Model of a permanent magnet motor with an inverter in the drive system of the vehicle.

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Abstract. The paper presents a mathematical model of a permanent magnet motor, powered by a three-phase source of sinusoidal voltage and a control method. The cooperation of the numerical integration algorithms of the system of differential equations of the motor and inverter has been verified. The results of numerical simulations are presented in graphic form. The work is a supplement to the publication [1], which proposed a model: a battery, a supercapacitor and a method of controlling these energy sources during the driving cycle of the vehicle.

1. Introduction

Most often, a permanent magnet motor is powered by a three-phase sinusoidal voltage, in which the phases and frequency depend on rotation angle of the rotor. Fig. 1 shows a three-phase stator winding connected in a star with four wires from the windings *a*, *b*, *c* and a star point *N*. In the case of three-wire systems, the resistance of the neutral conductor $R_N \to \infty$.

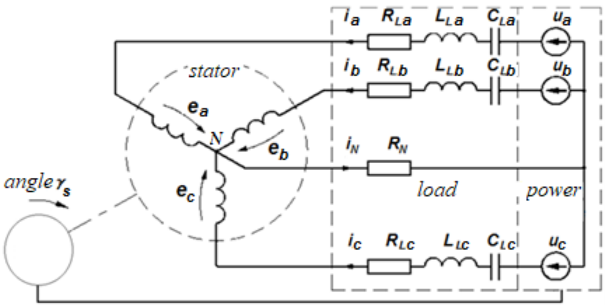

Figure 1. Electric scheme of a permanent magnet motor with supply from a voltage source [2]

As it is shown in Fig. 2, there is a connection between the motor and wheels of the vehicle through elastic elements.

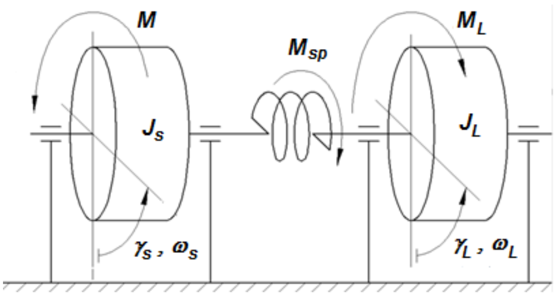

Figure 2. Mechanical model of the motor [2,3]

2. Mathematical model

The model of the motor is made up of the following equations [4]:

Voltage balance $$\mathbf{u} = \mathbf{R}\cdot\mathbf{i} + \frac{d}{dt}\boldsymbol{\psi} \tag{1}$$

Stator linkage flux $$\boldsymbol{\psi} = \mathbf{L}(\theta_e)\mathbf{i} + \boldsymbol{\psi_m}(\theta_e) \tag{2}$$

Electromagnetic torque $$M = p\,(\boldsymbol{\psi}\cdot\mathbf{i}) = p\,(\psi_a\cdot i_a + \psi_b\cdot i_b + \psi_c\cdot i_c) \tag{3}$$

Rotary motion equations $$J\frac{d}{dt}\omega_s + b\omega_s = M - M_o;\ \omega_s = \frac{1}{p}\frac{d\theta_e}{dt} \tag{4}$$

where:

$\mathbf{u} = [u_a, u_b, u_c]^{\mathrm{T}}$ – matrix of phase voltage stator
$\mathbf{i} = [i_a, i_b, i_c]^{\mathrm{T}}$ – matrix of phase currents stator
$\boldsymbol{\psi} = [\psi_a, \psi_b, \psi_c]^{\mathrm{T}}$ – matrix of linkage stator fluxes
$\mathbf{R} = \mathrm{diag}(R_s, R_s, R_s)$ - diagonal matrix of phase stator resistors
R_s - stator phase winding resistance
$\mathbf{L}(\theta_e)$ - general matrix of stator inductance
$\boldsymbol{\psi_m}(\theta_e)$ – matrix of fluxes from permanent magnets
θ_e – angle between the axis *a* of the stator and axis *d* of the rotor in electrical degrees
p – number of magnetic poles pairs in the motor
J - moment of inertia of the drive system reduced to the rotor shaft
ω_s - angular velocity of the rotor
b - sticky friction coefficient
M_o – load torque

In the developed form, the above system of equations can be written as [2]:

$$(R_b + R_{Lb})i_b - (R_a + R_{La})i_a + \frac{d\psi_b}{dt} - \frac{d\psi_a}{dt} + L_{Lb}\frac{di_b}{dt} - L_{La}\frac{di_a}{dt} + u_{Cb} - u_{Ca} + u_a - u_b = 0 \tag{5}$$

$$(R_c + R_{Lc})i_c - (R_b + R_{Lb})i_b + \frac{d\psi_c}{dt} - \frac{d\psi_b}{dt} + L_{Lc}\frac{di_c}{dt} - L_{Lb}\frac{di_b}{dt} + u_{Cc} - u_{Cb} + u_b - u_c = 0 \tag{6}$$

$$(R_c + R_{Lc})i_c + \frac{d\psi_c}{dt} + L_{Lc}\frac{di_c}{dt} + u_{Cc} - u_c + R_N i_N = 0 \tag{7}$$

$$i_N = i_a + i_b + i_c \tag{8}$$

$$i_a = C_{La}\frac{du_{Ca}}{dt} \qquad i_b = C_{Lb}\frac{du_{Cb}}{dt} \qquad i_c = C_{Lc}\frac{du_{Cc}}{dt} \tag{9}$$

$$u_a = U_m\sin(\theta_e + \varphi_a) \quad u_b = U_m\sin(\theta_e + \varphi_b) \quad u_c = U_m\sin(\theta_e + \varphi_c) \tag{10}$$

$$\psi_a = \psi[i_a, i_b, i_c, \theta_e] \qquad \psi_b = \psi[i_a, i_b, i_c, \theta_e] \qquad \psi_c = \psi[i_a, i_b, i_c, \theta_e] \tag{11}$$

