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THE PMSM TRACTION DRIVE – A CONTROL STRUCTURE DESIGN

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Abstract: The aritcle deals with the control structure design for a permanent magnet synchronous motor (PMSM) traction drive. Design and implementation of the speed servosystem under the field weakening conditions. The results of the design were verified on hybrid simulation model in Matlab-Simulink environment.

Keywords: Permanent magnet synchronous motor – PMSM, vector control, torque generator, traction drive, speed servosystem.

1 INTRODUCTION

Permanent magnet synchronous motor – PMSM is a relatively new unit in the area of the traction drives. The PMSM applications in the traction applications started to emerge broadly within the last ten years. The electrical and hybrid drives of the road vehicles were the first such applications. The PMSM applications spread to the city transport track vehicles and to the railway transport vehicles up to 300 kW [1, 3].

The PMSMs are more than twice less in both, the dimensions and the weight, then an asynchronous motor of the same type range (as to the power and the speed). This advantage stands out mainly in the low-floor rail vehicles. The PMSM is then constructed like a low-speed, multipole machine (with more than 40 poles). A higher efficiency, caused by absence of the both, the joul losses in the rotor and the gear losses, is another benefit of the gearless PMSM drive [2].

In comparison with AM drives, the PMSM traction drives are characterized by more complex and more expensive design and also by problematical working out of the emergency conditions caused by impossibility of the field suppression. But in spite of these partial drawbacks, the utilization of the PMSMs is still growing, mainly in the light traction area.

2 DYNAMICAL CONTROL OF PMSM

A PMSM control structure design goes from the demand of the electromagnetic torque dynamical control not only in the constant magnetic flux region, but also in the field weakening region. The nonlinear control methods may be applied, where the field oriented control (FOC) method belongs too.

The principle of the method represents the PMSM torque equation (1), from which results that the torque maximum value is possible to achieve if the equation (3) is valid. The motor torque is a scalar function of the stator current, because $\psi_f = konst$.

The PMSM torque equation contains the synchronizing and reactive torque components. The reactive torque component is zero under constant flux vector control.

$$M_m = \frac{3}{2} p' \left(\psi_d i_q - \psi_q i_d \right) = \frac{3}{2} p' \left[\psi_f i_q + \left(L_d - L_q \right) i_d i_q \right] \tag{1}$$

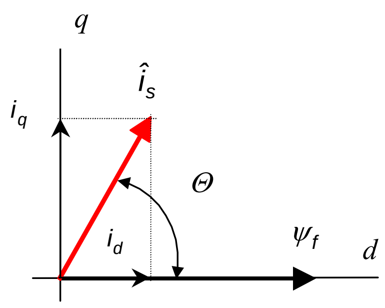

Figure 1: The vector diagram of the stator current represented in the d,q reference frame

The stator current vector represented in the d, q reference frame (see Fig.1), is composed of the torque (i_q) and flux (i_d) components.

$$i_q = i_s \cos\Theta$$
$$i_d = i_s \sin\Theta \tag{2}$$

The whole stator curent can contain only the torque component in the constant magnetic flux mode. It holds:

$$\hat{i}_s = i_q \quad \text{and then} \quad \Theta = 90^0$$
$$i_d = 0 \tag{3}$$

$$M_m = M_{ms} = \frac{3}{2} p' \psi_f i_q = K_m i_q \tag{4}$$

$$K_m = \frac{3}{2} p' \psi_f$$

The PMSM vector control makes possible the field weakening by flux component of the stator current i_d (see Fig.2).

