“Looking Through Walls” – Actuator Position Measurement Through a Conductive Wall

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In many industrial applications of actuators a contactless position sensing of an object is of main importance. In the case of a conductive object, a popular choice are Eddy Current Sensors (ECSs), which due to their non-contact nature offer a clear advantage over e.g. resistive or capacitive sensors [1]. Moreover, ECSs can also operate under harsh conditions, in dirty environments- or in vacuum, which makes them suitable for many different applications for the stationary, quasi-stationary [2] or dynamic position measurement [3] in the mm- to mm-range, e.g. in Magnetic Bearings (MBs) of actuator systems employing magnetic levitation [4]–[6], where the ECSs located on the stator are used to capture the position of the levitated rotor or mover (cf. Fig. 1 (a)). Besides position sensing, ECSs are also extensively used in non-destructive testing for inspection of irregularities or damages on the surface of conductive materials [7]. Typically, ECSs are either realized as a single- or dual coil configuration, where either the distance-dependent variation of the excitation coil impedance or the distance-dependent induced voltage in a second pick-up coil is measured. The design of such single- and dual-coil ECSs is thoroughly discussed in the literature [8]–[11]. In contrast, there are also applications where the moving conductive target is located behind another conductive material. For example, in entirely sealed actuator systems used for high purity food, medical or chemical applications, the stator and the magnetically levitated mover are fully encapsulated e.g. with a stainless steel housing (cf. Fig. 1 (b)). Consequently, the ECS located on the stator has to measure the mover’s position through the stainless steel cover of the stator, which due to its conductivity highly degrades the magnetic coupling between excitation and/or pick-up coil and the mover, and in turn leads to a reduction of the measurement sensitivity, i.e. mover position accuracy.

In the literature, such an ECS measuring through a conductive wall is e.g. described in [12], where a commercially available eddy current flaw detector is used for periodical inspections of nuclear power plants by measuring the distance between two conductive tubes. There, the well established Dodd and Deeds analytic models [13], [14] are validated with measurements, and sensitivity results of the sensor output are given for different excitation frequencies up to 16 kHz. For stationary applications, such low excitation frequencies are clearly sufficient, however, for an ECS designed for the dynamic position measurement in entirely sealed MBs this results in a too low measurement bandwidth. In order to achieve a position control bandwidth of around 1-2 kHz, sensor measurement bandwidths of at least 10-20 kHz are needed [4], [6], which again roughly lead to 10 times higher excitation frequencies of above 100-200 kHz. However, for ECS looking through a conductive wall, it also has to be considered that above a certain excitation frequency - defined by the material properties and dimensions of the intermediate layer - the skin and proximity effects of this layer start to play an important role and in the worst case can lead to a complete loss of any mover position information. Consequently, the ECS’s excitation frequency has to be selected properly, such that on the one hand a sufficient position measurement bandwidth is achieved, and on the other hand the position sensitivity is not too much affected by the intermediate conductive layer.

Therefore, this paper aims to provide a general understanding and design guidelines of the ECS measuring through a conductive layer, i.e. how the applicable excitation frequency range is influenced by the material properties and dimensions of both the intermediate layer and the mover, and how the frequency is optimally selected to achieve the highest position sensitivity. The presented FEM simulation based analysis is conducted for an ECS used as position sensor of the aforementioned entirely sealed linear-rotary actuator systems with magnetically levitated mover, whose specifications are given in Tab. I. Even though for this application the stator and mover materials are already defined, the influence of other sealing materials is investigated. The paper is structured as follows. Sec. II starts from the basic operating principle of a conventional ECS, where based on an equivalent transformer model and corresponding FEM simulations it is shown how the minimum applicable excitation frequency is influenced by the mover’s material properties and dimensions. Afterwards, in Sec. III, the influence of a conductive intermediate layer is analyzed with FEM simulations for different material combinations and an upper limit for the excitation frequency is given. Furthermore, a sensitivity analysis is conducted, in order to define the optimal operating frequency.
range for the selected materials. With the provided guidelines, then in Sec. IV the hardware setup with the excitation coil design, the measurement circuit and the test bench are discussed, while in Sec. V impedance measurements of the realized hardware prototypes are compared with the previously performed FEM simulations. In addition, the optional excitation frequencies are deduced for different material combinations and the corresponding sensitivities are listed, from which the achievable position accuracies can be calculated. Finally, Sec. VI concludes the paper.

