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Speed Sensorless Mixed Sensitivity

Lpv $H_\infty$ Control of the Induction Motor

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Abstract – The paper shows the design of a robust control structure for the speed sensorless vector control of the IM, based on the mixed sensitivity linear parameter variant (LPV) $H_\infty$ theory. The controller makes possible the direct control of the flux and speed of the motor with torque adaptation in noisy environment. The whole control system is tested by intensive simulations and according to the results shows good dynamic and robust performance. Implementation issues based on a DSP TMS320F243 development platform are also presented.

Index Terms – inductor motor, LPV, speed sensorless, gain scheduling control, $H_\infty$, mixed sensitivity.

I. INTRODUCTION

Induction motors (IM) are widely used in the industry due to their simple structure, low cost, and high reliability. Although they are the horsepower of industry, their control is significantly more challenging than of d. c. motors, because as a dynamical system they have a highly nonlinear nature with parameter disturbances. This is the reason, why IMs are still not rival to their d. c. cousins in a number of high precision applications. Nowadays, therefore, there is a great interest in developing high performance and robust controllers to make induction drives unbeatable in all fields of applications. Especially, these efforts concentrate on controllers that do not need speed sensors to operate, which greatly reduces costs and maintenance. (For details see [5, 13]).

Motivated by this goal, we show the design steps of a robust controller for speed sensorless operation of IMs. The designed system gives the opportunity of fast control of the speed of the motor and the magnetic field associated with the rotor flux ($\mathbf{i} = [\mathbf{u}_d, \mathbf{u}_m]^T$). This system also possesses the ability to operate in noisy environment and the online adaptation to the load torque ($T_{\text{load}}$), which is significant for dynamic tasks. The implemented control law is based on the linear parameter variant (LPV) theory of $H_\infty$ control with mixed sensitivity (MS), which has recently appeared in this field [8, 10]. The controller is supported by an I/O linearized reference model and a complex observer synthesized from an extended Kalman filter (EKF) [1, 6] and a $H_\infty$ observer [3, 11]. This structure needs only the measurements of the stator currents, and it shows robustness against system and measurement noises. Moreover, the proposed control law is designed to be easy to tune, that holds the possibility of the online tuning of the performance.

The paper is organized as follows. The LPV model of the induction motor is given in Section 2, the theory of MS LPV $H_\infty$ control in Section 3. The design steps of the controller are given in Section 4 and Section 5 includes the simulated results. The implementation with a digital signal processor (DSP) is presented in Section 6 and finally the conclusions are given in Section 7.

II. LPV MODEL OF THE INDUCTION MOTOR

In case of assuming that every variable is continually distributed inside of the machine and magnetic properties of the rotor are ideal, and using phasor theory to describe the density distribution of the electrical quantities and magnetic fields around the stator and the rotor. (For details see [4, 7]), than the mathematical model of the squirrel-cage IM can be easily derived. The relationship between the flux density, describing the magnetic field, the stator current ($\mathbf{i}$), and the stator voltage ($\mathbf{u}$) can be realized through 2 differential and 2 algebraic equation where the uncertainty of the rotor resistance ($R_r$) introduces nonlinearity into the system. From these equations the (2.1) system follows. This is called the stator oriented ($\alpha, \beta$) model of the IM, without the motion equation.
where the parameters are defined as follows:

\[ \sigma = 1 - L_r^2 / (L_s L_i) , \quad \lambda = L_m^2 / L_s , \quad \tau = L_m / (L_s L_i) , \]

\[ T_r = L_r / R_r , \quad T_s = L_s / R_s \]

and their nominal values are given in Table 1.

If we approximate \( R_r \) with equation (2.3) based on the theory of heating materials (aluminium) with linear convection (\( K_k \)) of heat [12]:

\[
\frac{dR_r}{dt} = \frac{0.86 \cdot R_r}{(245 + T_p) \cdot m \cdot c} (i_{st}^c)^2 \cdot R - K_k (R_r - R_n),
\]

then with the rotor field orientation (RFO) of the phasors [7] the LPV model of the IM is the following:

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix}
\Psi_{st}^r \\
\psi_{al}^r \\
i_{al}^r \\
i_{al}^r 
\end{bmatrix} &= \begin{bmatrix}
-\Psi_{st}^r / T_r - \omega \Psi_{al}^r + L_m i_{al}^r / T_r \\
\omega \Psi_{al}^r - \Psi_{al}^r / T_r + L_m i_{al}^r (t) / T_r \\
\tau \Psi_{st}^r / (\sigma T_r + \omega \cdot \Psi_{al}^r / (\sigma - (\lambda T_r + T_s) i_{al}^r) / \sigma) \\
-\omega (t) \cdot \tau \Psi_{st}^r / (\sigma + \tau \Psi_{al}^r / (\sigma T_r) - (\lambda T_r + T_s) i_{al}^r) / \sigma
\end{bmatrix} + \begin{bmatrix}
0 & 0 \\
0 & 0 \\
1 / \sigma L_s & 0 \\
0 & 1 / \sigma L_s
\end{bmatrix} \begin{bmatrix}
u_{st} \\
u_{al}^r
\end{bmatrix}.
\end{align*}
\]

\[ (2.1) \]

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix}
\Psi_{sl}^r \\
i_{al}^r \\
i_{al}^r \\
i_{al}^r 
\end{bmatrix} &= \begin{bmatrix}
-\psi_{al}^r / L_r \\
\lambda \psi_{al}^r / (L_r, \sigma) - (\psi_{al}^r / L_s + T_s) / \sigma \\
-\omega (t) \psi_{al}^r / \sigma - p_3 \\
- \omega (t) \psi_{al}^r / \sigma - (\lambda \psi_{al}^r / L_s + T_s) / \sigma
\end{bmatrix} + \begin{bmatrix}
0 & 0 \\
0 & 0 \\
1 / \sigma L_s & 0 \\
0 & 1 / \sigma L_s
\end{bmatrix} \begin{bmatrix}
u_{sl}^r \\
u_{al}^r \\
u_{al}^r \\
u_{al}^r
\end{bmatrix}.
\end{align*}
\]

\[ (2.2) \]

III. MIXED SENSITIVITY LPV \( H\infty \) THEORY

3.1 \( H\infty \) theory - From the germinal works of Zames [14] to the highly improved theories of the MS MIMO controls [27, 31], the \( H\infty \) theory has conquered great portion of today’s controller designs with lots of implemented examples [8, 9, 10]. Let us give a brief outline of this theory:

![Fig. 1. General problem definition](image)

Having a general control structure with system \( P \) as in Figure 1, we are searching an optimal, robust, and stabilizing controller \( K \) that minimizes the \( H\infty \) norm of the system:

\[
\|G(P(K))(s)\|_\infty = \sup_{s \in \mathbb{C}} \|G(s)\|, \quad t \in (0, \infty).
\]

This optimization is usually solved by a \( \gamma \)-iteration instead of a direct minimalization. In each recursive step of this iteration we are looking for a controller fulfilling (3.2).

\[
\|G(P, K)\|_\infty < \gamma^*.
\]

In practice, (3.2) is solved based on the Riccati equations or linear matrix inequalities (LMIs). The next \( \gamma \) is computed from the pervious step until we get close enough to its optimal value. It is proved that this algorithm converges and produces a robust controller which is stable and fulfills (3.2) on the whole frequency spectrum [15].

3.2 LPV systems - The LPV systems are such linear systems, where the \( A(.) \ldots D(.) \) matrices in the state-space representation are dependent on a \( \mathbf{p}(t) \in \mathbb{R}^n \) parameter vector.
Further an LPV system can be imagined as a point by point LTI system moving in a n dimensional system space. Supposing that the p vector is bounded (Condition#1) and the system is affine in p (Condition#2), then each of the matrices can be transformed into a form. These systems can be described by an n dimensional cube which can be transformed into a polytope (see Figure 3) defined (existing) on a 2 dimensional system space.

If system P can be represented with such a polytope, and the closed loop system fulfills (3.2) for a given γ than an LPV $H_\infty$ controller can be computed for each corner of the polytope set P. In practice this is done through LMIs. In this way, an LPV $H_\infty$ controller can be designed for the LPV model of the IM, because (2.2) fulfills Condition#1 and Condition#2.

