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A Simple Closed-Loop Flux-Weakening for Permanent Magnet Synchronous Motors

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Abstract: *This paper presents simple robust integrator-like flux-weakening regulator for permanent magnet synchronous machines, based on general closed-loop flux-weakening which uses reference stator voltage and DC link voltage feedback. Also, some of the open-loop flux-weakening schemes covering constant torque and power regions are briefly discussed and compared to the proposed scheme. It is seen that proposed regulator is independent of motor parameters and has superior performance. In addition, a method for tuning of the regulator gain is defined. Following analysis of the regulator dynamics and setup, the results of extensive simulation and experiment are given.*

Key Words: ***permanent magnet synchronous machines, control, flux-weakening***

1. INTRODUCTION

Permanent Magnet Synchronous Motors (PMSM) are widely used with current-controlled voltage source inverters for industrial and traction applications, because of their high power density, relatively small rotor inertia and high efficiency. In industrial applications, especially servo drives a constant torque operation is desired; whilst in case of traction applications, both constant torque and constant power operations are necessary. Since maximum inverter voltage is limited, PMSM motor cannot operate in speed regions where the back-electromotive force, almost proportional to permanent magnet field and motor speed, is higher than maximum output voltage of the inverter. An obvious consequence of reaching voltage limit is disturbed current dynamics, saturation of current regulators and degraded torque production which may cause system instability. Wide speed range can be achieved with appropriate reduction of rotation field. However, with PMSM motors direct control of permanent magnet flux is impossible. Instead, air-gap flux can be weakened with demagnetizing current in direct axis, which in turn gives indirect Flux-Weakening (FW) of permanent magnets.

2. OPEN-LOOP FLUX-WEAKENING

A PMSM model in synchronously rotating reference frame with electrical angular frequency $\omega_e = p\omega_r$ is considered (d-q coordinates):

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R + sL_d & -\omega_e L_q \\ \omega_e L_d & R + sL_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_e \psi_m \end{bmatrix} \quad (1)$$

$$T_e = 1.5 p i_q [\psi_m + (L_d - L_q) i_d] \quad (2)$$

where v_d, v_q – d, q-axis stator voltages; i_d, i_q – d, q-axis stator currents; R – phase resistance; L_d, L_q – stator d, q-axis stator inductances: ψ_m – permanent magnet flux linkage; s = d/dt; p – number of rotor pole pairs; ω_r – mechanical angular frequency, T_e – electromagnetic torque. Motor current and voltage constraints are

$$i_s = \sqrt{i_d^2 + i_q^2} \le I_{\max} \quad (3)$$

$$v_s = \sqrt{v_d^2 + v_q^2} \le V_{\max} = 2V_{dc}/\pi \quad (4)$$

where I_{max} is maximum inverter current and V_{max} is fundamental component of six-step voltage waveform i.e. situation when inverter enters over-modulation; V_{dc} is the rated DC link voltage. In constant torque region (below base speed), limit (3) is dominant. The goal of current trajectory control for PMSM in constant torque operation is to maximize motor torque with respect to the inverter current capabilities. In papers [1], [2] and [3] q-axis current is considered as main torque producing current, and it is assumed that the speed regulator output is the reference torque value, proportional to i_{qref}. If i_{qref} is expressed with (3) and substituted back to (2), then for dT_e/di_{dref} = 0 d-axis current trajectory which gives Maximum Torque Per Ampere (MTPA) operation is obtained as ($i_s \approx i_{qref}$):

$$i_{dref} = \frac{\psi_m - \sqrt{\psi_m^2 + 8(L_q - L_d)^2 i_{sref}^2}}{4(L_q - L_d)} \quad (5)$$

In case of PM machine with surface permanent magnets, rotor saliency is negligible ($L_d \approx L_q$), hence MTPA reference for d-axis current is zero ($i_{dref} = 0$). However, assumptions that speed regulator output corresponds to reference torque and that it is mainly comprised of i_{qref} make difficult current limiting in case of interior PM machines in MTPA regime, or any PM machine (with interior, inset or surface magnets) in FW regime (when $i_{dref} \ne 0$). A good idea, which provides better analogy between PM and DC machines, has been presented in [4], where the speed regulator output is considered as motor armature current. In case of PM machine that current can be magnitude of stator current