$$M = M[i_a, i_b, i_c, \theta_e] \tag{12}$$

$$-J_s\frac{d\omega_s}{dt} - b\omega_s + M - M_{sp} = 0 \qquad \frac{d\gamma_s}{dt} = \omega_s \qquad \theta_e = p\,\gamma_s \tag{13}$$

$$-J_L\frac{d\omega_L}{dt} + M_{sp} + M_L = 0 \qquad \frac{d\gamma_L}{dt} = \omega_L \tag{14}$$

$$M_{sp} = S_{sp}(\gamma_s - \gamma_L) \qquad J_t\frac{E_{sp}}{l_t} = M_{sp} \tag{15}$$

where:
R_a, R_b, R_c – active resistances of the phases of the stator windings
$R_{La}, R_{Lb}, R_{Lc}, L_{La}, L_{Lb}, L_{Lc}, C_{La}, C_{Lb}, C_{Lc}$, - resistances, inductances and capacitances
R_N – resistance of the neutral conductor
u_{Ca}, u_{Cb}, u_{Cc} – voltage on capacitors
u_a, u_b, u_c – phase voltages (for motor mode)
J_S, J_L – moments of inertia of the engine rotor and load
M_L – static moment of load
M_{sp} – torque of elasticity of the shaft connecting the motor with the load
U_m – maximum value of the phase voltage
$\omega_s, \omega_L, \gamma_s, \gamma_L$- angular velocities and rotor and load angles
E_{sp} – elastic of modulus
J_t – moment of inertia of the shaft from the engine to the load
l_t – length of the shaft

The above systems of equations are nonlinear due to their dependencies describing the magneto-mechanical characteristics of the motor. The main problem is the equation of the linkage flux in the stator. Its components depend on the angle of the rotor position and affect the value of the electromagnetic torque. The adoption of a sinusoidal magnetic induction distribution in the air gap allows the transition to the coordinate system dq associated with the rotor, Fig.3, and simplified the model.

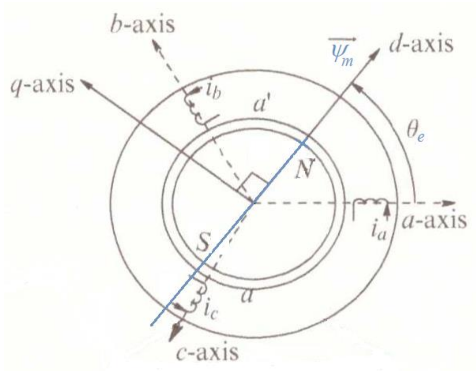

Figure 3. Distribution of currents and magnetic flux in a permanent magnet motor [5]

Taking into account the distribution of currents and magnetic flux as in Fig.4 and the transformation matrix:

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_e) & \cos\left(\theta_e - \frac{2\pi}{3}\right) & \cos\left(\theta_e - \frac{4\pi}{3}\right) \\ -sin(\theta_e) & -\sin\left(\theta_e - \frac{2\pi}{3}\right) & -\sin\left(\theta_e - \frac{4\pi}{3}\right) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{16}$$

determined for the three-wire power supply phases *a*, *b*, *c* of the stator, it is possible to determine dependencies in coordinates dq in the form:

$$\frac{di_d}{dt} = \frac{1}{L_d} u_d - \frac{R_s}{L_d} i_d + \frac{L_q}{L_d} p\omega_s i_q \tag{17}$$

$$\frac{di_q}{dt} = sign\left(\frac{dV}{dt}\right)\frac{1}{L_q} u_q - \frac{R_s}{L_q} i_q - \frac{L_d}{L_q} p\omega_s i_d - sign\left(\frac{dV}{dt}\right)\frac{\Psi_m p\omega_s}{L_q} \tag{18}$$

$$M = 1.5p\left[\Psi_m i_q + \left(L_d - L_q\right) i_d i_q\right] \tag{19}$$

where:
L_q , L_d - inductance in axes q and d
R_s - stator winding resistance
i_q , i_d - current along the axis q and d
u_q , u_d - voltage along the axis q and d
Ψ_m – flux coming from permanent magnet
V – velocity of the vehicle equals to the angular velocity of the motor shaft (constant gear ratio).

The positive *sign* value is for motor work and negative for generator.

The above system of equations illustrates the existing cross-coupling in the currents circuit i_d i i_q and block multiplier, as places of non-linearity in the mathematical description.

In the expression describing electromagnetic torque of the motor, there are two characteristic components: synchronous moment $(\Psi_m i_q)$ coming from the permanent magnets fluxes and reluctance torque $((L_d - L_q)\, i_d\, i_q)$ caused by magnetic asymmetry. A characteristic feature of this asymmetry is the significantly greater reluctance in the rotor's axis d from the reluctance in the q-axis. Depending on the design of the rotor, the component in the q-axis usually has a 5% to 30% share in the resultant torque produced by the motor [2].

Inductances L_q i L_d are the ratio between phase inductance and rotor's position, e.g. inductance measure between phase a and b (phase c is open), are linked by formulas:

$$L_{ab} = L_d + L_q - (L_d - L_q)cos\left(2\Theta_e + \frac{\pi}{3}\right) \qquad (20)$$

$$L_d = \frac{maxL_{ab}}{2} \qquad L_q = \frac{minL_{ab}}{2} \qquad (21)$$

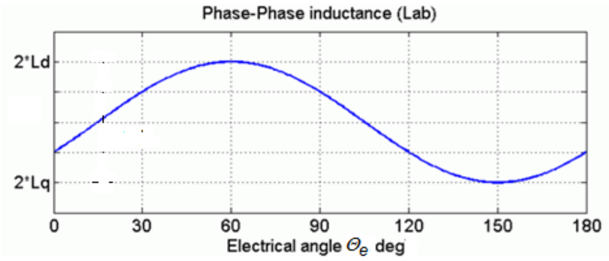

Figure 4. Change in inductance L_{ab} in the function of the angle θ_e [6]

The vector Ψ_m derived from permanent magnets changes as in Fig. 4 and is determined by the amplitude value.