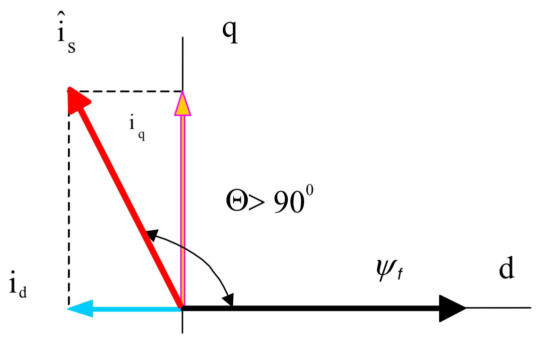

Figure 1: The vector diagram of the both, the stator current and the magnetic flux of the vector controlled PMSM, in the field weakening mode

The PMSM speed control in two regions is the subject of the speed servosystem controls tructure design, see Fig.3:

I – $\omega \in (0, \omega_n)$ – the constant magnetic flux region (or constant torque)

II – $\omega \in (\omega_n, \omega_{max})$ – the constant voltage region (or constant power)

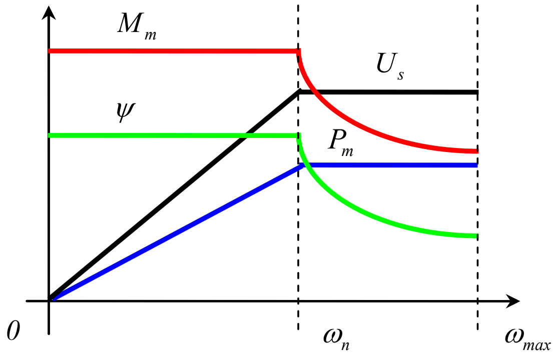

Figure 2: The speed servosystem control range

3 THE PMSM TRACTION SPEED SERVOSYSTEM WITH THE FIELD WEAKENING

There is a block diagram of the PMSM speed servosystem with the field weakening in Fig.4. The PMSM speed servosystem contains an electromagnetic torque generator (GM), with direct vector control. The control loops of the both, the magnetic flux (CCL_d) and the torque (CCL_q) components of the stator current vector, are the part of the electromagnetic torque generator. The magnetic flux control loop (FCL) makes possible a direct control of the magnetic flux vector magnitude ψ_s. The torque control loop (TCL) provides a direct electromagentic torque control. SM observer makes possible the state values control (flux, torque). The speed control loop is superior over the torque control loop.

The speed control loop utilizes an IRC position sensor and Luenberger observer (LO). The voltage control loop (VCL) is superior over the magnetic flux control loop (FCL). The voltage control loop makes possible the field weakening in the constant stator voltage region (U_s = const.). There is the need of the cross couplings compensation block (CB) in the control structure.

The torque and voltage controllers are the adaptive controllers. A design of the adaptive controller will be described in detail in the following section. The controller dynamics is designed from the point of view of the required frequency pass band.