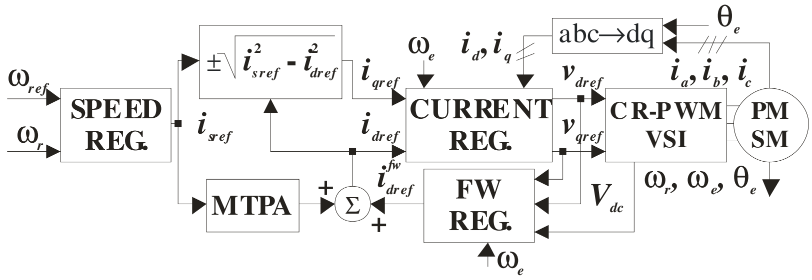

Fig. 1. Block diagram of vector controlled PMSM motor with closed-loop flux-weakening

space vector given in (3). Applying previous method, MTPA trajectory for i_{dref} can be obtained from (5) using i_{sref} instead of i_{qref}, where i_{sref} is magnitude of reference stator current. This way, trajectories of d, q-axis currents are defined with respect to the torque command, and also by clamping the speed regulator output it is possible to achieve simple motor current limiting. During motor operation in constant power region limit (4) is dominant. For steady state ($i_d \approx$ const, $i_q \approx$ const, $\omega_e \approx$ const), and small stator resistance ($R \approx 0$), (1) becomes

$$v_d \approx -\omega_e L_q i_q \tag{6}$$

$$v_q \approx \omega_e (L_d i_d + \psi_m) \tag{7}$$

After using (6)-(7) in (4) voltage limit becomes current and speed dependent.

$$L_q^2 i_q^2 + (L_d i_d + \psi_m)^2 \le (V_{\max} / \omega_e)^2 \tag{8}$$

In case of reference torque output ($T_{eref} \sim i_{qref}$), this yields

$$i_{dref} = -\psi_m / L_d + \sqrt{(V_{\max} / \omega_e)^2 - L_q^2 i_{qref}^2} \tag{9}$$

Alternatively, for reference armature current output, i_{dref} is given with

$$i_{dref} = -\frac{L_d}{2L_{dq}}\psi_m + \frac{1}{L_{dq}}\sqrt{L_{dq}\left[\left(\frac{V_{\max}}{\omega_e}\right)^2 - L_q^2 i_{sref}^2\right] + L_q^2 \psi_m^2} \tag{10}$$

where $L_{dq} = L^2_d - L^2_q$. For $L_d = L_q = L_s$, (10) is written as

$$i_{dref} = -\frac{\psi_m}{2L_s} + \frac{1}{2L_s\psi_m}\left[(V_{\max} / \omega_e)^2 - L_s^2 i_{sref}^2\right] \tag{11}$$

Trajectory (5) is valid for constant torque region, while (9)-(11) are valid for constant power region (between base and crossover speed); where crossover speed is $\omega_c = V_{max} / p\psi_m$. If, for speeds above ω_c, (9), (10) or (11) become complex numbers; then i_{dref} should be limited to $-\psi_m/L_d$, $-L_d\psi_m/(2L_{dq})$ and $-\psi_m/(2L_s)$ respectively, providing that these values are lower than I_{max}. Previous methods are known as open-loop FW methods, where a predefined reference current trajectory is enforced through current regulation, and they have been thoroughly analyzed in [5]. However, these schemes are highly motor parameter dependent, complex, and there is no feedback with respect to the DC link voltage. In practice, to avoid computational burden, open-loop FW schemes are often implemented by using look-up tables (LUT) and current regulators.

3. CLOSED-LOOP FLUX-WEAKENING

The idea of closed-loop FW for interior permanent magnet synchronous machines was originally developed by Jahns [6], who used a phase advance principle and distortion of reference d-axis current trajectory to achieve expansion of speed region. Kim and Sul [4] presented closed-loop FW which uses reference stator voltage from current regulators and available DC link voltage. The difference between those two voltages makes an error which is fed to the FW regulator. Output of FW regulator gives corrective term which is superimposed on reference MTPA d-axis reference. Similar idea, which uses feedforward stator flux correction signal, was introduced by Patel [7]. Harnefors [8] and Wallmark [9] defined procedures for tuning parameters of FW regulators for induction and PM machines, respectively. Practice showed that using proportional term in FW regulator is superfluous, since it increases regulator sensitivity to rapid changes in motor speed and DC link voltage and causes system instability. In most circumstances, usage of plain integrator with limited output is sufficient. System given in Fig. 1 is modelled with the following set of equations:

- Flux-weakening regulator

$$\dot{i}_{dref} = K_{ifw}(V'_{\max} - \sqrt{v_{dref}^2 + v_{qref}^2}) \tag{12}$$

K_{ifw} is FW integral gain; v_{dref}, v_{qref} – current regulator outputs; V'_{max} is boundary between linear modulation and nonlinear (over-modulation) regions set to $V_{dc}/\sqrt{3}$ for wye or $V_{dc}\sqrt{3}/2$ for delta connection and space vector PWM.

- Current regulators with cross-coupling

$$i_{ek} = i_{kref} - i_k, \quad k = d,q \tag{13}$$

$$v_{dref} = K_{pd} i_{ed} + K_{id} I_d - \omega_e L_q i_q - R_d i_d \tag{14}$$

$$v_{qref} = K_{pq} i_{eq} + K_{iq} I_q + \omega_e (L_d i_d + \psi_m) - R_q i_q \tag{15}$$

where K_{pd}, K_{pq} are proportional gains; K_{id}, K_{iq} – integral gains; R_d, R_q – damping gains; i_{ed}, i_{eq} – d, q current errors; I_d, I_q - PI integrating terms defined as

$$\dot{I}_d = i_{ed}; \quad \dot{I}_q = i_{eq} \tag{16}$$

Additionally, motor dynamics is modelled with matrix equations (1). According to [8], if current regulators have optimal setup, one can assume that ($k = d, q$):

$$K_{pk} = \alpha_c L_k; K_{ik} = \alpha_c^2 L_k; R_k = K_{pk} - R \tag{17}$$

$\alpha_c = 0.5 / (T_{PWM} + T_c)$ is inverse current loop time constant; T_{PWM} – PWM period, T_c – current loop sample period. Equation (12) can be linearized around operating point O (v_{kO} – steady state value, Δv_k – small signal value) for

$$v_{kref} = v_k = v_{kO} + \Delta v_k; \quad k = d,q \tag{18}$$

$$V'_{s\max} = v_{dO}^2 + v_{qO}^2 \tag{19}$$

For $\omega_e \approx$ const and using (6)-(7) and (18)-(19) with (12) gives

$$\dot{i}_{dref} = K_{ifw}\left[V'_{\max} - \frac{\omega_e}{V'_{\max}}\left(v_{qO}L_d i_d + v_{dO}L_q i_q + \psi_m\right)\right] \quad (20)$$

After using (1), (13)-(15) and (17) for well tuned current regulators ($v_{dref} = v_d$, $v_{qref} = v_q$), one can obtain

$$\dot{i}_d = \alpha_c i_{dref} - 2\alpha_c i_d + \alpha_c^2 I_d \quad (21)$$

$$\dot{i}_q = -2\alpha_c i_q + \alpha_c^2 I_q \quad (22)$$

If a state space vector is defined as $[x] = [i_{dref}\ i_d\ i_q\ I_d\ I_q]^T$, then model described with (16) and (20)-(22) can be written as state space system $[\dot{x}] = [A][x] + [B][u]$. After applying Laplace transform, system characteristic polynomial can be found as

$$p(\underline{s}) = \det(\underline{s}[I] - [A]) = (\underline{s} + \alpha_c)^3 \cdot \left[\underline{s}^2 + \alpha_c\left(\underline{s} + \frac{K_{ifw}\omega_e L_d}{V'_{\max}}\sqrt{V'^2_{\max} - \omega_e^2 L_q^2 i_{qref}^2}\right)\right] \quad (23)$$

For constant power region, torque is relatively low ($i_{qref} \approx$ 0) and (23) becomes

$$p(\underline{s}) \approx (\underline{s} + \alpha_c)^3\left[\underline{s}^2 + \alpha_c\left(\underline{s} + K_{ifw}\omega_e L_d\right)\right] \quad (24)$$