The transition from the actual parameters of phase windings to reduced magnetic couplings in the *dq* coordinate system is given below according to [7].

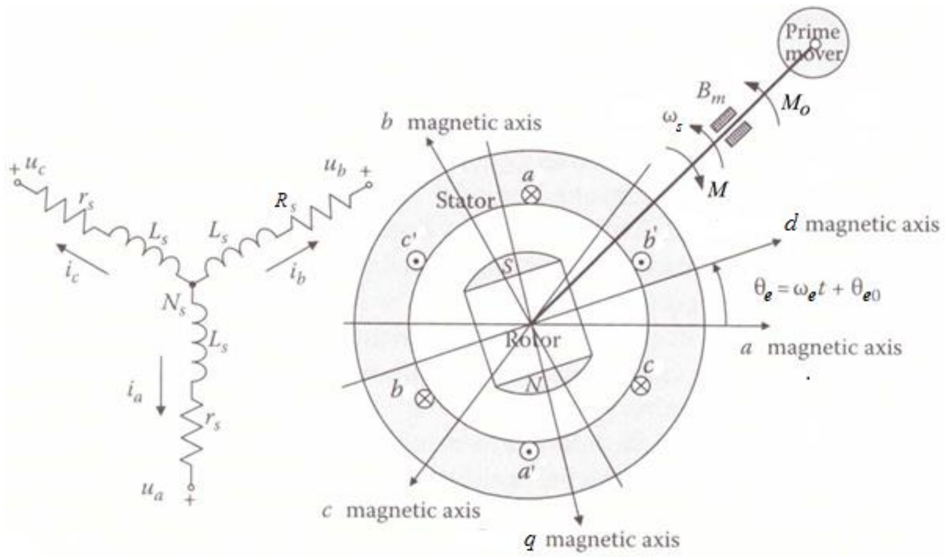

Figure 5. Parameters of three-phase motor winding with permanent magnet [7]

The vector of total magnetic coupling of the stator circuits is defined as:

$$\boldsymbol{\psi} = \boldsymbol{L_s i_{abc}} + \boldsymbol{\psi_m} =$$

$$\begin{bmatrix} L_s + L_m + L_{\Delta m} cos2\theta_e & L_m + L_{\Delta m} cos\left(2\theta_e - \frac{2\pi}{3}\right) & L_m + L_{\Delta m} cos\left(2\theta_e - \frac{4\pi}{3}\right) \\ L_m + L_{\Delta m} cos\left(2\theta_e - \frac{2\pi}{3}\right) & L_s + L_m + L_{\Delta m} cos\left(2\theta_e - \frac{4\pi}{3}\right) & L_m + L_{\Delta m} cos2\theta_e \\ L_m + L_{\Delta m} cos\left(2\theta_e - \frac{4\pi}{3}\right) & L_m + L_{\Delta m} cos2\theta_e & L_s + L_m - L_{\Delta m} cos2\left(2\theta_e - \frac{2\pi}{3}\right) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} +$$

$$\psi_m \begin{bmatrix} sin\theta_e \\ sin\left(\theta_e - \frac{2\pi}{3}\right) \\ sin\left(\theta_e + \frac{2\pi}{3}\right) \end{bmatrix} \tag{22}$$

Where:
L_m – average value of the magnetizing inductance, Fig.6,
$L_{\Delta m}$ - half of the sinusoidal amplitude,
ψ_m – permanent magnet flux.

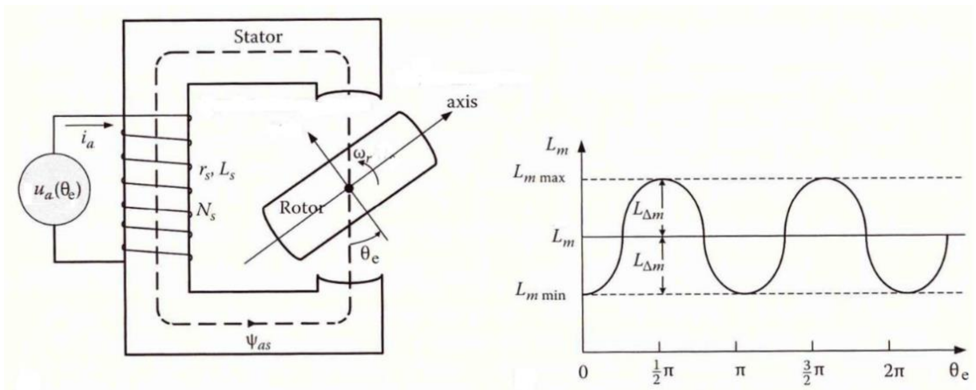

Figure 6. The course of inductance from permanent magnets as a function of angle Θ_e [7]
The physical meaning of L_s and L_m in [5] is presented below Fig.7.

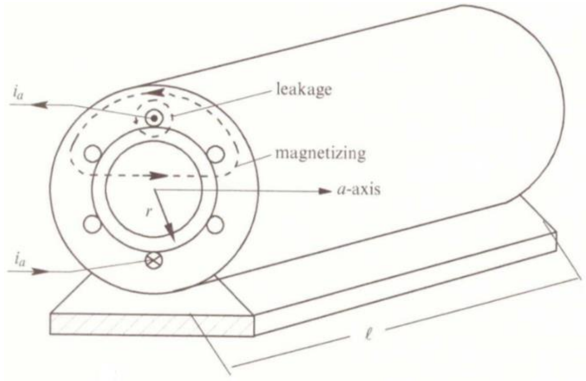

FIGURE 7. Inductances: magnetizing and leakage in the motor stator [5]

The current i_a in phase a, Fig.7, induce two fluxes: magnetizing, which cuts the air gap influencing on rotor and other phases of the stator, and leakage induced only in phase a. Due to this fact stator induction of phase a equals.

$$L_{sa} = L_s + L_{ma}$$

To estimate the leakage inductance L_s there are required test without load with blocked rotor.
Magnetizing inductance L_{ma} is calculated from amount of energy in air gap, then:

$$L_{ma} = \frac{\pi \mu_o r l}{l_g}\left(\frac{N_s}{p}\right)^2 \tag{23}$$

where:
r - average radius in the air gap,
l - length of the rotor along the axis of the shaft,
l_g – gap thickness,
N_s - number of coils in phase,
p - number of magnetic poles pairs in the motor,
μ_o - magnetic permeability.

