MODERN DIMENSIONING OF SWITCHED RELUCTANCE MACHINES

Abstract: In this paper a modern dimensioning procedure for Switched Reluctance Machines (SRMs) is presented. Self-prepared software makes it possible to find optimal set of construction parameters. Moreover first machine characteristics, losses and efficiency can be calculated. Goal is the optimization of the design by examination of special inner appearances. With the inclusion of the energy supply, converter, mechanical components and control strategy by a simulation model the system behaviour can be determined in advance.

1. INTRODUCTION

The various advantages of the SRM enables it an attractive alternative to existing dc and ac engines in the adjustable speed drives. It has a simple one construction, low rotor inertia, a high starting torque, high speed performance and low costs. The machine has salient poles on both stator and rotor. Concentrated coils are placed around the stator poles and there are no windings of any kind on the rotor. The SR Motor produces torque in consequence of variability of reluctance for magnetic flux excited by currents flowing in stator windings in accordance with the rotor position. The structure of a SRM is presented on Fig. 1 for example of a 3-phase machine with 6 stator and 4 rotor poles (6/4 machine). The main dimensions are the bore diameter \( D \), the rotor length \( L \), the stator outer diameter \( D_s \) and the stator- and rotor arc \( \beta_s \) and \( \beta_r \).

The angle \( \Theta \) describes the rotor position, the starting point \( \Theta=0 \) corresponds to the unaligned position, where the midpoint of the interpolar rotor gap faces the stator pole. In the aligned position the centre of the stator and rotor poles are coinciding. Magnetic cores of stator and rotor consist of laminated steel. Coils located on reciprocal poles are serial connected and creating a phase band. Each band of the stator winding is connected to supply by transistor switches.

The design of a SRM by numeric methods with FEM-programs (Finite Elements Method) provides the most precise and proof results. However, calculations by FEM are time consuming and require special software knowledge. For a complete design of new type series of machines respectively other dimensioning variants the exclusive way of using the FEM is at present not feasible. Therefore software is necessary that can abbreviate the preparation time for the following FEM-calculation considerable. Often the dynamic operational behaviour and the combination of the electric machine with other components like energy storage, converter or mechanic elements must be considered. In that case parameters calculated in the design have to be involved in a simulation model.
Following a new simulation model with SIMPLORER and a special analytical program for designing and computation SRMs of various constructions [6] shall be presented. This software is the basis of which a rapid design of the machine and the estimation of its parameters are possible without special software knowledge. The program is designed to be fast in operation, with an efficient user interface. Its main use is in
- a totally free new design or design of a similar machine with the same construction
- winding design
- calculation of the magnetic circuit, considering the nonlinear characteristics of the iron
- calculation of losses, including iron- and copper losses of the SRM and converter losses
- evaluation of SRM characteristics (torque, inductivity, flux linkage, ...)
- interface to a following FEM calculation and simulation

Starting from a performance requirement or after loading a machine dataset the computation of all results follows in seconds. A control window shows the program progress and guides the user through the design. The general calculating procedure shows Fig. 2:

![Fig. 2: Design process with the analytical design- and calculation program]

Parameters calculated by the design program can be involved in a simulation model to examine the dynamic operational behaviour and the combination of the SRM with other components like the power supply, converter and of course mechanic elements. The simulation is carried out by the program SIMPLORER. An own motor model was developed [6], consisting of a mechanical and an electrical part, including the converter connection.

2. DESIGN

In general a specification of an electrical machine consists of requirements (e.g. torque, speed) and constraints (e.g. dimensions or supply voltage). Moreover the design is based on the compliance of a number of prescriptions or agreements at the valid heating, mechanical stress, operational safety and the compliance of electrical limits. Target of the design is the determination of the main dimensions and the electric and geometric data of all electromagnetic claimed parts from the required properties. Note that the number of questions is very large because many of the constrains which simplify ac machines do not apply: the number of phases is open to choice between one and many, the ratio of phase number to stator pole number is not fixed and the ratio of rotor pole number to stator pole number is open to a wide variety of choices. The best values of rotor and stator pole arc have to be considered. Furthermore attention has to be paid to matters concerned with core losses, switching frequencies and varying flux distributions in different parts of the magnetic circuit. Of course
it is not possible to cover all these questions adequately in the context of this paper, but several are dealt with below. Attention is given to fundamental design considerations.