Inverse time constant for the FW loop α_{fw} is given with

$$\alpha_{fw} = K_{ifw}\omega_e L_d \quad (25)$$

Quality of FW loop response can be characterized with rise time [10], which is related to α_{fw} as $t_{rfw} = \ln 9/\alpha_{fw}$, after which K_{ifw} is defined as

$$K_{ifw} = \frac{\ln 9}{t_{rfw}\, p\, \omega_r L_d} \quad (26)$$

To make (26) more general, it is transformed to

$$K_{ifw} = \begin{cases} \dfrac{\ln 9}{t_{rfw}\, p\, \omega_b L_d}, & |\omega_r| \le \omega_b \\ \dfrac{\ln 9}{t_{rfw}\, p\, |\omega_r| L_d}, & |\omega_r| > \omega_b \end{cases} \quad (27)$$

Unlike in [9], this FW regulator gain is independent of maximum stator voltage V'_{max}. Also, dependence on L_d is negligible since there is no significant variation due to saturation along the d-axis ($L_d \approx$ const). Bandwidths of current and FW loop are, following dominant pole approximation, given respectively

$$BW_c = \frac{0.35}{\ln 9}\alpha_c;\ BW_{fw} = \frac{0.35}{\ln 9}\alpha_{fw} \quad (28)$$

In practice $BW_{fw} << BW_c$ and (24) is simplified further as

$$\begin{aligned} p(\underline{s}) &\approx (\underline{s} + \alpha_c)^3[\underline{s}^2 + (\alpha_c + \alpha_{fw})(\underline{s} + \alpha_{fw})] \\ &\approx (\underline{s} + \alpha_c)^4(\underline{s} + \alpha_{fw}) \end{aligned} \quad (29)$$

Dynamics of FW regulator is mainly governed with quadruple pole in α_c and single pole in α_{fw}. For higher torques in constant power region this dynamics is slower, keeping four poles in α_c, and fifth pole slightly moved rightwards to $x \cdot \alpha_{fw}$ ($x \approx 0.7$). Discrete implementation of FW regulator is shown in Fig. 2, where T_{fw} is FW loop sample time. In order to avoid deep saturation of the integrator when $V'_{smax} > (v^2_{dref} + v^2_{qref})^{1/2}$, a slight modification has been introduced. The interim limit of integrator state and output helps to average i^{fw}_{dref} correction signal towards 0 when FW is not required, and also allows quick response with negative correction signal when FW is necessary.

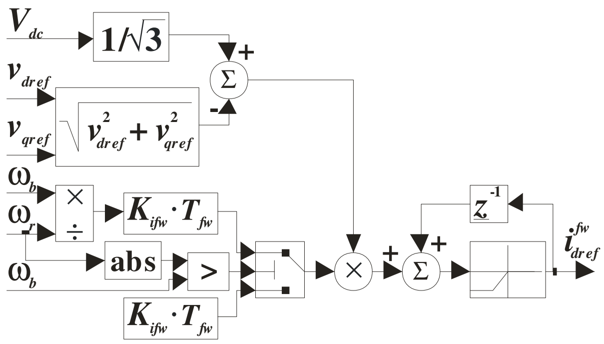

Fig. 2. Discrete implementation of FW regulator

4. EXPERIMENTAL RESULTS

For experimental verification of the proposed FW control scheme a laboratory setup consisting of PMSM motor and 50kW inverter has been used. Motor parameters are T_{en} = 190Nm, n_b = 1500rpm, I_n = 150A, p = 4, R = 32mΩ, L_d = 15mH, L_q = 17mH, ψ_m = 240.6mWb. The motor is driven by 3-phase current-controlled PWM voltage source inverter operating at 10 kHz switching frequency and rated DC link voltage of 340V. All control algorithms for speed, current and FW control have been implemented on TI DSP TMS320F2810 at 150MHz. Bandwidths of speed and current control loops are set to 12Hz and 398Hz, respectively. For FW control loop rise time is set to t_{rfw} = 200ms with BW_{fw} = 1.75Hz. For lower inertia PMSM typically t_{rfw} is 100-200ms, while for larger PMSM motors this time is 200-450ms. Generally, FW loop rise time should be directly proportional to the highest possible motor acceleration rate.

Fig. 3 depicts motor speed and i_d, i_q current waveforms for constant torque region with the motor accelerating from 0 to 1500rpm at 140Nm. As expected for the base speed region, i_q current is dominant an i_d current has relatively small negative value compared to q-axis current. Fig. 4 shows stator current space vector locus in constant torque region. Current space vector moves along MTPA trajectory having i_d = -7A and i_q = 69A in steady state. No noticeable action of FW regulator happened in constant torque region.