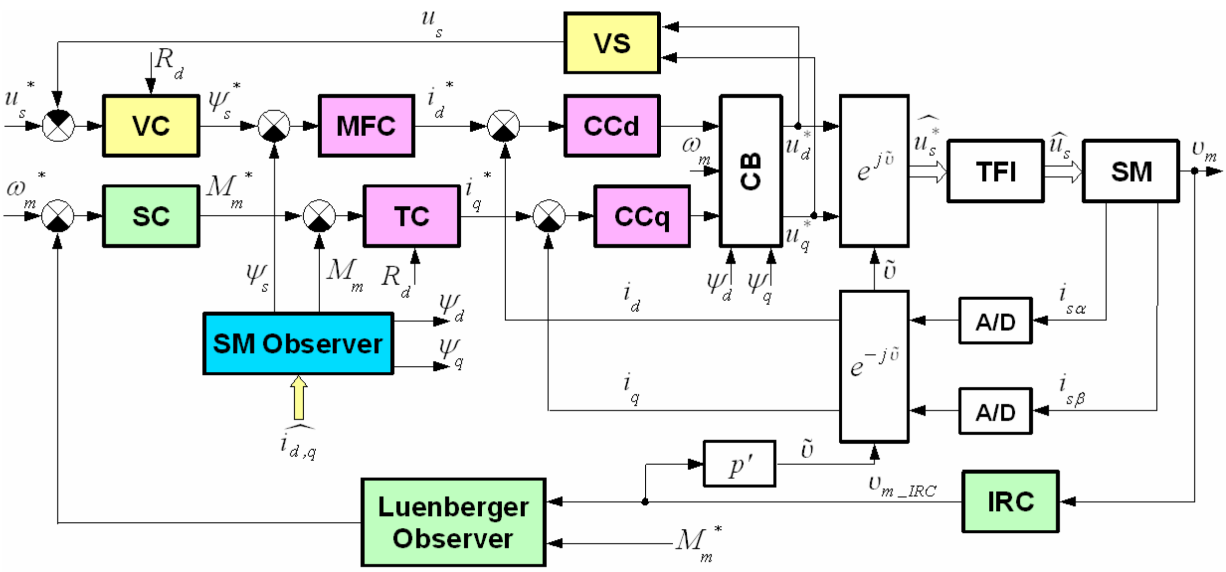

Figure 3: SMPM speed control loop with the field weakening

Key:

VC – voltage controller,

VS – U_s voltage sensor,

MFC – magnetic flux controller,

CCd – controller of the stator current vector magnetic flux component i_d,

CCq – controller of the stator current vector torque component i_q,

TC – torque controller,

SC – speed (velocity) controller,

IRC – incremental rotary encoder, N = 512 imp/rev,

SM – permanent magnet traction synchronous motor,

A/D – analog to digital converter.

PMSM observer from the diagram in Fig.4 is expressed by equations:

$$\begin{aligned} \psi_{sd} &= \psi_f + L_d i_d \\ \psi_{sq} &= L_q i_q \\ M_m &= \frac{3}{2} p' \left(\psi_{sd} i_q - \psi_{sq} i_d \right) \\ \psi_s &= \sqrt{\psi_{sd}^2 + \psi_{sq}^2} \end{aligned} \tag{5}$$

The cross couplings compensation block CB is expressed by equations:

$$\begin{aligned} u_d^* &= u_{d1}^* - L_q i_q \omega \\ u_q^* &= u_{q1}^* + L_d i_d \omega \end{aligned} \tag{6}$$

The design of the controllers for the PMSM stator current vector components

We neglect the cross couplings in the PMSM model within the controllers design. A simplified physical model of the control loop (see Fig.5), does not contain a dynamical model of the frequency inverter. A controlled plant behaviour is described by the first order plan.

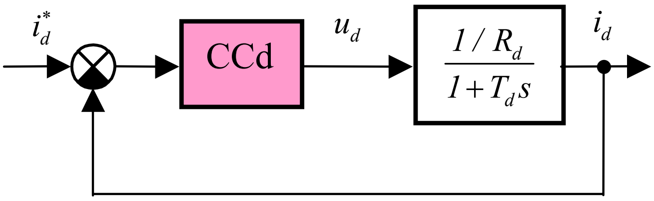

Figure 4: The control loop of the PMSM current i_d

The transfer function of the controlled plant:

$$G_d(s) = \frac{i_d(s)}{u_d^*(s)} = \frac{K_{pd}}{T_d s + 1} \text{ ; where } K_{pd} = \frac{1}{R_d} \text{ and } T_d = \frac{L_d}{R_d} \tag{7}$$

The following transfer function expresses the parameters of the discrete PS controller designed by inverse dynamics method.

$$G_{RP\_d}(z) = K_{id}\left(1+\frac{T_v}{T_{id}}\frac{z}{z-1}\right);\ K_{id} = \frac{2T_d}{K_{pd}\left(2T_{wd}+T_v\right)};\ T_{id} = T_d - \frac{T_v}{2} \tag{8}$$

T_v – sampling period,

T_{wid} – required speed of the control loop of the stator current vector flux component CCL_d.

The design of the current controller CCq is identical with the design of the current controller CCd

The design of the magnetic flux controller *MFC*

The magnetic flux controller *MFC* is superior to the current cottroller *CCd* in the direct vector control structure. The transfer function of the magnetic flux results from the PMSM model. A block diagram of magnetic flux control loop is in Fig.6.