Sizing the SRM starts with the determination of the main dimensions by the output equation [3], which relates the bore diameter $D$, rotor length $L$, speed $n$, and magnetic and electric loadings to the output $P$:

$$ P = C \cdot D^2 \cdot L \cdot n $$  \hspace{1cm} (1)

$C$ is the output coefficient and essentially depends on the machine dimensions and the cooling system. To determine the main dimensions, it is necessary to keep the rotor length as a multiple or submultiples of bore diameter.

Thereafter the sizing of internal dimensions follows, starting with the airgap. To maintain balanced phase currents and minimize acoustic noise, the SRM needs a uniform airgap. It also requires a small airgap to maximize specific torque and minimize the volt-ampere requirement in the converter. However, the bending of the shaft and the expansion of the material with increasing temperature must be considered during design in addition to manufacturing tolerances, so the air gap should be chosen in such a way that the machine works reliable under default operating conditions in every operating point. A typical airgap length $\delta$ is usually in the range of

$$ 0.2 \leq \delta \leq 0.6 \text{mm} \ . $$ \hspace{1cm} (2)

The airgap value influences the maximum torque value and also the flat torque range on the characteristic, as Fig. 3 shows for the example of a 4-phase 16/12-SRM with $P=18.5\text{kW}$.

The next design step is the selection of the pole arcs, depending on two criteria:

1. Self-starting requirement
2. Shaping of static torque vs. rotor position characteristics

These requirements can be incorporated into the machine design by computing the minimum rotor and stator pole arcs to achieve self starting:

$$ \min(\beta_s, \beta_r) \geq \frac{2\pi}{q \cdot N_r} $$ \hspace{1cm} (3)

with the number of rotor poles $N_r$ and the phase number $q$. The minimum pole arcs are equal to the stroke angle, ensuring that in the ideal case with no fringing flux, torque can be produced at all rotor positions. An upper limit is placed on the overlap of stator and rotor teeth:

$$ \beta_s \leq \frac{2\pi}{N_r} - \beta_r $$ \hspace{1cm} (4)
This ensures that in the unaligned position there is a clearance between rotor and stator poles. In practice is \( \beta_r \geq \beta_s \) providing a slightly larger slot area without sacrificing aligned inductance. Once the pole arcs are derived analytically to satisfy the aforementioned requirements, the width of the stator and rotor teeth, \( t_s \) and \( t_r \), follows in accordance to Fig. 1 from the airgap \( \delta \) and the rotor diameter \( D_r \):

\[
t_s = 2 \cdot \left( \frac{D_r}{2} + \delta \right) \cdot \sin \left( \frac{\beta_s}{2} \right) \quad \text{and} \quad t_r = 2 \cdot \frac{D_r}{2} \cdot \sin \left( \frac{\beta_r}{2} \right) \tag{5, 6}
\]

To research the influence of constructional parameters on the electromagnetic torque with the design program, the values of the rotor and stator pole width were changed. During changing one of these parameters, the other one remains constant. Increasing the rotor pole arc causes only changing of maximum torque with rotor position and influences the width of the torque impulse. The maximum value of torque is practical constant in the examined range (see Fig. 4). There is an optimum value of rotor pole wide when the torque integral (area under characteristic) has its maximum. Increasing the stator pole arc leads to a smaller slot area and limits the maximum magnetomotive force (mmf) and therefore the maximum torque (Fig. 5). On the other hand it has to be considered from the mechanical aspect that a narrower stator pole can be easier stimulated to oscillate, what influences acoustic noise negatively.

![Fig. 4: Influence of rotor pole arc on torque](image)

![Fig. 5: Influence of stator pole arc on torque](image)

The optimum stator and rotor pole arc is a compromise between various conflicting requirements, and there is no single value that is appropriate for all applications. For very high efficiency designs the slot area needs to be maximized, leading to a narrow pole arc. But the starting capability may be compromised because of torque dips and extreme torque ripple. Wider pole arcs can avoid these problems, but the price paid is a smaller slot area and higher copper losses. The choices depend on the entire torque/speed range and on the number of poles and phases.