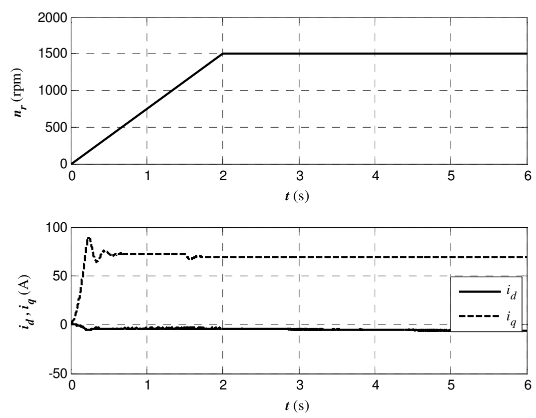

Fig. 3. Motor speed and i_d, i_q waveforms for constant torque region

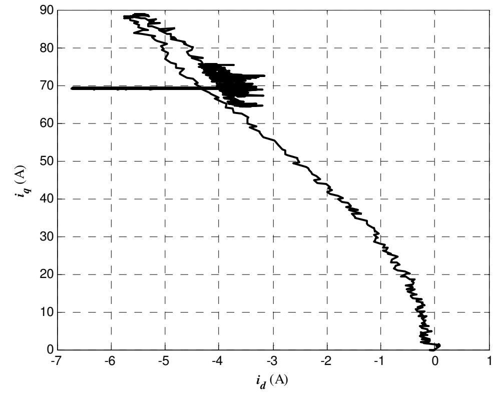

Fig. 4. Stator current space vector locus in constant torque region

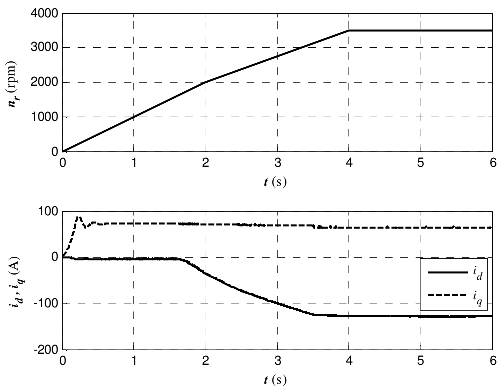

Fig. 5. Motor speed and i_d, i_q waveforms for constant power region

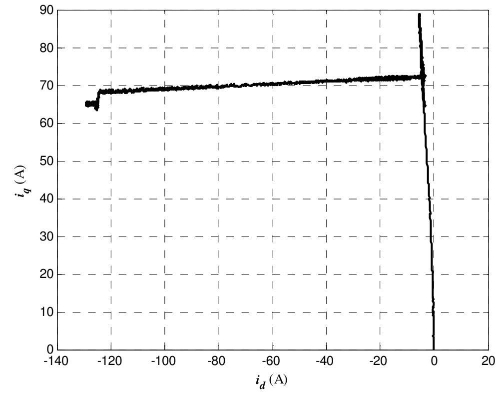

Fig. 6. Stator current space vector locus in constant power region

Fig. 5 shows motor speed and i_d, i_q current waveforms for constant power region with the motor accelerating from 0 to 3500rpm at 140Nm. After reaching speed of approximately 1600rpm, motor also reaches voltage limit and FW regulator starts its action. Direct axis current i_d steadily rises up to -120A allowing motor to double the speed range. As shown in Fig. 6 current space vector moves along MTPA curve and after reaching speed of 1600rpm it smoothly makes transition to the voltage limit ellipse allowing current i_d to rise.

5. CONCLUSION

In this paper, a simple, yet robust closed-loop flux-weakening regulator for permanent magnet synchronous motors is proposed. The important features of this regulator are that it has single gain, it is independent of motor parameters; and that the amount of regulator action is varied with the motor speed. Gain tuning is done with respect to the required rise time or bandwidth of the flux-weakening control loop. The experimental results verify the effectiveness of the proposed scheme. It operates satisfactory both in constant torque and power regions providing expansion of PMSM speed range. Regulator is immune to parameter detuning effect and gives good dynamics performance.

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