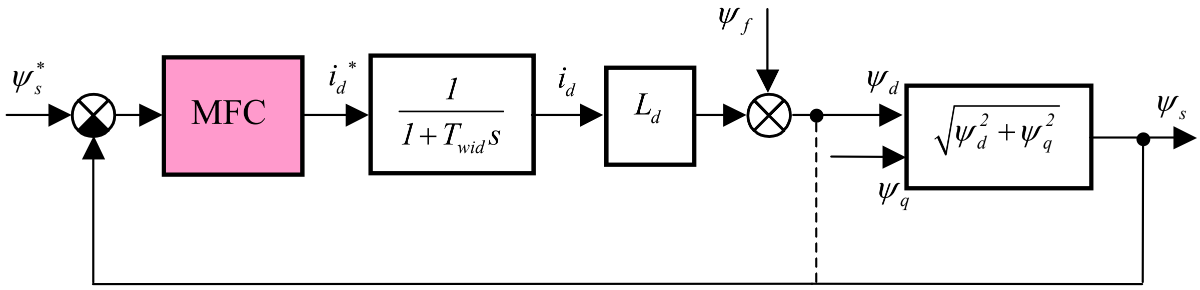

Figure 5: The magnetic flux control loop (MFCL)

A simplified transfer function of the controlled plant:

$$G_{mt}(s) = \frac{\psi_s(s)}{i_d^*(s)} \cong \frac{\psi_d(s)}{i_d^*(s)} = \frac{K_{mt}}{T_{mt}s+1};\quad \text{where } T_{mt} = T_{wid} \text{ and } K_{mt} = L_d \tag{9}$$

The PS magnetic flux controller transfer function designed by inverse dynamics method:

$$G_{RMT}(z) = K_{pmt}\left(1+\frac{T_v}{T_{imt}}\frac{z}{z-1}\right);\ K_{pmt} = \frac{2T_{imt}}{K_{pmt}\left(2T_{wmt}+T_v\right)};\ T_{imt} = T_{wid} - \frac{T_v}{2} \tag{10}$$

T_{wmt} – required speed of the magnetic flux control loop MFCL.

The design of the adaptive stator voltage controller

The stator voltage of the PMSM in steady state is expressed by simplified equation:

$$U_s = U_q \approx \omega\psi_d;\ \text{where}\ \ \omega = R_d\omega_o \tag{11}$$

Where $1 \le R_d\left(t\right) = \dfrac{\omega}{\omega_o} \le \dfrac{f_{s\,max}}{f_{so}} = 3$ expresses the dynamical control range of the field weakening.

Let's adapt the transfer function of the controlled plant in such a way that we exclude any influence of the angular velocity change. Then, the controlled plant is described by transfer function:

$$G(s) = \frac{K_o}{1 + T_{wmt}s}, \; K_o = \omega_o \tag{12}$$

The PS voltage controller parameters were designed by pole placement method and they are expressed by following equations:

$$\begin{matrix} K_{pn} > 0 \\ T_{in} > 0 \end{matrix} \quad \text{if} \quad \omega_0 > \frac{1}{2\xi T_{wmt}} \quad \text{then} \quad \begin{matrix} K_{pn} = \dfrac{2\xi\omega_0 T_{wmt} - 1}{K_o} \\ T_{in} = \dfrac{K_o K_{pn}}{T_{wmt}\omega_0^2} \end{matrix} \tag{13}$$

The voltage control loop VCL in Fig.7, contains the inner magnetic flux control loop with the dynamical properties set by time constant T_{wmt}. Parameter R_d determines the plant gain change and there is possibility to compensate its influence.

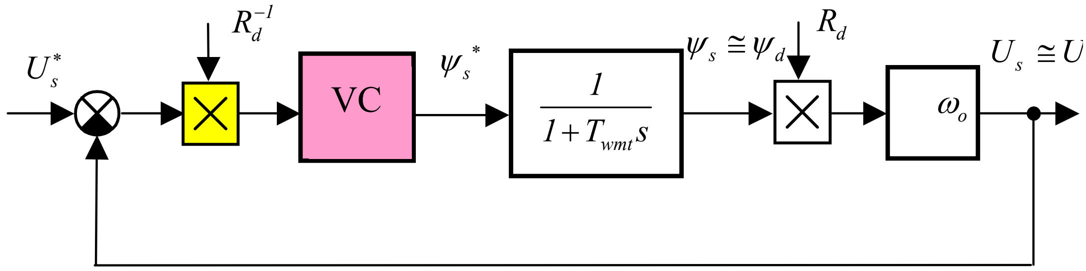

Figure 6: The block diagram of the PMSM adaptive voltage control loop

The design of the adaptive torque controller TC

The torque control loop without any adaptation is in Fig.8. It contains the inner stator current torque component control loop CCL_q.