### 3.3 Mixed sensitivity -

By introducing frequency filters (weighting functions) on the I/O signals of the system, not only the model of the IM can be more accurately defined, but the properties of the designed controller can be also directly influenced. The robust stability, disturbance and noise attenuation, and reference tracking of the whole system can be defined, with the frequency definition of the sensitivity function $S = (I + GK)^{-1}$, the inverse sensitivity function $T = I - S$, and the closed loop transfer function $K_S$. For a reference tracking objective, the structure presented on Figure 4 shall be considered. Here, each of the previously mentioned transfer functions are influenced by the $W_S$, $W_T$, $W_{K_S}$ filters, where $W_S$ must be a low-pass filter for good reference tracking, $W_T$ must be a high-pass filter for good noise attenuation, and $W_{K_S}$ must be a high-pass filter for robust stability and disturbance attenuation. Moreover, the presented $W_d$ and $W_r$ should be low-pass filters to define the frequency domain of the input signals. If such a structure considered for a $H_\infty$ control objective, then the $\gamma$-iteration will find a $K$ controller that minimalizes (3.5).

$$\|G(P,K)\|_{\infty} = \begin{bmatrix} W_S & S \\ W_T & T \\ W_{K_S} & K_S \end{bmatrix}$$

(3.5)

For estimation objective by $H_\infty$ observers, in a similar manner a MS structure is given on Figure 5.
IV. SPEED SENSORLESS CONTROLLER DESIGN

To fulfill the above requirements for an IM drive, the controller structure in Figure 6 has been proposed. This structure provides the independent control of the speed and flux based only on the measurement of the stator currents. The measured noisy 3-phase stator currents are transformed to their vectorial representation with the Clark transformation [7], and then they are cleaned from the noise by a complex estimation structure, which is the interconnection of a H∞ observer and a Kalman filter [1, 6]. The Kalman filter provides the estimation of ω and Rr from the nonlinear equations of the model: (2.3), (2.5), and the H∞ observers provides the stator oriented estimation of the rotor flux (Ψr), which is needed for the RFO. After RFO, the input reference signals, Ψref, ωref are transformed to current references, isq, isd, by the help of an I/O linearized model of the IM and based on the previously calculated ωr, Ψr, iad, iqs, and iqr. The H∞ controller gets the deviation isq(error), iad(error), from the current reference and calculates the new voltage phasor, which is transformed to 3-phase by the Space Vector Pulse Width Modulation (SVPWM) element that directly controls the firing impulses of the 3 phase inverter realizing the desired value of the stator voltage for IM.

The H∞ controller was designed with the following MS structure for the LPV model of the IM ((2.2)).

This structure inhabits a very rapid control ability of the flux and the speed through the stator current reference. During the design the sensitivity filter Ws was chosen to be
\[ W_s(s) = \text{diag}(10/(s+10),10/(s+10)) \]
providing good reference tracking on low frequency deviations and preventing the controller to be unnecessary aggressive beyond the cutting-off frequency. The amplification of the filter is only 1dB in the passing region, which gives the possibility to tune the speed and accuracy of the control by external amplification of the current reference signals. Because we greatly reduced the uncertainty by the estimation of Rr, there is no need to choose a dynamic filter for Ws. By trial and error Ws was designed to be
\[ W_s(s) = \text{diag}(0.8,0.8). \]

For Ws the following filter was introduced to restrict the speed of reference tracking which prevents the controller to be unstable even to the step like changes of the reference signals.
\[ W_s(s) = \text{diag}(15/s+15, 15/s+15), \]
Almost the half of T0 for T = Lr/Rr = 43msec for Tw, has provided good tracking without significant overshoots. It is important to note, that because this structure was designed without an integrator than an off-set error is expected. In opposite to the common practice, this makes possible the external tuning of the controller without destabilizing the whole system.

The optimalization was computed through the Matlab function hinfgs which is the part of the LMI toolbox. The resulted controller had 5 states, with two inputs and outputs, and it was described with 2LTI corner systems, with γ = 0.6247. This means, that without external amplification of isq(error) and iad(error), the steady state off-set error is 62.47%.