Taking into account the effect of mutual coupling with the other two phases leads to:

$$L_m = \frac{3}{2}\frac{\pi \mu_o r l}{l_g}\left(\frac{N_s}{p}\right)^2 \tag{24}$$

The reduction of inductance to dq coordinates calculated by formulas [7]:

$$L_d = \frac{3}{2}(L_m + L_{\Delta m}) \quad L_q = \frac{3}{2}(L_m - L_{\Delta m}) \quad oraz \quad L_m = \frac{1}{3}\left(L_q + L_d\right) \quad L_{\Delta m} = \frac{1}{3}\left(L_d - L_q\right) \tag{25}$$

3. Inverter model

In terms of reduction of the ripple of the electromagnetic torque, the motor has to be powered with a phase current that changes in function of the angle Θ_e. The ideal design of the motor and the ideal generation of a sinusoidal current i_{abc} or voltage u_{abc} by inverter is expected to lead to torque without ripples. In practice, it turns out that there are: ripples of torque and vibration. In general, modeling and control with high accuracy do not suppress ripple, which forces the development of various control concepts.

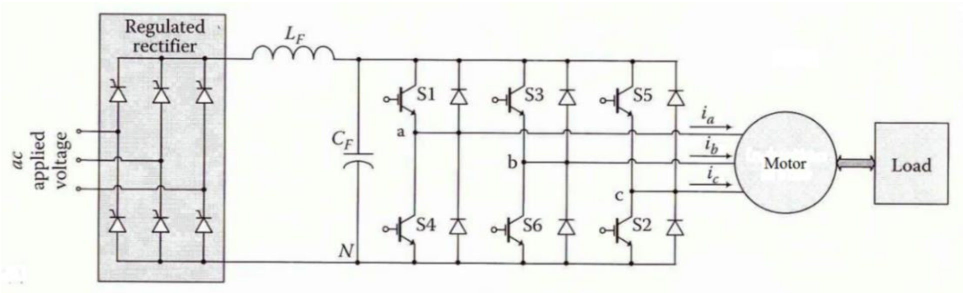

Figure 8. Power inverter with three-phase hard switch [7]

In the case of power supply from the grid through a diode rectifier, the capacitor C_F is charged in the DC circuit, Fig.8, connecting the rectifier with the transistor converter. The larger the capacity of the capacitor, the smaller the variation of the DC voltage, but the higher costs and size, as well as the greater unfavorable effect of the rectifier currents on the external network, exceeding the level allowed by the standards. The L_F induction coil suppresses the peak current [7].

It is common practice to use for the control of inverters the mathematical transformation *α,β,0*, known as Clark's transformation, because it simplifies the analysis of three-phase circuits and allows the generation of a signal used to control the modulation of spatial vectors in inverters in three-phase drives. Transformations [7] are justifiable:

$$\begin{bmatrix} u_\alpha \\ u_\beta \\ u_0 \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \\ 0 & \frac{1}{\sqrt{3}} & -\frac{1}{\sqrt{3}} \\ \frac{1}{3} & \frac{1}{3} & \frac{1}{3} \end{bmatrix} \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} \qquad \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \begin{bmatrix} 1 & 0 & 1 \\ \frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1 \end{bmatrix} \begin{bmatrix} u_\alpha \\ u_\beta \\ u_0 \end{bmatrix} \tag{26}$$

Similar formulas apply to currents. The voltages $u_{\square}$, $u_{\square}$ and currents $i_{\square}$, $i_{\square}$ are orthogonal components of the spatial vectors of voltage **U** and current **I**, same for the motor and the inverter, Fig.9.

The three-phase connection causes that the zero-current component i_o, which is the arithmetic mean of the phase currents, is always equal to zero and the zero components of the u_o voltages do not affect the phase-to-phase voltages. Taking into account this fact, the following formulas result in dependence on phase-to-phase voltages u_{ph} and u_c:

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & -\frac{1}{3} \\ 0 & \frac{1}{\sqrt{3}} \end{bmatrix} \begin{bmatrix} u_{ab} \\ u_{bc} \end{bmatrix} \qquad \begin{bmatrix} u_{ab} \\ u_{bc} \end{bmatrix} = \begin{bmatrix} \frac{3}{2} & -\frac{\sqrt{3}}{2} \\ 0 & \sqrt{3} \end{bmatrix} \begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} \tag{27}$$

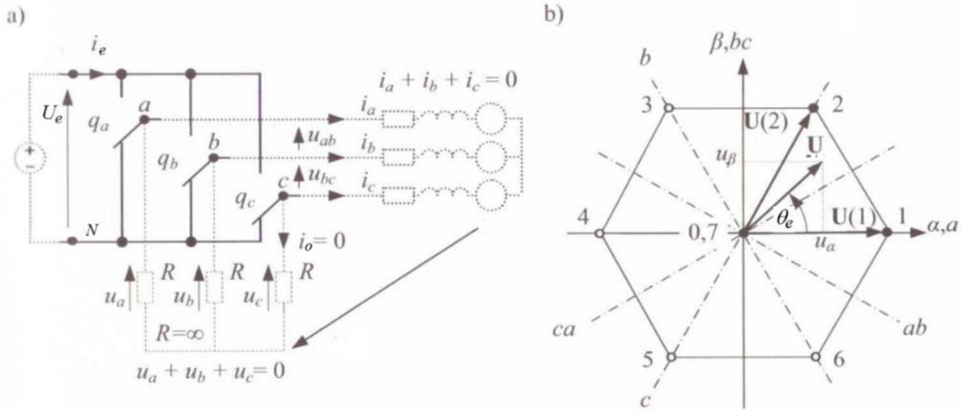

Figure 9. Diagram of an ideal three-phase bridge inverter and a graph of spatial voltage vector [7]