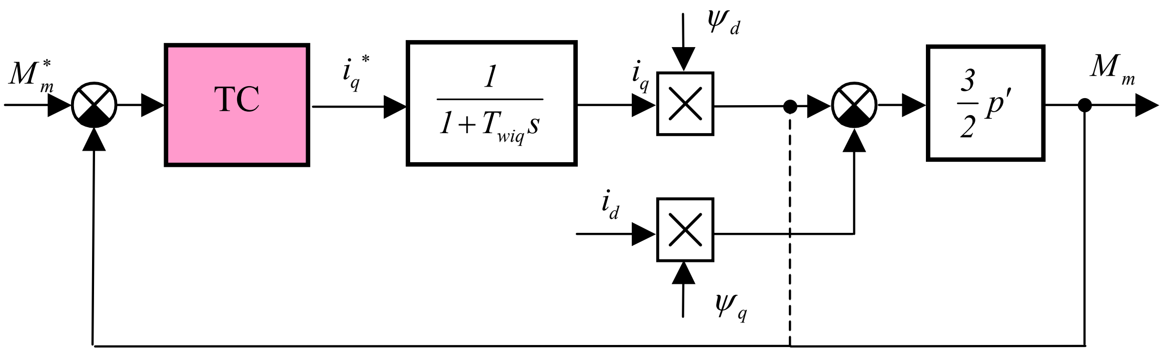

Figure 7: The torque control loop - TCL

There is a possibility to simplify the model of the controlled plant and to take into account only the coupling of ψ_d flux component, in the design. The loop dynamics designates the dynamics of the i_q current control loop with the time constant T_{wiq}. The controlled plant transfer function form will be modified for the purpose to eliminate influence of the magnetic flux variations. The controlled plant transfer function:

$$G_m(s) = \frac{M_m(s)}{i_q^*(s)} \frac{1}{\psi_{so}} = \frac{K_m}{T_m s + 1}\text{; where } T_m = T_{wiq} \text{ and } K_{mt} = \frac{3}{2} p' \tag{14}$$

The transfer function of the PS torque controller which is designed by inverse dynamics method:

$$G_{RM}(z) = K_{pm}\left(1 + \frac{T_v}{T_{im}} \frac{z}{z-1}\right) \quad K_{pm} = \frac{2T_{im}}{K_m\left(2T_{wm} + T_v\right)} \quad T_{im} = T_{wiq} - \frac{T_v}{2} \tag{15}$$

T_{wm} – desired speed of the torque control loop.

The magnitude of the magnetic flux ψ_s varies during the field weakening and (16) applies:

$$\psi_s(t) = \frac{\psi_{s0}}{R_d(t)} \tag{16}$$

The torque control loop with adaptation is in Fig.9.

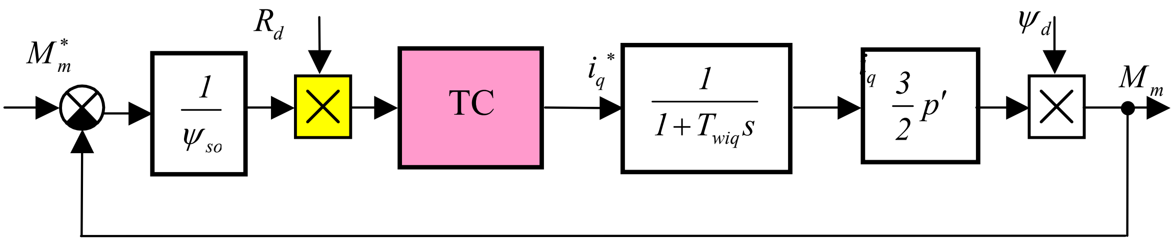

Figure 8: The torque control loop (TCL) with the adaptive controller

IP speed controller design

The speed controller design is simplified because of the fact that the dynamical properties of TCL are expressed by the first order plant.

