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Design, Analysis and Optimization of Magnetic Circuits for Linear Dynamic Actuators

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***Abstract* – Contactless braking methods (with capability of energy recuperation) are more and more widely used and they replace the traditional abrasive and dissipative braking techniques. In case of rotating motion, the method is trivial and often used nowadays. But when the movement is linear and fast alternating, there are only a few possibilities to break the movement. The basic goal of research project is to develop a linear braking method based on the magnetic principle, which enables the efficient and highly controllable braking of alternating movements. Frequency of the alternating movement can be in wide range, aim of the research to develop contactless braking method for vibrating movement for as higher as possible frequency. The research includes examination and further development of possible magnetic implementations and existing methods, so that an efficient construction suitable for the effective linear movement control can be created. The first problem to be solved is design a well-constructed magnetic circuit with high air gap induction, which provides effective and good dynamic parameters for the braking devices. The present paper summarizes the magnetostatics design of "voice-coil linear actuator" type actuators and the effects of structure-related flux scattering and its compensation.**

Keywords – magnetic brake, contactless brake, lifetime test, magnetic circuit analysis

I. INTRODUCTION

In the field of manufacturing, many methods of lifetime testing are used. In case of electronic devices, the stress- and lifetime testing is quite simple in most cases compared to the mechanical tests, when mechanical load emulation is very difficult and expensive. For example, power tools are tested with different loads and some of them have alternating linear movements which should be loaded. There are not fully developed contactless load emulating methods for this application. Sometimes hydraulic system is used with low efficiency. In case of traditional practice test methods, an operator works with the device under test (DUT), he/she cuts, sands, or planes different materials. This type of testing is expensive, not reliable, and less repeatable than the automated test solutions. In some tests the operators were replaced by industrial robots, but the test is still expensive and dirty due to the using of real materials. The best result would be a system which can emulate the loading force without any physical contact, abrasion, and dirt.

Research of braking method for fast alternating linear movements by using contactless magnetic braking methods is in the focus of our project. Project includes analysis of voice-coil type magnetic actuators, design of magnetic circuit to maximize efficiency and reliability and minimize weight and size of the brake. The first part of this paper includes introduction of a method for transformation of 2-dimension magnetic calculations to the cylindrical coordinate system and presents the analysis of the flux leakage and its effects to the results of calculations. The calculations are confirmed by finite element simulations, and the results are also used to correct the differences caused by the flux leakage. These calculations, simulations and corrections were solved for different shapes to realize a shape- and size independent model to calculate the correct average flux density in the air gap. The second part of paper presents the results of dynamic simulations, by which dynamic behaviour (relationship between the current and the force in dynamic cases, eddy current- and solid losses, etc.) of the voice coil-type actuator is analysed. [1] [3]

II. DESIGN OF CYLINDRICAL MAGNETIC CIRCUIT WITH TWO-DIMENSIONAL, PLANE CROSS-SECTION MODEL CALCULATIONS

The aim of the first part of the work was to theoretically establish, develop, and validate a method that transforms the dimensions of a cylindrical symmetrical magnetic circuit of a given size into an equivalent two-dimensional cross-sectional and constant depth model. In this way, cylindrical magnetic circuits can also be calculated. In this method the cylindrical magnetic circuit is "spread out" so that the vertical (h) dimensions are leaved unchanged, and the r values are transformed into x values to create a flat-section, fixed-depth model in which the volume of each part is the same, so the two models is connected by the unchanged value of the flux. However, inductions calculated in the planar model is also valid for the transformation, the calculated values correspond to the average values in the cylindrical model, since values of the magnetic induction in the cylindrical model are changing in radial direction. As a result, higher induction values are observed on the inner half of the cylindrical parts and lower on the outer half. Figure 1. shows the dimensions of a typical cylindrical model, and Figure 2 illustrates the corresponding x and y and depth (d) dimensions in a planar cross-sectional model for transformation. [6] [7]