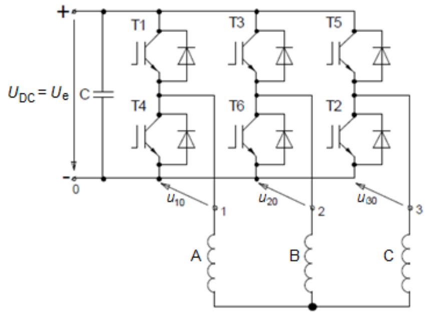

Figure 10. Three-phase bridge connected inverter [8]

Model of inverter with the structure as on Fig.10 is described by formulas [8]:

$$u_a = \frac{1}{3}(2u_{01} - u_{02} - u_{03}) = \frac{U_e}{3}\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(2c_{01n} - c_{02n} - c_{03n})e^{jk\theta_e} \tag{28}$$

$$u_b = \frac{1}{3}(2u_{02} - u_{03} - u_{01}) = \frac{U_e}{3}\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(2c_{02n} - c_{03n} - c_{01n})e^{jk\theta_e} \tag{29}$$

$$u_c = \frac{1}{3}(2u_{03} - u_{01} - u_{02}) = \frac{U_e}{3}\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(2c_{03n} - c_{01n} - c_{02n})e^{jk\theta_e} \tag{30}$$

where:

$$\begin{aligned} u_{01} &= \frac{U_e}{2} + r\frac{U_e}{2}\sin(\theta_e) \\ u_{02} &= \frac{U_e}{2} + r\frac{U_e}{2}\sin\left(\theta_e - \frac{2\pi}{3}\right) \\ u_{03} &= \frac{U_e}{2} + r\frac{U_e}{2}\sin\left(\theta_e + \frac{2\pi}{3}\right) \end{aligned} \tag{31}$$

r - DC voltage regulation coefficient equals to the ratio of the expected voltage amplitude to the DC input voltage (supply)

m - modulation factor; when it is high enough, the difference between real and discrete values is irrelevant

U_e – DC inverter input voltage

The output voltage from the inverter for the first phase can be mathematically expressed in Fourier's series as:

$$\begin{aligned} u_{01} &= U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}c_{01n}e^{jk\theta_e} \\ c_{01n} &= \frac{1}{j2k\pi}\left(e^{-jk\alpha_{01n}} - e^{jk\beta_{01n}}\right) \qquad dla\ k \neq 0 \\ c_{01n} &= \frac{\beta_{01n} - \alpha_{01n}}{2\pi} \qquad dla\ k = 0 \end{aligned} \tag{32}$$

Similarly:

$$u_{02} = U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}c_{02n}e^{jk\theta_e} \qquad u_{03} = U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}c_{03n}e^{jk\theta_e} \tag{33}$$

and phase-to-phase voltages:

$$\begin{aligned} u_{12} &= u_{01} - u_{02} = U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(c_{01n} - c_{02n})e^{jk\theta_e} \\ u_{23} &= u_{02} - u_{03} = U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(c_{02n} - c_{03n})e^{jk\theta_e} \\ u_{31} &= u_{03} - u_{01} = U_e\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(c_{03n} - c_{01n})e^{jk\theta_e} \end{aligned} \tag{34}$$

Taking into account Clark's transformation:

$$u_\alpha = \frac{1}{3}(2u_a - u_b - u_c) = \frac{U_e}{3}\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(2c_{01n} - c_{02n} - c_{03n})e^{jk\theta_e} \tag{35}$$

$$u_\beta = \frac{1}{\sqrt{3}}(u_b - u_c) = \frac{U_e}{\sqrt{3}}\sum_{k=-\infty}^{\infty}\sum_{n=1}^{m}(c_{02n} - c_{03n})e^{jk\theta_e} \tag{36}$$

The above transformations are used to build control systems as in Fig. 11 and 12. In both cases, the motor model in coordinates *dq* and PI controllers in the current circuits i_d and i_q were used. They differ in the way of determining the values of currents and appropriate voltages.

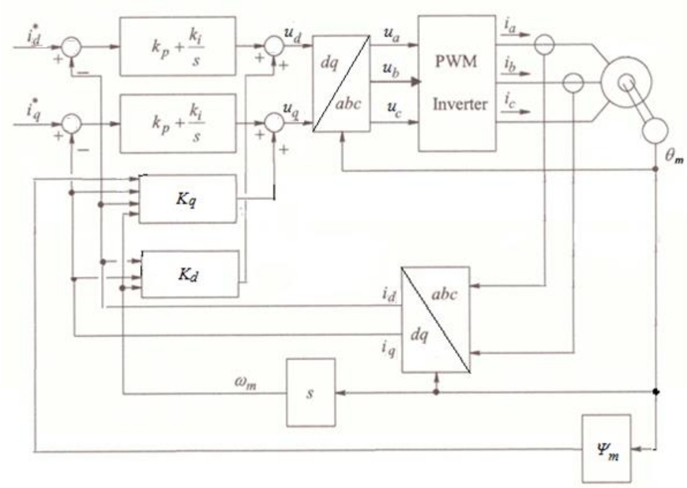

Figure 11. Schematic diagram of the motor control algorithm according to [5]

The following symbols have been introduced in the diagram above:

$$u_{do} = R_s i_d - L_q p \omega_s i_q + L_d \frac{di_d}{dt} = K_d + L_d \frac{di_d}{dt} \tag{37}$$

$$u_{qo} = R_s i_q + p \omega_s (L_d + \psi_m) + L_q \frac{di_q}{dt} = K_q + L_q \frac{di_q}{dt} \tag{38}$$