II. OPERATING PRINCIPLE OF THE CONVENTIONAL ECS

In general, as shown in Fig. 2 (a) for a simplified planar structure, a single-coil ECS consists of a stationary excitation coil which is placed at a certain distance $d$ from a moving target with specific conductivity $\sigma_t$, relative permeability $\mu_{r,t}$ and thickness $d_t$. Since in many applications a paramagnetic material as e.g. stainless steel, copper or aluminum is selected for the target, in the following a permeability of $\mu_{r,t} \approx 1$ is assumed.

The coil is typically excited with a high-frequency AC current, generating a surrounding magnetic field $H_c$ which also penetrates the moving target. Hence, the high-frequency magnetic flux linked with the target induces a voltage in the target and, due to the target’s conductivity $\sigma_t$, it leads to eddy currents circulating in the target, which generate a second magnetic field $H_t$ counteracting the magnetic field of the coil. For a fixed excitation current and given material properties of the target, the intensity of this interaction, i.e. the magnetic coupling between coil and target, only depends on the distance $d$, which for a single-coil ECS configuration finally reflects into a measurable change of the coil’s input impedance $Z_{c,in}(\omega, \delta)$. Hence, by knowing the relation between $d$ and the variation of the coil input resistance and/or inductance, the position of the target is accurately measurable. Instead of measuring this distance-dependent impedance variation, in a dual-coil ECS configuration a second stationary pickup coil is used, where a distance-dependent induced voltage is measured. However, in this paper only the single-coil ECS is considered, since the findings are also directly applicable to the dual-coil ECS configuration, as the induced voltage variation is directly related to the inductance variation.

A. Equivalent Transformer Model

For the conventional single-coil ECS, the impedance variation of the excitation coil can be described by an equivalent transformer model as shown in Fig. 2 (b), where $R_c$ and $L_c$ are the (DC) resistance and inductance of the excitation coil, $R_t$ and $L_t$ are the resistance and inductance of the target, $\omega$ designates the angular excitation frequency and $M = k\sqrt{L_c L_t}$ the mutual inductance between coil and target, with the coupling factor $k$. Both the mutual inductance $M$ and the coupling factor $k$ depend on the angular frequency $\omega$ and the distance $\delta$, i.e. $M = M(\omega, \delta)$ and $k = k(\omega, \delta)$. Hence, for a given ECS setup, the excitation frequency $f_{exc} = \omega_{exc}/2\pi$ is selected in such a way that on the one hand a high measurement sensitivity and on the other hand a large enough measurement bandwidth is achieved. The measurement sensitivity is proportional to the variation of the coil’s input impedance $Z_{c,in}$, which can be split into an equivalent input resistance $R_{c,in}$ (real part) and inductance $L_{c,in}$ (imaginary part, divided by $\omega$), which according to the equivalent model are expressed as

$$R_{c,in} = R_c + \frac{\omega^2 M^2}{R_t^2 + \omega^2 L_t^2} R_t$$

(1)

$$L_{c,in} = L_c - \frac{\omega^2 M^2}{R_t^2 + \omega^2 L_t^2} L_t = L_c + L_{c,var}(\omega, \delta)$$

(2)

As can be noticed, for low angular frequencies $\omega$, the variable parts $R_{c,var}(\omega, \delta)$ and $L_{c,var}(\omega, \delta)$ vanish, and a distance-independent impedance $Z_{c,in} = R_c + j\omega L_c$ given by the coil resistance and its self-inductance is measured. Hence, the excitation frequency has to be selected above a certain lower frequency limit, which is defined by a coupling-independent angular cutoff frequency

$$\omega_{co} = 2\pi f_{co} = R_t/L_c$$

(3)

since around this frequency ($\pm 1$ decade) the input resistance increases by

$$R_{c,var}(\omega, \delta) = \left(\frac{M^2}{L_c}\right)^2 R_t = k(\omega, \delta)^2 L_c/\omega_{co}$$

(4)

while the input inductance decreases by

$$L_{c,var}(\omega, \delta) = -\left(\frac{M^2}{L_c}\right) = -k(\omega, \delta)^2 L_c$$

(5)