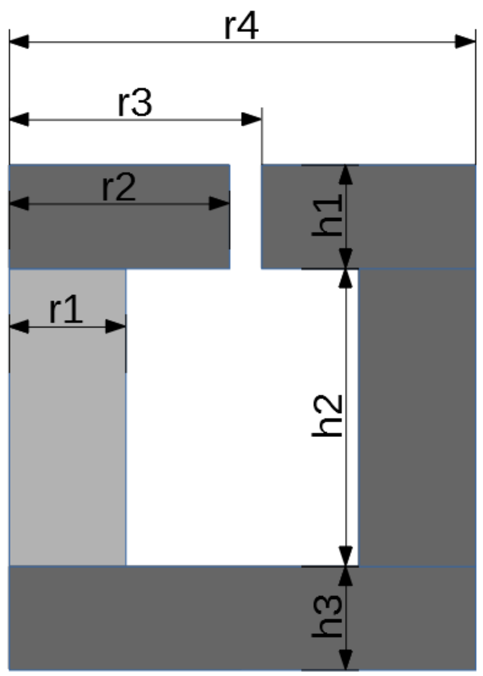

Fig. 1. The half cross-section of a typical cylindrical magnetic circuit

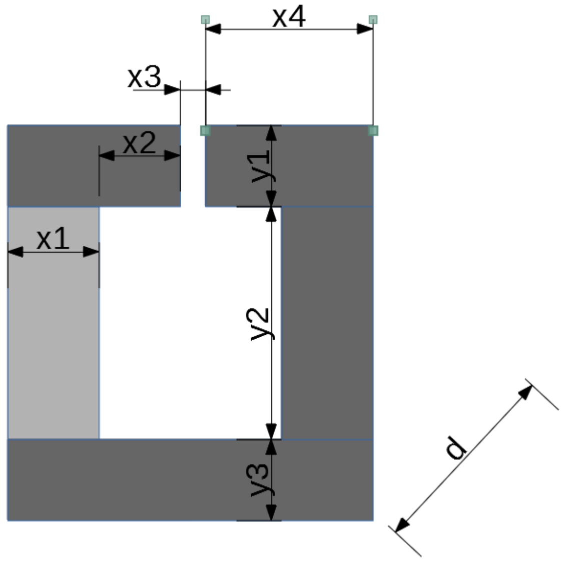

Fig. 2. The cross-section of the 2-D model

The depth (d) of the plane cross-section fixed-depth model should be taken so that x dimensions of the plane model are close to the radius differences of the cylindrical model in the area most affected by the test (this is practically the air gap). Thus, for practical reasons, depth dimension d is selected equal to the length of line at the air gap center circle which is as follows:

$$d = (r_2 + r_3)\pi$$

To determine the relationship between the radius and the x values, the volume equality was used as already mentioned above, which is the following:

$$V_{n_{rh}} = V_{n_{xy}}$$
$$({r_n}^2 - {r_{n-1}}^2)\pi h_m = x_n y_m d$$

Since the given h and y values in the two models are the same based on a previous condition, the final correlation for the x values after transformations is as follows:

$$x_n = \frac{({r_n}^2 - {r_{n-1}}^2)\pi}{d}$$

In the above relation if $n = 1$, then r_0 is 0, assuming that inner bar (iron cores and also magnets) is cylindrical. If the inner bar has ring section, value of r_0 is equal to radius of inner ring.

III. VERIFICATION OF THE RELATIONSHIPS RECEIVED BY FINITE SIMULATION

To verify transformation relationships introduced above, a transformation of a cylindrical magnetic circuit of a given size was performed. In determining the depth dimension d, the length of the center circle of the air gap was considered, which will result nearly equal air gap induction in plane model and in the cylindrical model. The dimensions of the initial cylindrical model and the plane model which is the result of the transformation are summarized in Table 1.