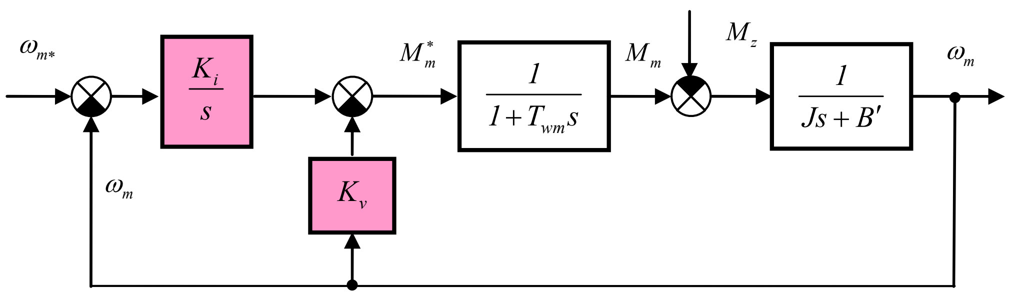

Figure 9: The speed control loop with IP controller

The speed control loop dynamics is of the lower order than the torque generator dynamics and so the torque generator dynamics is possible to neglect. The following K_I and K_V parameters of the IP controller are derived via the pole placement method:

$$K_I = J\omega_0^2 \quad K_V = 2\xi\omega_0 J - B' \tag{17}$$

The pass band ω_0 [rad/s] and damping ξ are the optional parameters.

Design of the Luenberger position observer

The speed control loop utilizes IRC and „tuned filter“ – Luenberger observer (LO). LO alows to get the quality speed signal with low content of the higher harmonics, from a low resolution IRC (N = 512 imp/rev). The speed signal is possible to use in the speed control loop, see Fig.11.

The block diagram in Fig.11 represents a discrete model of the Luenberger observer. The parameters of the PID controller are designed by pole placement method.

The PID algorithm – general form of the notation:

$$K_d = B_{20}\tilde{J}, \quad K_p = B_{10}\tilde{J}, \quad K_i = B_{00}\tilde{J} \qquad \begin{aligned} B_{20} &= \omega_0\left(2\xi + k\right) \\ B_{10} &= \omega_0^2\left(2\xi k + 1\right) \\ B_{00} &= k\omega_0^3 \end{aligned} \tag{18}$$

The pass band ω_0 [rad/s], damping ξ and the coefficient k are the optional parameters.

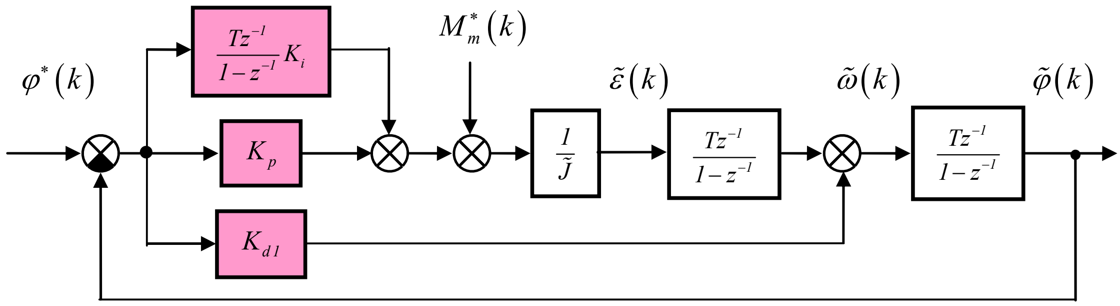

Figure 10: The closed loop observer of the angular speed with direct measurement of the position by digital sensor (forward method)