Shaping of torque vs. Rotor position characteristics is possible with different widths at the tip and base of the stator- and rotor poles (tapered poles). As a result, the bases of the poles are magnetically disburdened, leading to a higher flux linkage at a given mmf and a smoother
torque. The demand on the torque-smoothing controller and hence on the current controllers is minimized by this torque shaping. Further, for many low-performance applications, any additional torque smoothing other than the torque shaping provided by variation of pole arc base may not be necessary. A hybrid method of both pole shaping and torque control using modern techniques will enhance the SRM drive system performance to suit almost all high performance applications.

All of the described geometric dimensions are determined in the presented design program considering the required properties. Fig. 7 shows a dialog for the example of an 18.5kW SRM with 16 stator and 12 rotor poles.

3. CALCULATION

3.1 Magnetic circuit

The calculation of the magnetic circuit forms the focal point of the design. The first step is a rough estimating of the number of turns per phase. Assuming that at a specific speed the conduction angle of the power transistors has a certain value, equal to the stroke angle of the SRM, the peak flux linkage per phase peak at constant supply voltage \( u \) is

\[
\psi_{\text{peak}} = \frac{u \cdot \Delta}{\omega}. \tag{7}
\]

At full speed \( \psi_{\text{peak}} \) occurs before the aligned position. Assuming that the ampere-turns are sufficient at this point to bring the stator pole to the flux density \( B_{sp} \) across the entire width at its root, the peak flux linkage per phase results with the stack length \( L_{stk} \) in:

\[
\psi_{\text{peak}} = t_s \cdot L_{stk} \cdot B_{sp} \cdot N_{ph} \tag{8}
\]

Combining equations (7) and (8) the number of turns per phase \( N_{ph} \) is

\[
N_{ph} = \frac{u \cdot \Delta}{\omega \cdot L_{stk} \cdot t_s \cdot B_{sp}} \tag{9}
\]

According to Fig. 1 the magnetic circuit of the SRM can be subdivided into the stator pole, airgap, rotor and stator back iron and rotor pole. If the stator flux density is \( B_{sp} \) and the
area of cross section for the flux path is \( A_{sp} \), their product gives the stator pole flux \( \Phi_{sp} \). With the stator pole flux the flux densities for the other parts are likewise obtained. The magnetomotive force (mmf) \( F \) for each segment of the flux path is evaluated by finding the magnetic field densities \( H \) using the B-H-characteristics of the lamination material and average length \( l \) of the flux path in these segments. Then, the ampere circuital equation is

\[
F = N_{ph} \cdot i = \sum H \cdot l \quad ,
\]  

(10)

where \( i \) is the phase current. If the computed mmf is not equal to the applied mmf, the error between them can be reduced by an iterative adjustment of the stator pole flux density, and then recalculating all other variables. From the final flux densities in the various segments of the flux path and the magnetic field intensities, the reluctances \( R_m \) are calculated. The reluctance \( R_{m,sp} \) of one stator pole segment with the length \( l_{sp} = h_s \) is

\[
R_{m,sp} = \frac{l_{sp}}{A_{sp} \mu_0 H_r} = \frac{H_{sp} l_{sp}}{B_{sp} A_{sp}} \quad .
\]  

(11)

Similarly, the reluctances of the air gap, rotor and stator back iron and rotor pole are obtained.

The requirement for finding performance characteristics of the SRM is to generate the relationships between the flux linkages \( \psi \) vs. current \( i \) as a function of the rotor position \( \Theta \) (see Fig. 8). In the unaligned position the air gap and therefore the reluctance is maximal. Since no saturation occurs, the flux linkage is a linear function of the current in this position. During movement of the rotor tooth to an excited stator pole the airgap decreases. That leads to saturation of the magnetic circuit. Because of the high induction values in the tooth tips, these parts saturate at first and cause an enlargement of the effective airgap. As a result, the linear relationship is lost between current and flux linkage.