Accordingly, for a selected excitation frequency $f_{exc}$ around or above $f_{co}$, the resistance as well as the inductance variation can be used to measure the distance $\delta$ between target and coil. In this case, the coupling factor $k(f_{exc}, \delta)$ only depends on $\delta$ and thus for larger $\delta$, i.e. lower $k(f_{exc}, \delta)$, the measured resistance $R_{c,in}$ decreases, while the measured inductance $L_{c,in}$ increases. Furthermore, it is interesting to note that $\omega_{co}$ can be modified by the target material and thickness, i.e. the target resistance $R_t$ and the dimensions of the ECS, since $L_t$ scales proportionally to the coil’s self-inductance $L_c$. Consequently, for a given ECS geometry, a target material with higher conductivity $\sigma_t$ or larger thickness $d_t$ shifts $\omega_{co}$ to lower frequencies and in cases where the inductance variation is measured, either a lower $f_{exc}$ could be selected or a higher sensitivity can be achieved. In case of resistance variation, however, a lower $R_t$ also means lower sensitivity.

B. Impedance Variation obtained by FEM Simulations

Fig. 3 shows these relationships between cutoff frequency $f_{co}$, resistance and inductance variation $R_{c,var}(\omega, \delta)$ and $L_{c,var}(\omega, \delta)$ for a fixed target thickness of 0.5 mm and two different materials, i.e. stainless steel and aluminum. The distance $\delta$ is gradually increased between 0 mm (min. position), 1 mm (nom. position) and 2 mm (max. position), whereas here - for the conventional ECS without intermediate conductive layer - an offset of 0.5 mm has to be considered, which means that for $\delta = 0$ mm the effective distance between coil and target is 0.5 mm. The impedance curves of Fig. 3 are obtained from axisymmetrical 2D FEM simulations (ANSYS) for the ECS structure shown in Fig. 2 (e), whose excitation coil design is discussed in Sec. IV-A.

As can be noticed from Fig. 3, since aluminum has an approximately 26 times higher conductivity than stainless steel, its cutoff frequency...
The described behavior is verified in Fig. 5, which shows the simulated resistance and inductance curves for the distances $d = 0\text{ mm}, 1\text{ mm}, 2\mm$, and $\delta \rightarrow \infty$, where only the shield layer is in close proximity to the excitation coil, the shield acts like a target for the conventional ECS discussed in Sec. II. This means that $\pm 1$ decade around the cutoff frequency - now defined by the shield's $R_s$ and $L_s$ as $\omega_{co,s} = 2\pi f_{co,s} = R_s/L_s$ - the coil input resistance and inductance increase/decrease by the amount given in (4) and (5), respectively. As mentioned, since the shield is closely attached to the coil (there is no offset of 0.5 mm as considered for the targets of the conventional ECS), the magnetic coupling $k(\omega, \delta)$ is increased, resulting in a larger coil resistance and inductance variation around $\omega_{co,s}$. This is highlighted in Fig. 3, where in addition to the previous targets also the resulting coil resistances and inductances for shields made of 0.5 mm stainless steel ($\sigma = 1.45 \text{ MS} \cdot \text{m}^{-1}$) and aluminum ($\sigma_{alu} = 37.7 \text{ MS} \cdot \text{m}^{-1}$) are shown. As can be noticed, in order to achieve a high impedance variation, i.e. measurement sensitivity, a shield layer with high $R_s$ must be selected. This is also clear from the fact that with a lower $R_s$, larger eddy currents are induced in the shield, and thus the remaining flux linked with the target is small.

**TABLE II:** Different material and thickness combinations.

<table>
<thead>
<tr>
<th>Design</th>
<th>Shield Material</th>
<th>Shield Thickness</th>
<th>Target Material</th>
<th>Target Thickness</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1</td>
<td>Stainless Steel</td>
<td>0.5 mm</td>
<td>Aluminum</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>D2</td>
<td>Stainless Steel</td>
<td>0.2 mm</td>
<td>Aluminum</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>D3</td>
<td>Stainless Steel</td>
<td>0.5 mm</td>
<td>Stainless Steel</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>D4</td>
<td>Stainless Steel</td>
<td>0.2 mm</td>
<td>Stainless Steel</td>
<td>0.5 mm</td>
</tr>
</tbody>
</table>

If now a target is approaching the shield, the magnetic coupling to the target is increasing and additional eddy currents are induced in the target. At the coil terminals this is noticed as a reduction of the input impedance, where the variation is intensified with lower $R_s$. Hence, for an ECS measuring through a shield, the resistance $R_s$ shown in the equivalent circuit of Fig. 2 (b) for the conventional ECS, could be substituted by some equivalent resistance $R_{eq}$, which is a sort of a parallel combination of $R_s$ and $R_t$, however, where also the distance- and frequency-dependent couplings between coil, shield and target has to be considered. In other words, $R_{eq}$ can also be seen as some special target at some equivalent position having some mixed material properties of shield and target.