Parameter	Value (mm)
r1	40
r2	70
r3	72,5
r4	130
r5	150
x1	11,23
x2	23,16
x3	2,49998 (~2,5mm)
x4	81,71
x5	39,3
h1 = y1	20
h2 = y2	60
h3 = y3	20
d	447,68

Table 1. Basic dimensions of the cylindrical model and calculated dimensions of the transformed model

The next step was to determine the air gap induction based on static magnetic calculations in which dimensions of the plane model were used. Initial data had to be determined, these are the main magnetic parameters of the applied soft ferromagnetic and permanent magnetic material. These data were received from the finite element simulation software database for better comparison. These material properties are as follows (the setpoint values of the magnet have been determined graphically):

- Permanent magnet
  - Material: Y25 ferrite magnet
  - Residual induction: 0,378 T
  - Coercive force: 153035 A/m
  - Relative permeability at the setpoint section of the demagnetization curve: 1,9
- Soft steel parts
  - Material: 1010 low carbon soft steel
  - Relative permeability: ~10^3
  - Maximum induction: 1T (in linear part of B-H graph)

The summarized relative permeability of the permanent magnet and the air gap is three order of magnitude smaller than relative permeability of the body, therefore the permeability of the soft iron was neglected in the calculations. In this phase of the research the magnetic calculations focused to the permanent magnet and the air gap. Based on the above described initial data, the main steps of the calculations were as follows.

1.) Determining the setpoint of the magnet by equation of the air gap line (based on data of the transformed geometry):

$$H_m = -\frac{1}{\mu_0} \cdot \frac{x_1}{y_1} \cdot \frac{x_3}{y_2} \cdot B_m$$
$$H_m = -7000\,\frac{A}{m}$$
$$B_m = 0{,}3726\ T$$

2.) Determination of the flux:

$$\Phi = B \cdot A = B_m \cdot A_m = B_m \cdot x_1 \cdot d = 1{,}873 \cdot 10^{-3} Wb$$

3.) Determination of the air gap induction:

$$B = \frac{\Phi}{A} \rightarrow B_\delta = \frac{\Phi}{A_\delta} = \frac{\Phi}{y_1 d} = 0{,}2092T$$

After performing the calculations, both the original cylindrical and the transformed magnetic circuits were simulated by FEMM 4.0 finite element simulation software. The simulation results and the calculated values are summarized in Table 2.

Parameter	Calculated	Simulated (cylindrical)	Simulated (plane)
B_m	0,3726T	0,3739T	0,3744T
Φ_m	$1{,}873 \cdot 10^{-3}$ Wb	$1{,}879 \cdot 10^{-3}$ Wb	$1{,}886 \cdot 10^{-3}$ Wb
B_δ	0,2092T	0,1509T	0,138T
Φ_δ	$1{,}873 \cdot 10^{-3}$ Wb	$1{,}351 \cdot 10^{-3}$ Wb	$1{,}24 \cdot 10^{-3}$ Wb

Table 2. Comparison of the calculated and the simulated results

The air gap induction value was approx. 75% compared to the calculated one in the simulation (the reason is the leakage flux around the air gap). In spite of this, the average value of the induction in the magnet and the flux of the magnetic circuit showed only a very small difference from the calculated data. According to these results we can state that the model and transformation provided good results and can be used in the next stage of the research. Most of the differences are due to the leakage flux, correction of which needs further studies. Although flux in the air gap is also 72% less, including leakage induction lines, simulation gives the original calculated flux value.

IV. CALCULATION OF MAGNETIC CIRCLE FOR REQUIRED AIR GAP INDUCTION AND AIR GAP DEPTH

While the previous calculations illustrated the transformation of a magnetic circuit with given dimensions, in practice, developing a so-called “voice-coil-actuator”, a much more common problem is adjusting dimensions of the structure, especially dimensions of the magnet to be used, to the given air gap height and air gap induction value. In such a case, the minimum air gap diameter has to be defined at which the given induction can be performed at the specified air gap height.