The energy \( W \) supplied by the inverter is \( W_F + W' \). Physically \( W_F \) corresponds to the stored field energy in the case of a specific rotor position and constant current in the machine. It is returned minus hysteresis losses to the supply after commutation. The coenergy \( W' \) that is available to be converted into mechanical work in each working stroke is equal to the area enclosed by the trajectory of the operating point in the \( \psi - i \) diagram. An "energy ratio" \( \lambda \) has been defined in [2] which tells how much energy conversion \( W' \) is obtained for a given input energy:

\[
\lambda = \frac{\Delta W'}{\Delta W_F + \Delta W'}
\]  

(12)

As Fig. 7 shows, the ratio of the coenergy \( \Delta W' \) increases for the field energy \( \Delta W_F \) due to increasing saturation. As a result the energy ratio increases. Furthermore it is shown in [4] that the effect of saturation is to reduce the energy conversion capability of a motor of given dimensions, but at the same time it reduces the inverter volt-ampere-requirement for a given torque and speed by a greater factor. The degree of saturation will influence the balance
between motor and inverter size. Saturation is desirable from this point of view, but must be localized on the tooth tips, for instance by a tapered shape of the teeth, since the electromagnetic energy conversion occurs in the air gap. Consequently, the demand to operate the SRM partially in the range of saturation is not in the contradiction to the generally usual procedure to avoid the non-linear operation at electrical machines.

Within the design program a procedure for analytically deriving the machine characteristics given the motor dimensions and excitation conditions is realized. Figure 9 and 10 show the calculated $\psi$-i diagram and the inductivity as a function of phase current for the example of an 18.5kW SRM. Because of the influence of saturation, inductance strongly decreases in the aligned position with increasing current. In the unaligned position the inductance is nearly independent of the current due to the larger airgap.

3.2 Coenergy and electromagnetic torque

The next step of the performance computation is determining the electromagnetic torque $T$. After calculating a discrete data set of flux linkage vs. current vs. rotor position angle, the extraction of flux linkage for any current has been developed and the electromagnetic torque can be obtained from the change in coenergy $W'$:

$$T = \left[ \frac{\partial W'(i, \Theta)}{\partial \Theta} \right]_{i=\text{const.}}$$
with  $$W'(i, \Theta) = \int_{i_1}^{i_2} \psi(i, \Theta) di$$  (13, 14)

In the presented program the flux linkages are integrated with respect to the phase current for every rotor position. The area under each $\psi$-i-characteristic at constant rotor position, corresponding to the coenergy, is calculated by a rectangle approximation. The result is a discrete data set of coenergy vs. rotor position angle as a function of the current (Fig. 11). If the inductance is nearly varying with rotor position for a given current, which is in general not the case in practice, then the torque can be derived as:

![Fig. 9: Calculated $\psi$-i-diagramm](image)

![Fig. 10: Calculated L-$\Theta$-diagramm](image)

![Fig. 11: Calculated coenergy vs. rotorposition](image)
\[ T = \frac{dL(i, \Theta)}{d\Theta} \cdot \frac{i^2}{2} \]  

Figure 12 shows the calculated torque vs. rotor position characteristic as a function of phase current under real saturation conditions (a) and for ideal case with linear magnetization curve (b). Neglecting saturation effects, torque is proportional to the square of the current; hence the current can be unipolar to produce unidirectional torque. Such feature greatly reduces the number of power switches in the converter and thereby makes the drive economical.

3.3 Iron losses

Losses in SRMs consist mainly of iron losses and stator copper losses. The copper losses are proportional to the square of the rms current whereas the iron losses are a function of the excitation frequency and flux density. Estimating the iron losses is complicated in comparison to conventional a.c. variable speed machines by the fact that the frequency of flux reversals is different for different parts of the core. Figures 13 and 14 show the stator and rotor flux density waveform for a half rotor revolution as a function of the rotor position angle.