4 THE SIMULATION RESULTS OF THE PMSM SPEED SERVOSYSTEM WITH THE FIELD WEAKENING

The PMSM parameters:

rated power	Pn = 150 kW
line to line voltage	U = 425 V
phase current	I1 = 332,2 A
rated torque	Mn = 1018,5 Nm
rated frequency	fn = 67 Hz
rated speed	nn = 2010 1/min
pole number	2p = 4

The objective of the experiment is the examination of the sampling period on the contents of the higher harmonics in the PMSM state values, that is the stator current, the stator voltage, torque, magnetic flux and power. The pass bands of the individual control loops are set as follows:

i_d current control loop: f_{wd} = 1 kHz
i_q current control loop: f_{wq} = 500 Hz
Torque control loop: f_{wm} = 250 kHz

Magnetic flux control loop: f_{wmt} =10 Hz
Voltage control loop: f_{on} = 15 Hz
Speed control loop: f_{02} = 2 Hz
Luenberger observer: f_{0e} = 0,2 Hz

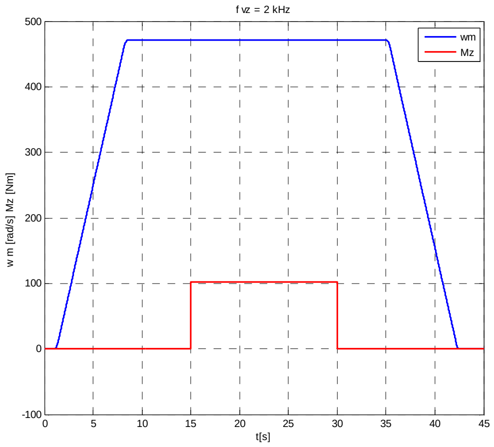

Figure 11: The run up to maximum speed f_s = 150 Hz and run down to the zero speed. The speed reference is generated in the starting unit. The load torque step M_z = 0.1 M_n= 100 Nm is applied in the steady state when the speed is maximum.

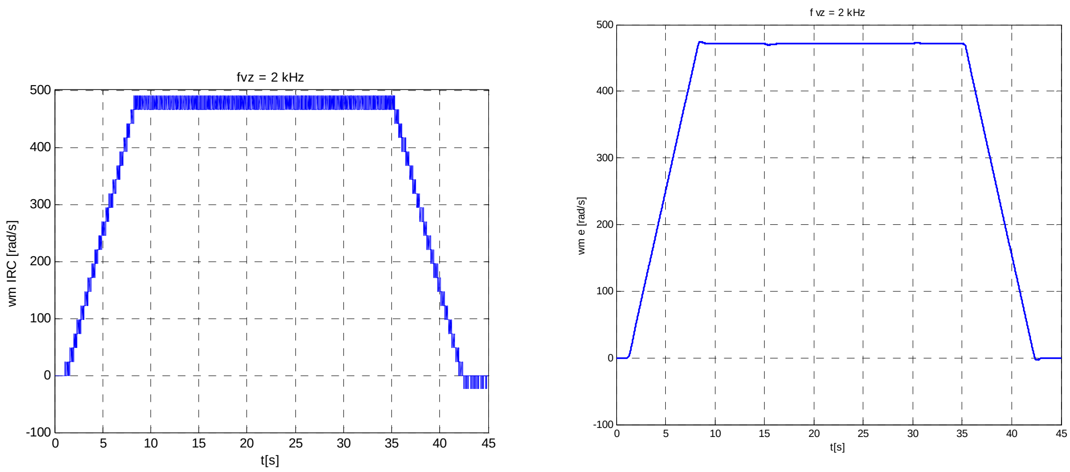

Figure 12: The curve of the angular velocity estimated by Luenberger observer from the position measured by IRC