Also, if the diameter of the air gap is fixed, feasibility of the desired induction with the given permanent magnet type at the specified sizes should be checked. In addition, the effect of leakage flux must be considered when calculating these data (the study and correction of leakage is discussed in chapter VI.

Calculation steps:

1.) Definition of setpoint values of magnets

The real demagnetization curve of permanent magnets is linear in a relatively wide range, it has nonlinearity only near the coercive force. Therefore, setpoint of the magnet should be defined to provide maximum value of the product B x H, which is in the point $B_m = B_r / 2$ in the practice.

2.) Determination of the cross section of magnets perpendicular to flux

The air gap flux can be determined from the air gap cross section and the desired induction. Since the air gap and the flux of the magnet are the same in theory, the cross section of the magnet can be determined from the equation $\Phi = B * A$. The value of the radius of this surface must be checked to ensure that it is smaller than the inner circle line of the air gap (in the case of a plane model the test can also be done, in which case the x value at the beginning of the air gap must be greater than or equal to x).

If the evaluation shows that the desired induction is not feasible at a given air gap circle, the air gap diameter must be chosen to be higher, or if it is not possible, the magnet can be used with a different (higher) induction than the optimal setpoint, which will result an increase in the length of the magnet (see point 3).

In practice, the air gap and the flux of the magnet do not match due to scattering, so the correction described later should be applied.

3.) Determination of the optimal length of permanent magnets

Since the reluctance of the iron body is neglected according to an earlier condition, the equation $\oint H\, dl$ in the magnetic circuit can be defined as follows (without considering leakage correction):

$$H_\delta \delta + H_m y_m = 0$$
$$\frac{B_m}{\mu_0 \mu_m} y_m = -\frac{B_\delta}{\mu_0} \delta$$

where μ_m is the relative permeability of the permanent magnet (typically will be between 1 and 2), B_m is the setpoint induction of the permanent magnet, y_m is the length of permanent magnet to be defined, B_δ required induction, δ length of air gap.

Calculation example shows that the required air gap induction can be achieved without operating the magnet at the setpoint. The demagnetizing field strength is less than the setpoint field strength, but in this case the magnet length must be greater than the optimal value for the equation to be realized.

If the goal is to use a permanent magnet with the smallest possible volume (and at the same time the lowest cost), it is advisable to use the optimal dimensions determined by the setpoint. To achieve this, custom-made permanent magnets are necessary in practice. Experience show that when using commercial permanent magnets from catalogs some we should accept some comptonization.

4.) Determination of cross section of soft iron body

Based on the data sheets of various commonly used soft iron materials, we can state that approx. up to 1T, their B-H curve is linear, so it is not recommended to design above this value. Otherwise, especially near saturation, the relative permeability of the iron decreases, and in this case the reluctance of the given section is no longer negligible. [6]

The previously determined flux value and the maximum induction can be calculated based on the cross-sectional dimensions.

V. APPLICATION OF FERRIT MAGNETS IN THE OUTER RING AS A FLUX CONDUCTOR

In the earlier stage of the research, we have performed dynamic simulations on "voice-coil actuator" constructions of different designs. Results of these simulations proved, that the reluctance of the magnetic circuit and the properties of the materials in the vertical columns and rings greatly influence the value of the inductivity of the moving coil and the magnitude of the eddy current losses during dynamic operation. We have also examined an initial experimental design that included a permanent magnet both inside and outside. A static study of this construction was also carried out, during which it was found that the external permanent magnet does not substantially increase the air gap induction in the magnetic-air gap-magnet series magnetic circuit, it only increases the demagnetization field strength of the inner, so-called "working" magnet, which results setpoint down shifting of this magnet.

