

ELECTROMAGNETIC NON-DESTRUCTIVE EVALUATION (XXI)



ELECTROMAGNETIC NON-DESTRUCTIVE EVALUATION (XXI)

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Electromagnetic Non-Destructive Evaluation (XXI)

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Preface

The ENDE international workshop has been held every year since 1995. Its aim is to bring together engineers and scientists from universities, research institutions and industry active in the research, development and industrial application of Electromagnetic Nondestructive Evaluation.

The 22nd edition of the workshop was held in CEA SACLAY DIGITEO LABS, Saclay, France, September 6–8, 2017. It was co-organized by the CEA LIST Institute and the Laboratoire des Signaux et Systèmes (L2S), with the sponsorship and support of the Digiteo-DigiCosme 2016 program, the University Paris-Saclay, the COFREND organization, and the JSM learned society. It was conducted under the co-chairpersonship of D. Lesselier (L2S) and C. Reboud (CEA LIST), assisted by a Scientific Committee and a Local Organizing Committee, under the auspices of the ENDE International Standing Committee chaired by Professor Zhenmao Chen, School of Aerospace, Xi'an Jiaotong University.

The ENDE topic is crucial to a great number of engineering activities, environmental evaluation and protection issues, and it is recognized for theoretical insight, efficient models and simulations, robust data interpretation, and proper instrumentation combined accordingly. A broad set of themes follows, including non-destructive testing and evaluation (NdT-NdE), static to THz electromagnetics, smart models and high-performance computations, advanced sensors, from design to usage, involving single- or multi-physics, inverse problems, imaging and signal processing in the context of uncertain and relatively scarce data, surrogate models, adaptive databases, model selection, and the qualification of uncertainty, multi-sensor data fusion, towards automatic approaches and decision making, complex material characterization from small to large scale, the monitoring and diagnosis of mechanical structures, and innovative industrial applications.

Ten oral sessions were organized at ENDE 2017 during the two-and-a-half days of the workshop: high frequency applications I & II; instrumentation I and II; material characterization I & II; theory and modeling; eddy current testing; low frequency inversion; and time domain methods. Two poster sessions enabled very interesting discussions in a more informal fashion. A total of 114 participants were registered (from Europe, the United States and Canada, China, Korea, Japan, and Thailand), and 3 invited keynotes, 47 oral contributions and 43 poster contributions were given in all, short versions of all such contributions being published in the Workshop Digest.

The first keynote was entitled “Product Uniformity Control – A Research Collaboration of European Steel Industries to Non-Destructive Evaluation of Microstructure and Mechanical Properties” and was given by Frenk van den Berg, Principal Scientist at Tata Steel, R&D IJmuiden, Netherlands. The second keynote, “Intrinsic modeling and inversion of ultra wideband microwave radar data for nondestructive testing of soils and materials”, was given by Sébastien Lambot, Professor in Agricultural and Environmental Engineering at UCL & FNRS Scientist, Louvain-la-Neuve, Belgium, and the third, “Numerical tools for the ultrasonic Non Destructive Testing of anisotropic plates”, was given by Sonia Fliss, Assistant Professor at ENSTA ParisTech, member of the Poems Laboratory (UMR 7231 CNRS-INRIA-ENSTA), Palaiseau, France.

On condition of being first authors and attending the workshop, students were proposed for the best presentation (oral or poster). The first prize went to Cesar Camerini for the communication he co-authored with João Marcos Rebello, Rafael Wagner Santos, João Marcio Santos, Tomasz Chady and Gabriela Pereira, entitled “Eddy Current System for Clad Pipe Inspection”. The second went to Takanori Matsumoto for the communication he co-authored with Tetsuya Uchimoto, Toshiyuki Takagi, Gabor Vertesy, Hidehiko Kage and Gerd Dobmann, entitled “Nondestructive Evaluation of Mechanical Properties of Cast Iron by Magnetic Incremental Permeability Method”, and the third prize went to Almpion Ratsakou for the communication he co-authored with Christophe Reboud, Anastassios Skarlatos and Dominique Lesselier, entitled “Fast Models Dedicated to the Simulation Of Eddy Current Thermography”.

In addition to an array of premier contributions, the 21 previous ENDE have been marked by edited peer-reviewed proceedings published by IOS Press as Electromagnetic Nondestructive Evaluation, within the series Studies in Applied Electromagnetics and Mechanics. The first (with the slightly different title of Non-Destructive Testing of Materials) appeared in 1995 under R. Collins, W.D. Dover, J.R. Bowler, and K. Miya. Twenty more followed. This suite of proceedings provides a unique perspective on the many challenges and advances of electromagnetic non-destructive evaluation, and adds significantly to the expert exchanges that have taken place at each workshop. The 22nd proceedings, following ENDE 2017, are proposed herein, as an Open Access publication on the IOS Press e-book platform, including 1 keynote and 34 peer-reviewed extended contributions.