Fig. 13: Flux density waveform in the stator of a 16/12-SRM
The complexity increases with the number of phases. Further, the current waveform is not sinusoidal and is depended on operating conditions. The fact that SRMs mostly operate in varying degrees of saturation also complicates the loss calculation. An approximate but simple procedure has been realized in the presented design program for determining the flux density waveforms for any pole combination. The specific iron losses \( p_c \) corresponding to each segment are then determined by a new calculation method described in [5]. This procedure subdivides the actual flux density waveform into linear sections. The specific iron losses are calculated using a modified Steinmetz’s equation for each part of the core. In order to include the magnetic reversal rate the equation is extended with a frequency equivalent \( f_{eq} \) and multiplied for regard of repetition of the sequence with the frequency \( f_T \):

\[
p_c = f_T \cdot \left( C_m \cdot f_{eq}^{(a-1)} \cdot B^\beta \right)
\]

The coefficients \( C_m, \alpha \) and \( \beta \) are empirically extracted from loss curves of the iron. The frequency equivalent is derived by subdividing the actual induction waveform into linear sections. The magnetic reversal rate \( B = dB/dt \) of these parts is weighted with maximum modulation \( \Delta B = B_{max} - B_{min} \). From a comparison of the sum of all partial sections with the weighted rate of change of a sine-wave oscillation with the same modulation \( f_{eq} \) becomes

\[
f_{eq} = \frac{2}{\pi^2} \cdot \frac{1}{(B_{max} - B_{min})^2} \cdot \int_0^{T} \left( \frac{dB}{dt} \right)^2 dt
\]

For piece wise linear induction sections, as they are available at SRMs, \( f_{eq} \) can be simplified:

\[
f_{eq} = \frac{2}{\pi^2} \cdot \sum_{k=2}^{K} \left( \frac{B_k - B_{k-1}}{B_{max} - B_{min}} \right)^2 \cdot \frac{1}{t_k - t_{k-1}}
\]

With reference to Fig. 13a the specific iron losses \( p_{c,SP} \) of one stator pole can be determined:

\[
p_{c,SP} = \frac{N_r}{T} \cdot C_m \cdot \left( \frac{2}{\pi^2} \cdot \left[ \frac{1}{\Delta t_r} + \frac{1}{\Delta t_f} \right] \right)^{(a-1)} \cdot \left( \frac{B}{2} \right)^\beta,
\]

where \( t_r \) and \( t_f \) are the rice and fall time of the stator pole flux density, \( T \) is the time for one
rotor revolution. Similarly the specific iron losses in the rotor pole are calculated. The estimation of core losses in the stator- and rotor yoke is more complicated since the induction waveforms in these segments are a sum of stator- and rotor pole induction waveforms. They can be split into two basic figures whose frequency equivalent can be obtained with (18).

\[ \Delta t_0 \]

Now the specific iron losses in the other parts of the magnetic circuit can be calculated. For the example of the stator yoke segments with bipolar flux waveform \( p_{cSY} \) becomes

\[
p_{cSY} = \frac{N_r}{T} \cdot C_m \cdot \left[ s \cdot \left( \frac{2}{\pi^2} \cdot \left[ \frac{1}{\Delta t_1} + \frac{1}{\Delta t_2} \right] \right)^{(a-1)} \cdot \left( \frac{B_{sub}}{2} \right)^\beta + \right. \]

\[
\left. + \left( \frac{2}{\pi^2} \cdot \left( \frac{1}{2 \cdot B_0} \right)^2 \cdot \left[ 2 \cdot \frac{B_{sub}^2}{\Delta t_1} + 2 \cdot \frac{B_{sub}^2}{\Delta t_2} + 2 \cdot \frac{(B_0 - B_{sub})^2}{\Delta t_0} \right] \right)^{(a-1)} \cdot B_0^\beta \right] \quad (20)
\]

The parameter \( s \) describes the number of the small figures according to Fig. 15. A value \( s=1 \) is valid for a three-phase machine, \( s=2 \) in the case of a 4-phase machine. The flux in the rotor yoke is bipolar, hence the specific iron losses can be calculated similarly.

Once the specific iron losses \( p_{ck} \) and the iron weights \( m_k \) of each part \( k \) of the magnetic circuit are known, the total iron losses \( P_c \) are calculated according to

\[
P_c = \sum p_{ck} \cdot m_k \quad (21)
\]

To verify the computation algorithm, the calculated iron losses are compared with measuring results of a 4-phase 18.5kW SRM as a function of speed, torque is constant. The results match very well, but note that the calculation procedure may be used for an approximate estimation but not for applications where precision in predicting efficiency or thermal rise is required. Then, finite element analysis has to be resorted too.