However, dynamic studies have discovered some advantages, which are the reduced inductance of the moving coil and reduced power dissipation of iron loss. Results show that the inductance is approx. half of the level when using soft iron instead of an external magnet in such a structure. The conclusion is that if the dimensions of the external magnet are determined so that the magnetic field strength inside of it is close to 0, then the magnetic ring behaves as a "flux conductor" like iron with low relative permeability.

This operating state is also characteristic of soft iron materials in the near-saturation state, with significant flux leakage. However, in the case of permanent magnets, the leakage is minimal. The results are better dynamic parameters caused by reduced inductance of the moving coil as well as reduced eddy current losses, but in turn it and does not reduce the performance of the correctly calculated dimensions working magnet.

If this solution is used for the sizing of the magnetic circuit, the last point of the design steps is modified as follows: the cross section of the external magnetic ring at which the induction is equal to the value of the residual induction of the magnet must be determined. For practical reasons, it is recommended to choose the length of the external magnet as equal to the length of the inner magnet (this simplifies the construction).

While experience show that neodymium iron-boron (or samarium cobalt at higher operating temperatures) is the most suitable material for the internal working magnet due to its high energy density, conventional strontium ferrite magnets can also be used for the external magnetic ring used as a flux conductor. They have lower prices than the two types of magnets listed above and, since they are used as external elements, their relatively large size is not limited by critical parameters affecting the moving mass, such as the diameter of the moving coil.

VI. ANALYSIS AND CORRECTION OF FLUX LEAKAGE

The leakage of magnetic induction lines in a magnetic circuit with an air gap is a complex problem that depends on several design and operating parameters. Practical experiences show that in optimal situation more than 90% of the leakage flux is present around the air gap, but in case of permanent magnets operating out of the optimal setpoint or at soft iron sections near the saturation, significant part of lines may close outside the magnetic circuit. There are several estimation work-help documents of leakage calculations, including air-gap leakage calculations are available from permanent magnet manufacturers to magnetic circuit designers. In these documents, the air gap and the field around it are divided into several areas depending on the type of magnetic circuit, which can include semicylindrical, semi-spherical, quarter-spherical and/or prismatic areas. The magnetic reluctances of these parallel fields are defined using exact, empirical formulas. However, these definitions help only in the design of certain magnetic circuits with frequently used structures.

In general cases, the effects of leakage in an arbitrary magnetic circuit can be estimated most accurately by finite element simulation. This requires exact and accurate information about geometric model and characteristics of the applied materials.

In order to make method outlined earlier applicable in practice for design "voice-coil actuator" type electromagnetic actuators, correct levels of leakage flux should be estimated. In the first step we have worked out an algorithm for automated static magnetic calculations. The main input parameters of this algorithm are the residual induction of the applied magnet types (internal and external), its coercive force, the height of the air gap, and the radius of its internal and external sections. The required air gap induction can be given by the following values:

- start value
- stop value
- number of steps

The algorithm is able to generate table including the most important geometric parameters of the parametric simulation sequences, which are the following:

- r_0: the inner radius of the inner magnet
- r_1: the outer radius of the inner magnet, equal to the inner radius of the air gap
- r_2: the inner radius of the outer magnet, equal to the outer radius of the air gap
- r_3: the outer radius of the outer magnet

In the first simulation a magnetic circuit according to Figure 1. was used, in which the radius of the central circle of the air gap is 70 mm and the height of the air gap is 20 mm. The start value of induction is 0.1T and the stop value is 1T, with 0.05T steps. The 1T stop value is maximized by the given air gap center circle radius. Using the received geometric dimensions, a parametric simulation was prepared to investigate the difference between the average induction value experienced in the air gap and the initial air gap induction value due to flux leakage. In the simulation deviation of working range from the setpoint was examined for internal magnets (induction value is 0.67T for N35 type magnets) and deviation of the average induction from the residual induction value (Br = 0.38T) was analyzed. The cross-sections of the soft iron elements of the construction were set large enough to avoid saturation or close to saturation operation. Initial values and the results of the simulation are summarized in Table 3.

