sufficient for a self-commissioning.

The reason of this enormous error can't be found in the identification method or the measuring set-up. The supposition that the model (1) describes the IM not accurately in this case is more plausible.
A closer examination shows that the hysteresis of the iron mustn't be neglected.
Fig. 8 depicts a schematic hysteresis loop of a ferromagnetic material [7]. A major loop, a minor loop and the normal magnetization curve are plotted. Such a minor loop will be described during the frequency

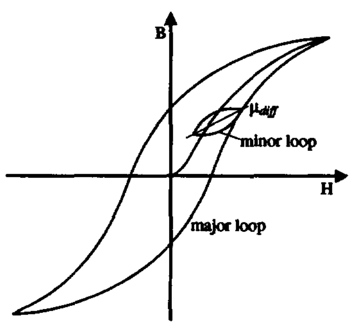

Figure 8: Schematic hysteresis loop

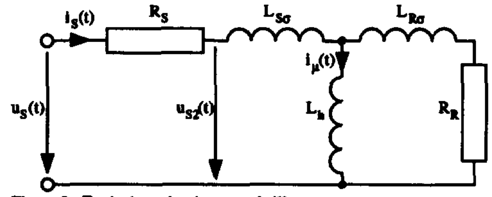

Figure 9: Equivalent circuit at standstill

response test with current offset. The frequency response test estimates the mean differential inductance which is proportional to the differential permeability μ_{diff} of the minor loop. In every operating point μ_{diff} is less than the gradient of the normal magnetization curve. So, the errors of the estimated values of L_{D0} are plausible.

6 A SOLUTION OF THE HYSTERESIS PROBLEM

At the commissioning phase the alloy of the rotor and stator lamination is usually unknown. Thus, a reconstruction of the normal magnetization curve with the measured values L_{D0} seems to be hopeless.
The frequency response test delivers good estimations of R_R and L_σ. The voltage drop across both, R_S and the inverters transistors, can be measured with a simple d.c. test. Therefore a promising solution is to calculate the flux Ψ and the magnetizing current i_μ directly from $u^*_{S\alpha}$ and $i_{S\alpha}$.
For that, the IM is fed sinoidally with 2Hz and without a current offset. In the steady state one period of the current and the voltage are sampled and stored. The voltage u_{S2} (compare Fig. 9) can be calculated with the help of the known voltage drop across R_S and the inverter. Afterwards u_{S2} and $i_{S\alpha}$ can be represented with the first coefficients of their Fourier series. The limitation on the first coefficients suppresses the noise with a higher frequency.

$$u_{2S}(t) = \Re\left\{\sum_{i=1}^{N} u_i \; e^{j\omega_0 it}\right\}$$
$$i_{S\alpha}(t) = \Re\left\{\sum_{i=1}^{N} i_i \; e^{j\omega_0 it}\right\} \tag{13}$$

The flux and the magnetizing current depend linearly on u_{S2} and $i_{S\alpha}$.

$$\Psi(t) = \Re\left\{\sum_{i=1}^{N} \frac{u_i - j\omega_i L_\sigma i_i}{j\omega_i} \; e^{j\omega_0 it}\right\}$$
$$i_\mu(t) = \Re\left\{\sum_{i=1}^{N} \left(i_i - \frac{u_i - j\omega_i L_\sigma i_i}{R_R + j\omega_i L_\sigma}\right) e^{j\omega_0 it}\right\} \tag{14}$$

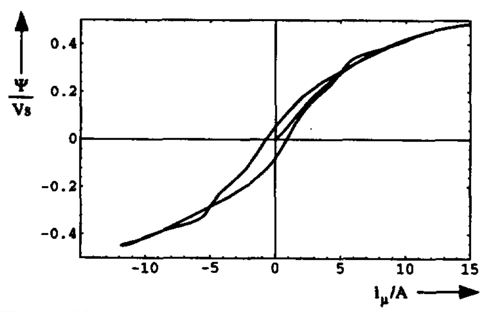

Figure 10: Measured hysteresis loop and the normal magnetization curve measured with the no-load test

Since the signals $\Psi(t)$ and $i_\mu(t)$ are known the curve $\Psi(i_\mu)$ can be calculated, too. The result of one examinations is shown in Fig. 10. The different magnetizing behaviour in dependency of the sign of $d i_\mu / d t$, caused by the hysteresis, is recognizable.
The distortion of one branch of the hysteresis loop appears simultaneously with the zero cross of the phase current. The results at this region can be improved with a more accurately measured voltage drop across R_S and inverter.
The error between the maximum of $\Psi(i_\mu)$ and the curve, determined with the no-load test is very small and this result is valid for other current amplitudes as well.
So, the identification of the whole magnetization curve is possible in this way with a sufficient accuracy.