Fig. 15: Splitting the stator- and rotor yoke flux density waveform

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\[
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\[
\left. + \left( \frac{2}{\pi^2} \cdot \left( \frac{1}{2 \cdot B_0} \right)^2 \cdot \left[ 2 \cdot \frac{B_{sub}^2}{\Delta t_1} + 2 \cdot \frac{B_{sub}^2}{\Delta t_2} + 2 \cdot \frac{(B_0 - B_{sub})^2}{\Delta t_0} \right] \right)^{(a-1)} \cdot B_0^\beta \right] \quad (20)
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Fig. 16: Iron losses of a 18.5kW SRM as a function of speed
4. SIMULATION

In order to investigate the operational behaviour as well as special operating states of the entire drive system, a simulation model of the SRM has been developed with the program SIMPLORER. The simulation model contains the inverter with power supply, a current regulation and the motor model, consisting of an electrical and a mechanical part (Fig. 17). The electrical part represents the voltage equation assuming that there is no mutual coupling to other phases:

\[ u = R_i \frac{\partial \psi(i, \Theta)}{\partial i} \frac{di}{dt} + \frac{\partial \psi(i, \Theta)}{\partial \Theta} \frac{d\Theta}{dt}, \]  

where \( u \) is the supply voltage and \( R \) is the winding resistance. The dependence of the inductance \( L \) of the current and the rotor position angle can be considered by including the \( L(i, \Theta) \) characteristic from a preceding analytical computation or a FEM calculation. The mechanical part of the motor model corresponds to the mechanical equations with the load torque \( T_L \), the friction torque \( T_{Fr} \), the moment of inertia \( J \), the number of rotor poles \( N_r \) and the angular frequency \( \omega \):

\[ T - T_L - T_{Fr} - \frac{J}{N_r} \frac{d\omega}{dt} = 0, \quad \omega = \frac{d\Theta}{dt} \]  

The inverter model is an asymmetric bridge converter. It offers the maximum flexibility during the current regulation, caused by a tolerance band procedure. Fig. 18 shows the phase currents and the average torque of a 3-phase motor as a function of time. The turn-on and turn-off angle of current can be adjusted. With it two essential parameters are available for torque control. The 3rd control parameter, the width of the current band, has influence on the inverter losses as well as the vibrations and the resulting acoustic noise in the SRM. The simulation of various control schemes is possible. Note that the simulation can not consider the real saturation conditions, but it gives acceptable results multiple faster than the FEM.
5. CONCLUSION

A modern method for designing and determining Switched Reluctance Machines has been presented. The machine design consists of the specification of all dimensions and the stepwise optimization of geometry, winding and material. There are embedded special demands like the optimization of torque characteristics, efficiency or acoustic noise. Some fundamental guidelines for an optimized machine design can be derived from the results:

- The airgap should have the lowest value, which is accepted due to mechanical reasons. The airgap length influences hardly the maximum torque value as well as the volt ampere requirement of the SRM, provided by the converter.
- Width of pole not influences the maximum torque, but there exists an optimum value, which causes maximum energy conversion during phase band operating cycle.
- Torque smoothing by variation of pole arc base can save additional electronic torque control techniques for many low-performance applications.
- A combination of pole shaping and torque control using modern techniques will enhance the SRM drive system performance to suit almost all high performance applications.
- Saturation reduces the inverter volt-ampere-requirement for a given torque and speed and influences the balance between motor and inverter size. Saturation is desirable from this point of view, but must be localized on the tooth tips, for instance by a tapered shape of the teeth.
- The iron losses are a function of flux density excitation frequency. Reducing the iron losses at high speed is possible by decreasing the number of stator phases. Three phases is preferred over four phases for high speed applications.

A number of experience values can be effectively used in the design of SRMs, as they could use the commonality between these and the conventional machines to start with. The resulting machine dimensions form the starting point in design evaluation, and final design is achieved through an iterative process of steady-state performance calculations.

6. SUMMARY

The design of new electric drives requires modern design tools. The presented procedures - analytical design by a design program and simulation of the whole drive system - allow the examination of essential machine parameters as basis for the design of an optimized SRM. They are prepared for low- and high speed applications. The validity of the procedures is evaluated by comparing with measurement results obtained from an 18.5kW SRM. The analytical results are close to the measurements.

7. REFERENCES