Required air-gap induction [T]	**Simulated air-gap induction [T]**	**Ratio of simulated and the required induction [%]**	**Setpoint induction of inner magnet [T]**	**Setpoint induction of external magnet [T]**
0,1	0,071	71,00	0,793	0,335
0,15	0,109	72,66	0,784	0,344
0,2	0,147	73,50	0,779	0,351
0,25	0,185	74,00	0,774	0,356
0,3	0,223	74,33	0,770	0,362
0,35	0,261	74,57	0,767	0,367
0,4	0,299	74,75	0,764	0,371
0,45	0,338	75,11	0,761	0,375
0,5	0,376	75,20	0,758	0,379
0,55	0,415	75,45	0,755	0,382
0,6	0,453	75,50	0,752	0,386
0,65	0,492	75,69	0,750	0,389
0,7	0,532	76,00	0,746	0,393
0,75	0,571	76,13	0,743	0,395
0,8	0,611	76,37	0,740	0,399
0,85	0,651	76,59	0,736	0,402
0,9	0,692	76,88	0,732	0,405
0,95	0,734	77,26	0,727	0,409
1	0,778	77,80	0,717	0,413

Simulation results prove slight increase (between 72% and 78%) in the ratio of theoretical and simulated air gap induction when varying the value of the air gap induction from 0.1T to 1T.

Examining the simulation results, we can find that that the setpoint of the working magnet on the demagnetization curve B-H always shifts to the right of the ideal setpoint. This phenomenon is caused by the modelling error, that is the computational models do not consider either the alternative reluctances caused by leakage or the real magnetization curve of the body. However, the difference in practice is small enough to be neglected in the general case. For this reason, the setpoint of the external magnets also differs slightly from H = 0.

The next step of the work is to determine a correction factor for the required initial air gap induction, resulting corrected geometric parameters at which the value of the simulated air gap induction will be equal to the originally required (not corrected) air gap induction. The correction relationship determined from the simulation results is expressed by the following equation. This equation is the first order polynomial form which is given by the previous simulation results.

$$B_{\delta_{korr}} = B_{\delta} \cdot 1{,}281 + 0{,}018$$

Including this correction into the original algorithm, the parametric simulation was repeated with initial values of 0.1T and 0.75T. The higher values is selected according to the maximum possible 1T corrected initial value of induction. The results are shown in Table 4.

Required induction [T]	**Corrected value of induction for simulation input[T]**	**Simulated induction [T]**
0,1	0,139	0,106
0,15	0,205	0,154
0,2	0,271	0,202
0,25	0,337	0,251
0,3	0,402	0,3
0,35	0,468	0,35
0,4	0,533	0,399
0,45	0,598	0,449
0,5	0,663	0,499
0,55	0,728	0,55
0,6	0,792	0,6
0,65	0,857	0,652
0,7	0,919	0,705
0,75	0,984	0,759

VII. CONCLUSIONS AND OUTLOOK

Results of the research show that air gap induction correction is an effective method to calculate geometry of magnets and checking the real air gap induction for the calculated geometry for magnetic circuits of so-called "speaker-type voice coil actuator" actuators. Accuracy of air gap induction simulation can be increased if considering further construction details, like join deviations or detailed material properties. Difference between the magnetic properties of real soft magnetic material and simulated material may also cause some simulation error. Converting cylindrical, axially symmetrical magnetic constructions to plane model by defined geometric transformations, the induction and field strength values of the magnetic circuit's sections can be determined with acceptable accuracy. The acceptable accuracy highly depends on the compensation capacity of the control system to be used in the system, therefore checking and correcting the calculations by finite element simulations can be still useful.

Validation of the developed calculation and simulation methods is in progress. A prototype is designed and built with the geometrical dimensions defined by the described methods, tests will be performed in the near future. The validated methods will be used for development and optimization of industrial testing processes.

VIII. ACKNOWLEDGEMENTS

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