As shown in Fig. 11, the transformation of signals in two directions is used in the control system: *dq→abc*

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_e) & -\sin(\theta_e) \\ \cos\left(\theta_e + \frac{4\pi}{3}\right) & -\sin\left(\theta_e + \frac{4\pi}{3}\right) \\ \cos\left(\theta_e + \frac{2\pi}{3}\right) & -\sin\left(\theta_e + \frac{2\pi}{3}\right) \end{bmatrix} \begin{bmatrix} u_d \\ u_q \end{bmatrix} \tag{39}$$

and in the reverse direction *abc→dq:*

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta_e) & \cos\left(\theta_e - \frac{2\pi}{3}\right) & \cos\left(\theta_e - \frac{4\pi}{3}\right) \\ -sin(\theta_e) & -\sin\left(\theta_e - \frac{2\pi}{3}\right) & -\sin\left(\theta_e - \frac{4\pi}{3}\right) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{40}$$

In the algorithm shown in Fig.12, the signal given by the driver ω_{ref} together with the actual velocity ω_s gives the error, which after passing the proportional-integral regulator is the set value of the $i_{q\text{-}ref}$ current for the adder in the *d* axis. From the outputs of the current regulators i_d and i_q by applying the limiters, the voltage set in the *dq* system, taking into account the measured instantaneous angle θ_m converted to the electrical angle θ_e, is transformed into the stationary coordinates system a,b,c.

The PI regulators in Fig.11 directly affect the changes of fluxes associated with the inductances in the axes *q* and *d*.

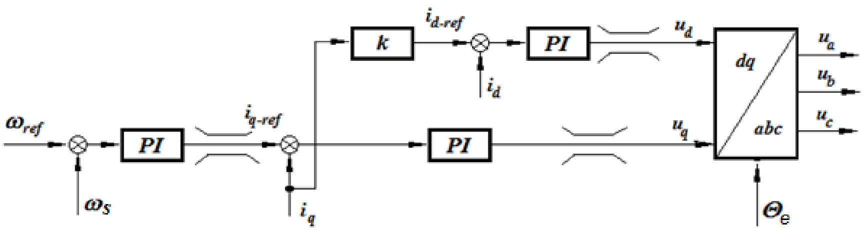

Figure 12. Block diagram of the control algorithm according to [6]

4. Sample results

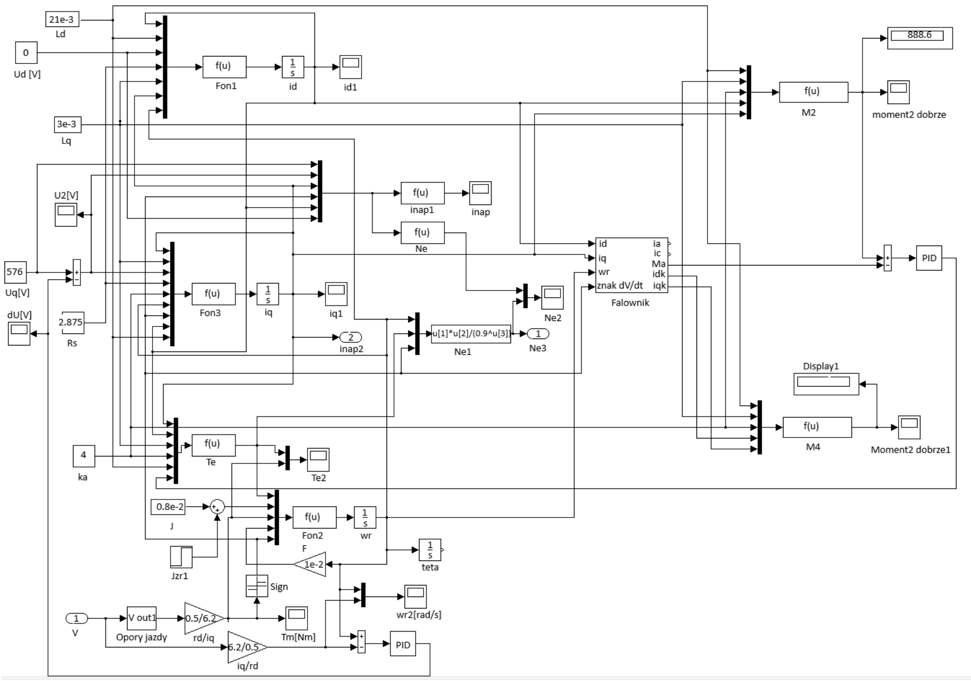

Figure 13. Model of motor with inverter

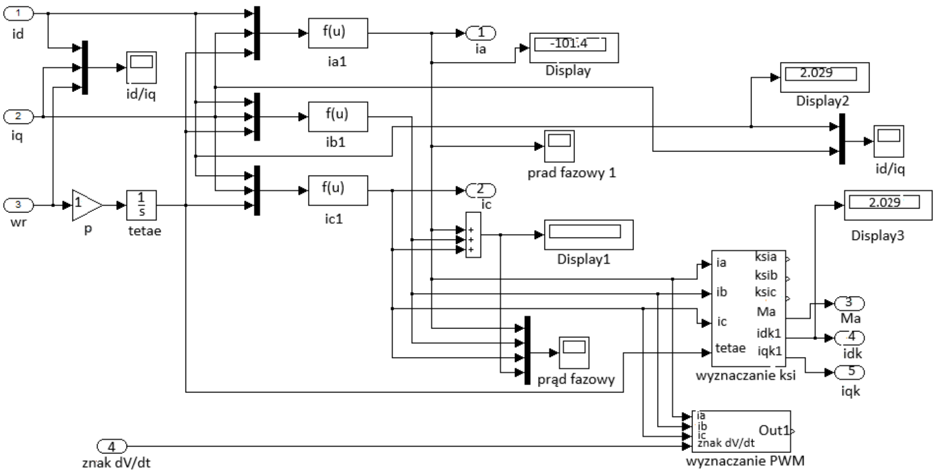

Figure 14. Inverter model - determination of phase currents and stator linkage fluxes