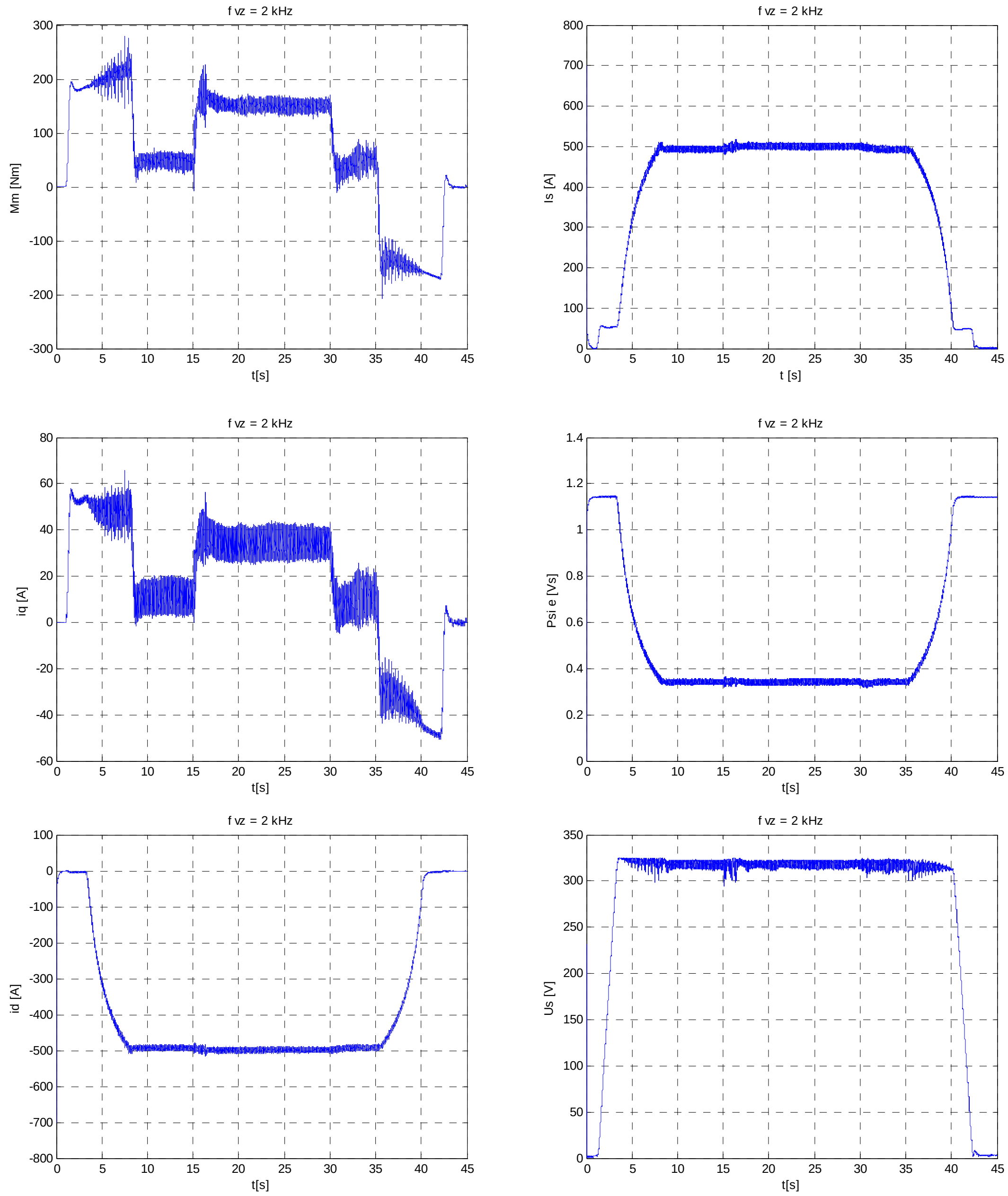

Figure 13: The state values of PMSM

CONCLUSION

The speed control loop which is able to work also in the field weakening area, is designed in this application. The speed servodrive is controlled in the Master-Slave mode. The application of the starting unit is a simplified solution if this control mode. The experimental results confirm the assumption that because of the lower content of the higher harmonics is effective the higher sampling period. The simulation results confirm satisfactory operating behavior of the traction drive.

We will focus on investigation of the influence of the PWM and the change of parameters of the vehicle-load on the quality of control in the following research.

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