7 CONCLUSION

Two methods were presented which allow a parameter identification of an induction machine at standstill.
The first method is a frequency response test which delivers very accurate values of R_R and L_σ in spite of the use of an inverter without measuring the phase voltage and in spite of the saturation. With these parameters and the values of $i_{S\alpha}$ and $u_{S\alpha}$ for one period the second method allows the calculation of the magnetization curve.
The accuracy of the determined parameters is more than sufficient for self-commissioning and the requirement of hardware is minimal. So, these methods are suitable for commercial self-commissioning drives.
A further remarkable result is the magnetic hysteresis which can be observed with both methods.

NOMENCLATURE

$d^*_{a,b,c}$	Reference duty cycles of the inverter transistors
$d_{a,b,c}$	duty cycles of the inverter transistors
i_μ	Magnetizing current in the stator reference frame
I_{S0}	Stator current at the operating point
$i_{S\alpha,\beta}$	Stator current components in the stator reference frame
$i_{R\alpha,\beta}$	Rotor current components in the stator reference frame
I_{R0}	Rotor current at the operating point
L_D	Differential inductance $d(\|i_\mu\| L_h) / d\|i_\mu\|$
L_{D0}	Differential inductance at the operating point
L_h	Mutual inductance
$L_{S,R\sigma}$	Leakage inductance
U_D	D.c. link voltage
$u_{S\alpha,\beta}$	Stator voltage components in the stator reference frame
U_{S0}	Stator voltage at the operating point
R_S	Stator resistance
R_R	Rotor resistance
T	Sampling period
$Y(j\omega)$	Transfer function
γ	Angle between i_μ and the α-axis
Ψ	Air-gap flux
ω_{RS}	Electrical angular rotor speed

REFERENCES

1. Boussak, M., Capolino, G. A.,1988, Conf. Rec. ICEM, Pisa,215-220

2. Brown, J. E., Kovacs, K. P., Vas, P., 1983, IEEE Transaction on Power Apparatus and Systems, PAS-102, 96-103

3. Bünte, A., Grotstollen, H., 1992, Conf. Rec. 37. Internationales Wissenschaftliches Kolloquium, Ilmenau, 198-203

4. Heinemann, G., Leonhard, W., 1990, Conf. Rec. Proceedings of IPEC Tokyo, 465-472

5. Klaes, N., 1992, "Identifikationsverfahren für die betriebspunktabhängigen Parameter einer wechselrichtergespeisten Induktionsmaschine", VDI-Verlag GmbH, Düsseldorf

6. Koch, U., Wiesing, J., 1991, "Ein Programmsystem zur Simulation von Drehfeldantrieben", Conf. Rec. 11. Internationale Fachtagung Industrielle Automatisierung - Automatisierte Antriebe, Chemnitz

7. Nitzsche, K., Ullrich, H. J., 1986, "Funktionswerkstoffe in der Elektrotechnik und Elektronik", Hüthig Verlag, Heidelberg

8. Ruff, M., Grotstollen, H., 1993, Conf. Rec. EPE, Brighton

9. Schierling, H., 1988, "Self-Commissioning-a novel feature of Modern Induction Motor Drives", IEE Conference on Power Electronics and Variable Speed Drives, London

10. Soliman, S. A.; Christensen, G. S., 1992, Electric Machines and Power Systems, 20, 123-136

11. Unbehauen, R.,1990, "Systemtheorie", R. Oldenburg Verlag München Wien

12. Willis, J. R., Brock, G. J., Edmonds, J. S., 1989, IEEE Transactions on Energy Conversion, Vol. 4, 608-615