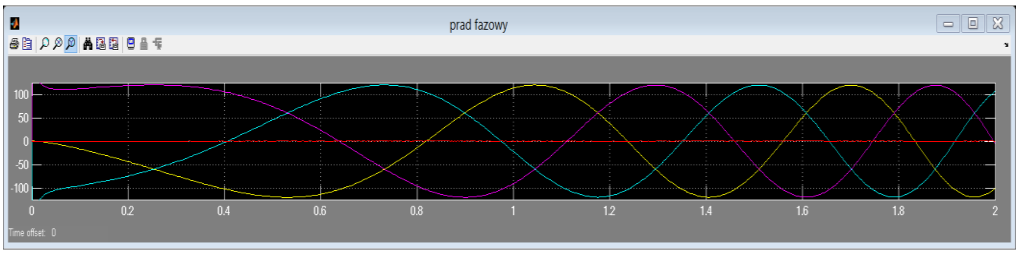

Figure 15. Phase currents $i_{a,b,c}$

The model and the waveforms $i_{a,b,c}$ shown in Fig. 14 and 15 are the result of mathematical relations discussed earlier. Amplitude values of phase currents correspond to the torque at the motor shaft. The initial period of 2 seconds is shown to check the phase shift. Fig.17 and 18 are the presentations of PWM signals determined according to the diagram from Fig.14 and 16.

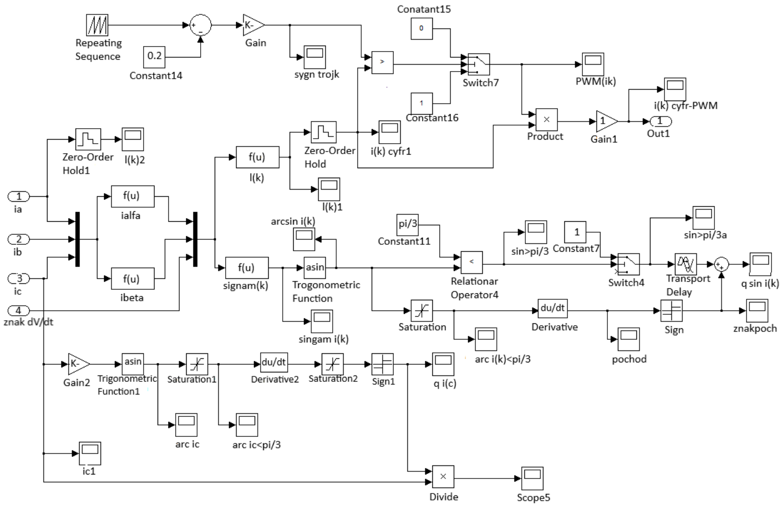

Figure 16. Inverter model - determination of PWM signals for phase currents and stator linkage fluxes