The I/O linearization of (2.2) gives us the possibility to transform Ψref, ωref into iad, iqs. If the derivatives of Ψref and ω are chosen to V1, V2 virtual inputs equal to Ψref, ωref than the following algebraic equation system provides the reference computation:
\[
\begin{align*}
    i_{d_{r}}^s &= L_{r} \cdot i_{d_{r}}^\alpha / (L_{m} R_{r}) + \Psi_{d_{r}}^\alpha / L_{m}, \\
    i_{q_{r}}^s &= T_{mab} v_{c} / \left( p \cdot \Psi_{r}^\alpha \right) + T_{mab} (T_{mab} + \text{Re}) \omega, \quad \text{where} \quad T_{mab} = 2L_{r} / 3L_{mr},
\end{align*}
\] (4.4) (4.5)

The (4.4) and (4.5) equations can handle the transformation task when \( \Psi_{d_{r}}^\alpha \neq 0 \) which only occurs when the system is at zero energy. At this point, any value can be assigned to the flux in equations (4.4) and (4.5), because this situation exists only for a very short time, during startup.

The flux orientation is handled through the Park transformation [7], to which the needed flux angle is computed from the real and imaginary part of the estimated flux vector. It is clear that for accurate operation we need a very good estimation of the real flux. This is the main reason, why such a complex structure is used for the estimation task. Even in noisy environment, the \( H_{\infty} \) observers are capable for this very accurate estimation because of their low-pass property. Thus for the stator oriented LPV flux model of the IM:

\[
\begin{bmatrix}
    \Psi_{r}^\alpha \\
    \Psi_{r}^\beta \\
    L_{r} \cdot i_{d_{r}}^\beta \\
    L_{r} \cdot i_{q_{r}}^\alpha
\end{bmatrix}
\begin{bmatrix}
    -p_{r} / L_{r} & 0 & 0 & 0 \\
    0 & 0 & 0 & -p_{r} / L_{r} \\
    L_{r} R_{r} / L_{r} & 0 & 0 & 0 \\
    0 & L_{r} R_{r} / L_{r}
\end{bmatrix}
\begin{bmatrix}
    \Psi_{r}^\alpha \\
    \Psi_{r}^\beta \\
    L_{r} \cdot i_{d_{r}}^\beta \\
    L_{r} \cdot i_{q_{r}}^\alpha
\end{bmatrix}
\] (4.6)

the MS structure on Figure 8 was used to calculate an \( H_{\infty} \) observer with the \texttt{hinfgs} function. In this structure we chose the frequency of the nonfiltered deviations to be greater than 300Hz, so the introduced sensitivity filter was

\[
W_{s}(s) = \text{diag} \left( 300 / (s + 30) , 300 / (s + 30) \right). \quad (4.7)
\]

Because any kind of disturbance can shock the system \( W_{s}(s) \) was omitted for wide interval of functioning. The resulted observer had a \( \gamma = 8.49 \cdot 10^{-1} \). Although, this observer calculates the flux vector we still need \( \omega \) and \( R_{r} \). To obtain their value, an EKF is attached to the observer. This EKF based on (2.1), (2.3), (2.5) nonlinear equations, where \( u_{ss}, u_{dl}, T_{load} \) is used as known inputs and \( i_{ss}, i_{dl} \) as the measured outputs of the system. Because of the strong dynamical properties of the resulted model, the prediction phase (see [2]) of this EKF is computed through a 3rd order recursive Adams-Bashforth numerical method [10], and in its correction phase only the diagonal elements of \( Q \) (expected variance of the system noise) and \( R \) (expected variance of the measurement noise) were chosen to be nonzeros. It is not a strict assumption, because there is no significant cross coupling between these noises in the real environment. For this reason

\[
Q = 0, \quad \text{expect} \ Q(i_{ss}, i_{dl}) = Q(i_{dl}, i_{ss}) = 0.0117 h / L_{r}; \quad i, j \in \{0, 1, \ldots, 6\}
\]
(4.8)

\[
R_{i_{ss}, i_{dl}} = R_{i_{dl}, i_{ss}} = 13.85; \quad i, j \in \{0, 1, \ldots, 4\}
\]
（4.9）

where \( h \) is the step size of the numerical algorithm. The whole estimation structure was tuned to be perfectly functioning with only 0.5% of prediction error, while heavy measurement noise (Figure 9), inverter noise (figure 10) and 5% parameter uncertainties was introduced into the Matlab simulations, during the design.

\[ \text{Fig. 9. Modeled measurement noise during design} \]
\[ \text{Fig. 10. Modeled inverter noise during design} \]

V. SIMULATION AND TEST RESULTS

The controller was tested in Matlab with the help of an IM Simulink model. During a very dynamic
task where the load torque changed as in Figure 13, the reference tracking for speed occurred as in Figure 11 when the rotor flux was changing as in Figure 12.