Nevertheless, independently from the exact value $R_{eq}$, this means that the impedance curve measured with only the shield (cf. Fig. 3), continuously decays if a target approaches the shield. Furthermore, as $R_{eq}$ is smaller for closer targets, also the cutoff frequency is shifted to lower frequencies. Unfortunately, the amount of reduction depends on the ratio of $R_s$ and $R_t$ as well as on their magnetic couplings to the coil, which are difficult to calculate. However, if for the extreme case $\delta = 0$ mm it is assumed that the shield and target have the same coupling to the coil, then the minimum cutoff frequency is roughly given by the parallel connection of $R_s$ and $R_t$ as

$$\omega_{co,min} = 2\pi f_{co,min} = R_s + R_t/((R_s + R_t)/L_s).$$

Consequently, in cases where the target has a much higher conductivity than the shield, $\omega_{co,min}$ is much lower than $\omega_{co,s}$, while for equal shield and target resistances, $\omega_{co,min}$ is roughly $\omega_{co,s}/2$.

**A. Impedance Variation obtained by FEM Simulations**

The described behavior is verified in Fig. 5, which shows the simulated resistance and inductance curves for the distances $\delta = 0$ mm, $1\text{ mm}, 2\text{ mm}$ and $\delta \rightarrow \infty$ (i.e. only the shield) as well as for different material and thickness combinations as summarized in Tab. II. The impedance curves are again obtained from FEM simulation, where the arrangement of Fig. 4 (b) is used.
the frequency shift from sensitivity is lost. Consequently, variation is visible, because reduction of the shield improves the sensitivities (cf. design D4).

Again, depth related frequency of the shield is reduced (cf. design D2), while the inductance variation is limited to a lower frequency range given by $f_{co,min}$ and $f_{co,s}$. As a rule of thumb, assuming that the optimal excitation frequency has to be located symmetrically within these boundaries, i.e. the geometric mean, the optimal excitation frequencies for the resistance and inductance variation can be roughly calculated as

$$f_{ext,opt,R} \approx \sqrt{f_{sk,s} \cdot f_{co,R}}, \quad f_{ext,opt,L} \approx \sqrt{f_{co,min} \cdot f_{co,s}}.$$  \hspace{1cm} (8)

As can be noticed from Tab. III, the calculated optimal excitation frequencies based on (8) are in good agreement with the optimal excitation frequencies obtained from Fig. 5. The discrepancy between FEM simulation and calculation is within 10\%, only $f_{ext,opt,L}$ of design D2 differs by 20\% and $f_{ext,opt,R}$ of design D3 differs by 30\%. However, even if for D3 the calculated frequency is selected, the max. resistance variation decreases by only 13\%.

C. General ECS Design Guidelines

Summarizing, from the FEM results and the considerations above it is possible to provide the following guidelines:

- In order to achieve a high measurement sensitivity, for both resistance and inductance variation, the shield resistance $R_s$ should be much larger than the target resistance $R_t$, which is either obtained by selecting different materials or by using different thicknesses.
- A large shield resistance $R_s$ increases both the cutoff frequency $f_{co,R}$ and the skin depth related frequency $f_{sk,s}$, resulting in a larger measurement bandwidth.
- The resistance variation starts around $f_{co,R}$ and ends around $f_{sk,s}$, which are both defined by only the shield layer.
- The range of inductance variation is defined by $f_{co,min}$ and $f_{co,s}$.
- The optimal excitation frequency is roughly calculated with (8), which is the geometric mean of the frequency boundaries.

IV. EXPERIMENTAL SETUP

A. PCB-embedded Excitation Coil Design

In general, the excitation coil has to be designed in such a way that on one hand, based on (4) and (5), a maximum impedance variation is achieved, and on the other hand this impedance variation caused by the target is not affected by the coil impedance itself, i.e. for example a strongly increasing AC resistance or a low coil self-resonance frequency.