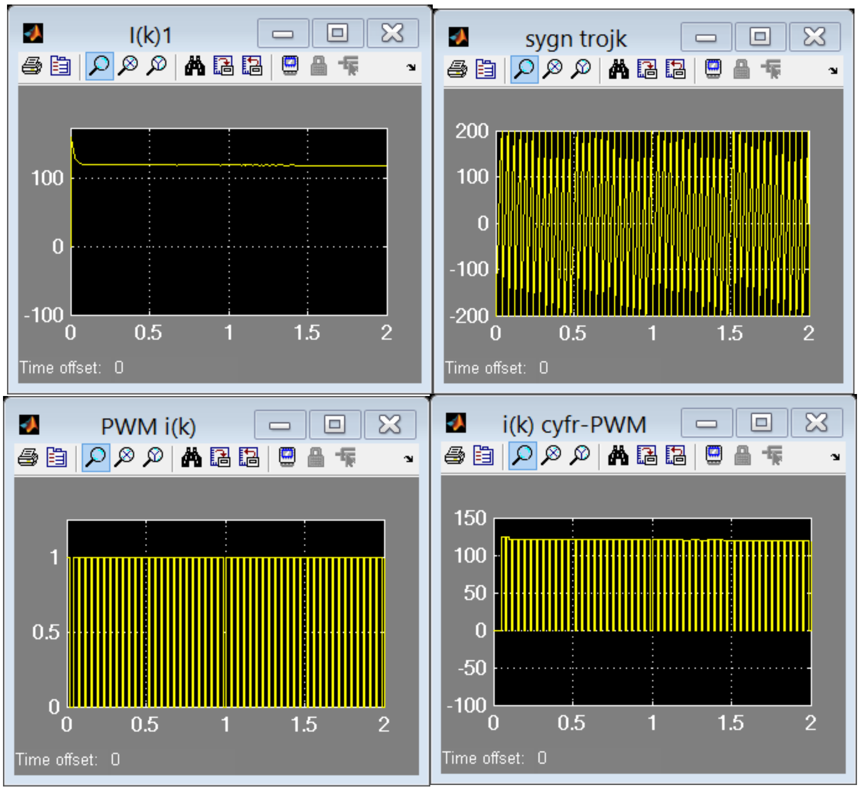

Figure 17. Inverter model – determination of the spatial current

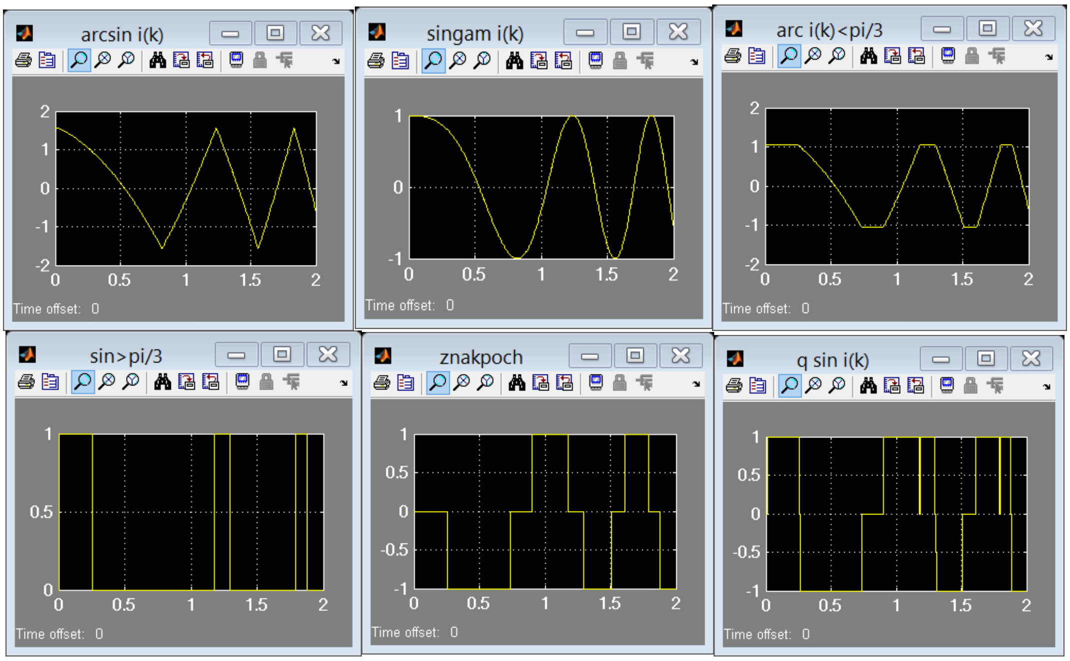

Figure 18. Inverter model - determination of operating states q_{ik} of switches for spatial current

Examples of determination of operating states of switches in a 3-phase inverter according to the scheme from Fig. 9 are presented in Fig. 19. The attached control signals are only special cases of vector modulation, indicating what tasks resulting from the above formulas and descriptions must be solved by inverters. An overview of modulation methods can be found in the extensive literature on converters.

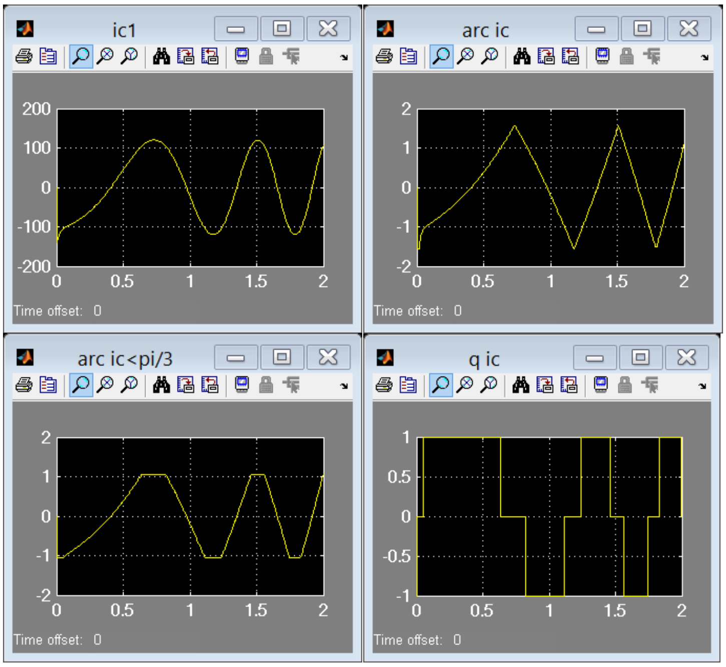

Figure 19. Inverter model - determination of the operating state of q_{ic} switches for phase current i_c

5. Conclusions

In vector control, the mathematical model of a synchronous machine in the *dq* coordinate system is the most common one. The most important feature of this control is the fact that the component i_q of the rotor current vector determines the size of the motor torque, and the component i_d the magnitude of the magnetic flux.

The basic task of the inverters is to create such currents iabc or voltage uabc to obtain the torque without ripple. It leads to development of different control concepts for achieving this goal, which are related to the modeling of magnetic fluxes in stator and inverter.

The presented relations describing models of motor and inverter allow development of control algorithms for electric drives of various configurations, even at the stage of theoretical considerations. It is obvious that the algorithm should also take into account the system of protection against overloads, e.g. electric and temperature, that may occur in the electrical system. Many of these elements already exist in the default software of inverters and hence the "insight" in its algorithm is necessary for proper configuration of the so-called "overlay".

References

[1] Pawelski Z.**,** Zdziennicki Z. 2017 Vehicle electric drive system model. *Journal of KONES Powertrain and Transport*, Vol. 24, No. 3, p.211-220
[2] Lis M., Makarchuk O. 2013 Model matematyczny elektromechanicznych stanów nieustalonych silnika typu BLDC zasilanego ze źródła napięcia sinusoidalnego. *Maszyny Elektryczne*, No. 1, p.183-188
[3] Zaskalicky P. 2014 Calculation of a torque ripple a three-phase asynchronous motor supplied by a PWM controlled inverter. *Maszyny Elektryczne*, No. 2, p.53-58
[4] Zawirski K., Deskur J., Kaczmarek T. 2012 Automatyka napędu elektrycznego. Wydawnictwo Politechniki Poznańskiej, Poznań[
[5] Matlab-simulink MATLAB software library
[6] Załęski J., Dadana M., Młodzikowski P., Twerd M. 2013 Przekształtnikowy napęd samochodu elektrycznego. *Maszyny Elektryczne*, No. 1, p.89-93
[7] Giurgiutiu V., Lyshevski 2009: Micromechatronics. Modeling, Analysis, and Design with MATLAB. CRC Press, by Taylor & Francis Group, LLC
[8] Mohan N. 2014 Advanced Electric Drives. Wiley&Sons