![Fig. 11. Reference tracking for \( \omega \)](image1)

![Fig. 12. Reference tracking for \( \psi_{rd} \)](image2)

![Fig. 13. Change of \( T_{load} \)](image3)

Fig. 13. Change of \( T_{load} \)

Fig. 14 and Figure 15 shows the stator voltages, during the simulation.

![Fig. 14. Change of \( u_{s\alpha} \)](image4)

Fig. 14. Change of \( u_{s\alpha} \)

Looking to these results can be concluded that the controller works well even in rapidly changing load conditions (like at 0.5 msec) and its tracking accuracy and dynamics even for large reference steps (like at 3.5 msec) are good. The controller was also tested for robustness. With 5% of parameter variance the maximum tracking error in speed was no more than 10%.

IV. IMPLEMENTATION WITH TMS320F243

The proposed controller is under implementation on a Digital Spectrum motion control development kit which is powered by a TMS320F243 DSP. This fixed point DSP processor is capable of 20Mips and the processors board contains 8K word Flash ROM. This hardware directly connects to an inverter interface card which produces the PWM signals for a 300Vpp AC capable inverter that empowers the IM seen on Table 1. This interface card also contains analog to digital converters (ADC) which are used to get to know the values of the stator currents and of course is responsible to directly give the PWM signals to the inverter. The connection of the structure is presented on Figure 11.

![Fig. 15. Change of \( u_{s\beta} \)](image5)

![Fig. 16. Implemented AC drive with DSP control](image6)

Fig. 15. Change of \( u_{s\beta} \)

The program of the DSP card is developed in Code Composer. During implementation the continues controller system was discretised with
the Euler method with a chosen step size of 1msec, considered enough to represent the continuous controller. Even with the drawback of the fixed point calculations the implementation shows acceptable performance. (The measurements of the performance will be given in the final paper)

VII. CONCLUSION

In this paper our aim was to show the design steps of a state of the art controller for speed sensorless robust operation of the IM, taking into account the load torque changes without the loss of reference accuracy and (effectiveness) of the whole drive. It is clearly turned out, that with the use of the MS LPV H\textsubscript{\infty} control theory the proposed task can be handled and even implemented on a cheap hardware. However, this structure gives the opportunity of control of a given IM with a parameter variance no more than 5%, its usage would be greatly improved with an online tuning algorithm which is in the focus of our future research.

NOTATIONS

\begin{tabular}{ll}
(A, B, C, D, E) & Matrices of the state space representation of a system \\
IM & Induction Motor \\
MS & Mixed Sensitivity \\
EKF & Extended Kalman Filter \\
DS \texttt{p} & DSP \\
LPV & Linear Parameter Variant \\
RFO & Rotor Field Orientation \\
LTI & Linear Time Invariant \\
PWM & Pulse Width Modulation
\end{tabular}

REFERENCES


\begin{tabular}{|c|c|}
\hline
\textbf{Description} & \textbf{Nominal value} \\
\hline
Lumped stator 3 phase inductance & 0.13 H \\
Lumped rotor 3 phase inductance & 0.13 H \\
Lumped mutual 3 phase inductance & 0.12 H \\
Leakage factor & 0.15 \\
Stator 3p. resistance & 1.86 Ω \\
Rotor 3p. resistance & [3Ω, 6Ω] \\
\textbf{R}_s, \text{ at } T_0 \text{ temperature} & 3Ω \\
Shaft rotor speed & [-85Hz, 100Hz] \\
Rotor flux frequency & [-50Hz, 50Hz] \\
Load torque & [-100Nm, 100Nm] \\
Rotor winding weight & 4 kg \\
Specific heat ct. (Al) & 0.21 \\
Nominal temperature & 18º \\
Moment of inertia & 0.21 J/kgK \\
Number of pole pairs & 3 \\
Fraction & 0.001 \\
Linear heat convection & 3.5 \\
\hline
\end{tabular}
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
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<tr>
<td>(i_{sx}(t))</td>
<td>Stator x phase current [A]</td>
<td></td>
</tr>
<tr>
<td>(u_{sx}(t))</td>
<td>Stator x phase voltage [V]</td>
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<td>(\Psi)</td>
<td>Magnetic flux [Wb]</td>
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<td>(i_{sx}^*(t))</td>
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<tr>
<td>(u_s^*(t))</td>
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