A maximum impedance variation is obtained with a large self-inductance $L_e$, which in general means a maximum diameter and maximum number of turns. For the underlying application, the outer diameter of the coil (also a square coil shape would be possible) is limited to $d_{e,out} = 11.5\, \text{mm}$, while based on [15] the inner diameter should not be smaller than $d_{e,in} = 3\, \text{mm}$, since the most inner turns don’t enlarge the inductance noticeably but lead to a larger high-frequency coil resistance and lower quality factor. With the given PCB design constraints like minimum track width and distance of $150\, \text{μm}$, the maximum number of turns per layer is

As can be noticed for all designs, above a certain frequency, the skin and proximity effects of the shield become dominant and hinder the magnetic field to pass the shield layer, i.e. the target position sensitivity is lost. Consequently, $f_{exc}$ must stay below a certain skin depth related frequency $f_{sk,s}$ of the shield, deduced from the skin depth $\delta_{sk,s}$, where it is assumed that $\delta_{sk,s} = d_t$, given as

$$\delta_{sk,s} = \sqrt{\frac{1}{\pi \mu_0 \mu_r \sigma d_t}}, \quad f_{sk,s} = \frac{1}{\pi \mu_0 \mu_r \sigma d_t^2}. \hspace{1cm} (7)$$

### TABLE III: Characteristic and optimal excitation frequencies.

<table>
<thead>
<tr>
<th></th>
<th>D1</th>
<th>D2</th>
<th>D3</th>
<th>D4</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{co,min}$</td>
<td>4</td>
<td>4</td>
<td>71</td>
<td>71</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{co,s}$</td>
<td>126</td>
<td>282</td>
<td>126</td>
<td>282</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{sk,s}$</td>
<td>699</td>
<td>4367</td>
<td>699</td>
<td>4367</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{exc,opt,R}$ (FEM)</td>
<td>297</td>
<td>1109</td>
<td>297</td>
<td>1109</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{exc,opt,L}$ (FEM)</td>
<td>24</td>
<td>35</td>
<td>94</td>
<td>168</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{ext,opt,R}$ (FEM)</td>
<td>136</td>
<td>1259</td>
<td>430</td>
<td>1122</td>
<td>[kHz]</td>
</tr>
<tr>
<td>$f_{ext,opt,L}$ (FEM)</td>
<td>25</td>
<td>45</td>
<td>89</td>
<td>178</td>
<td>[kHz]</td>
</tr>
</tbody>
</table>

### Fig. 5: Normalized $R_{exc}(\omega, \delta)$ and $L_{exc}(\omega, \delta)$ versus frequency for the designs listed in Tab. II at $\delta = 0\, \text{mm}$, 1 mm, 2 mm and $\delta \rightarrow \infty$ (i.e. only shield), as obtained from FEM simulations. The obtained values of $R_t$ and $L_t$ are 2.812 and 13.9μH, respectively. The cutoff frequency $f_{co,s}$, the skin depth related frequency of the shield $f_{sk,s}$, and the optimal excitation frequencies $f_{exc,opt,R}$ and $f_{exc,opt,L}$, are also indicated (cf. Tab. III).
found as $N = 13$. The resulting self-inductance of a single-layer excitation coil is calculated as described in [16]–[20] and can be further increased if multiple layers of the PCB are used. In order to obtain the overall self-inductance, the magnetic coupling between the different layers has to be considered as given in [17], [18], [20]. In addition, a high-frequency ferrite core can be added on top of the coil to further increase $L_c$ and the magnetic coupling. However, besides maximizing $L_c$, it always must be ensured that the coil’s self-resonance frequency stays well above the desired excitation frequency range. Hence, the parasitic coil capacitance $C_{cp}$ - mainly consisting of the inter-winding and inter-layer capacitances as well as the capacitance to the core - must be small. There, especially the inter-layer capacitances lead to a strong increase of $C_{cp}$, and in order to keep the coil’s self-resonance frequency above 10 MHz, i.e. around one decade above $f_{exc}$, three layers are selected. A picture of the realized PCB-embedded excitation coil is shown in Fig. 6 (a).

B. Differential Sensor Configuration

According to Tab. III, the optimal excitation frequencies for inductance variation $f_{exc,opt,l}$ are considerably smaller than the ones for resistance variation $f_{exc,opt,r}$. Therefore, in order to achieve a sufficient measurement bandwidth for the underlying application, the resistance variation $\Delta R_{exc,vat}(f_{exc,opt,r}, \delta)$ is measured. Furthermore, since the nominal mover position is at $\delta = 1$ mm, and the position sensor should only output a voltage if the mover is displaced, for the measurement of the x- and y-displacement two identical ECS $ECS_x$ and $ECS_y$ are placed on opposite sides of the rotor, which are then used as a differential sensor in a Wheatstone bridge configuration (cf. Fig. 6 (b)) as proposed in [21]–[23]. This also eliminates any thermal drifts and dependencies of $\Delta R_{exc,vat}(f_{exc,opt,r}, \delta)$, and only leads to a temperature-dependent sensitivity, which can be compensated by measuring e.g. the stator temperature.

C. Mechanical Test Bench

Fig. 6 (c) shows the mechanical test bench used to verify the results obtained from the FEM simulations. It consists of two rigid plastic frames on which the 100 mm × 100 mm shield samples and the PCB-embedded coil are mounted. In between them, a pair of square-shaped 80 mm × 80 mm target samples of the same kind is glued on the two faces of a non-conductive support of thickness 10 mm,
As already expected from the measured impedance variations, in the achieved values are also reported for each design in full scale of the sensor, which in this case is D4.

The selected values of \( C_{\text{rel-opt}} \) and \( R_{\text{tot}} \) for the considered designs are reported in Tab. IV. Fig. 8 shows the measured \( U_{\text{br,out}} \) for each of the designs of Tab. II with respect to \( \delta_{\text{diff}} \), which corresponds to the displacement from the center position. The sensor output can be characterized with two main parameters, namely its sensitivity and percentage non-linearity. The former is directly computed by dividing the measured values of \( U_{\text{br,out}} \) by the corresponding value of \( \delta_{\text{diff}} \) and averaging the results. The latter is usually given in percentage as 

\[
(\Delta U_{\text{br,max}}/FS) \times 100\% \text{, where } \Delta U_{\text{br,max}} \text{ is the maximum absolute error between the measured data and its best linear fit, and } FS \text{ is the full scale of the sensor, which in this case is } 2 \text{mm for each ECS}. 
\]

The achieved values are also reported for each design in Tab. IV. As already expected from the measured impedance variations, in Fig. 8, design D2 with a thin shield and a highly conductive target offers the lowest sensitivity of about 0.9 mV/(μm) \( \delta_{\text{diff}} \). Design D1 and D4 feature similar sensitivities, which are already roughly half of design D2. However, design D4 has a 4 times higher optimal excitation frequency than D1, and thus enables also a larger measurement bandwidth. Finally, design D3 has the lowest sensitivity and also the lowest bandwidth, i.e., both almost 4 times smaller than in design D2. Furthermore, it can be noted that a larger sensitivity results in a higher non-linearity. Based on these considerations, the most suitable design for the targeted application would be design D2, however, for a pure stainless steel housing, design D4 is a good alternative offering a only \( \approx 1.6 \) times lower sensitivity but improved non-linearity, while the achievable bandwidth is the same as with aluminum. Furthermore, it also has advantage in terms of manufacturability in industrial applications, and can be even preferred to aluminum for its properties, corrosion resistance and sterilizability.

**VI. Conclusion**

This paper investigates the design of an eddy current sensor (ECS) measuring the position of a moving conductive target located behind another fixed conductive shield. The performed FEM simulations allow to characterize the position-dependent resistance and inductance variations measured at the excitation coil terminals versus frequency for different combinations of shield and target materials and/or thicknesses. In addition, design guidelines are provided to select the optimal excitation frequencies based solely on the properties of the shield and target materials. The analysis is verified with measurements on a sensor prototype, showing that the frequency selection guidelines are valid. For the best case of a 0.2 mm thick stainless steel shield and 0.5 mm thick aluminum target, a sensitivity of up to 0.9 mV/(μm) \( \delta_{\text{diff}} \) for a measuring range of 2 mm with a sensor measurement bandwidth in the order of 100 kHz can be achieved in a differential configuration. Therefore, the sensor is applicable for active control of MBs.

**REFERENCES**