To conclude, the organizers would like to thank all those Committees members who have tirelessly contributed to the workshop, and the many reviewers of the extended papers who have played a key role in the making of these proceedings.

Saclay, March, 2018
D. Lesselier, L2S
C. Reboud, CEA LIST
Editors



ENDE 2017 – Saclay, France

Conference Organization

The 22nd International Workshop on Electromagnetic Nondestructive Evaluation was held in Saclay, France, 6–8 September 2017.

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Evaluation of Hardened and Non-Hardened Regions Using Eddy Current Testing

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Abstract. A steel plate having hardened and non-hardened regions in one plate is promising material in automobile field owing to its high performance. This steel plate has to be evaluated whether the non-hardened region was prepared properly at the desired position with desired hardness. However, the conventional evaluation method of hardness measurement is destructive testing and requires a long inspection time. Therefore, a reliable and fast nondestructive inspection method is required. In this study, the hardened and non-hardened regions were evaluated using an eddy current testing (ECT) to distinguish each region. The measured resistivity and permeability at each region were different and these differences were found to be caused by the crystal structure change of heat treatment. Therefore, the magnetic response between the hardened and non-hardened regions was different and this difference was correlated with the hardness of each region. These results indicate that the hardened and non-hardened regions can be distinguished using ECT.

Keywords. Eddy current testing, hardness, permeability, magnetic sensor, phase

1. Introduction

Steel plate is widely used in various industrial fields and the hardness is one of the important factors to ensure the durability and safety of steel plate. In particular, in the auto mobile field, the hardness of steel plate is controlled depending on the car components. To control the hardness of steel plate, a heat treatment method is generally employed and the desired hardness is realized by selecting the optimum conditions such as temperature, cooling method, and so on. Recently, a new hot press forming method was developed which can prepare a non-hardened region at the desired position in a one hardened steel plate [1-4]. Because of the different hardness in the hardened and non-hardened regions, this steel plate can reduce the welding fracture and control the bending position in the case of automobile crash. Therefore, the steel plate with a partially non-hardened region in a hardened steel plate is expected to be used in the auto mobile field.

Considering the practical use of steel plate with a partially non-hardened region, it should be evaluated that the hardened and non-hardened regions are prepared properly at the desired position. However, the conventional evaluation method is a hardness measurement. This method is destructive testing and the oxide film generated during heat treatment should be removed for hardness measurement. Moreover, the multipoint measurement is also required to evaluate whether the non-hardened region is prepared at the desired position or not. For these reasons, the hardness measurement requires a

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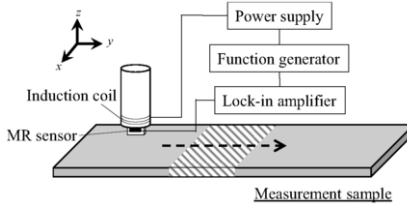


Figure 1. Schematic of measurement system. using eddy current testing.

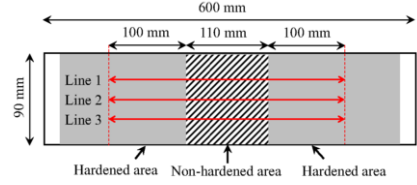


Figure 2. Measured steel plate with hardened and non-hardened regions.

long inspection time and is difficult to evaluate all samples. Therefore, a reliable and fast nondestructive evaluation method for the hardened steel plate with a partially non-hardened region is required. Although some non-destructive evaluation methods of hardness have been attempted [5-7], it is unknown that these method are applicable to the evaluation of a steel plate with hardened and non-hardened regions.

In this study, we investigated to distinguish the hardened and non-hardened regions using an eddy current testing (ECT) which can solve the above mentioned problems. This is because the electric and magnetic properties of each region are considered to be different owing to the structure change of steel plate and the magnetic signal response of ECT is expected to change at each region. To clarify this phenomenon, the electrical resistivity and the magnetization curve (B - H curve) of each region were measured and a magnetic signal change at the interface of hardened and non-hardened region was calculated by a finite element method (FEM). Then, a suitable condition for the evaluation of the steel plate with a partially non-hardened region was investigated by measuring the steel plate using ECT.

2. Experimental procedure

2.1 Measurement system

The developed measurement system of ECT used in this study is composed of an induction coil, a magnetic sensor, a function generator, a power supply, and a lock-in amplifier (Fig. 1). To evaluate the hardened and non-hardened regions inside a steel plate, a low frequency magnetic field was applied to avoid skin effect and magnetoresistive (MR) sensor was used because it shows high sensitivity in the low frequency range [8].

The magnetic field was applied to the measurement sample using an induction coil and the secondary magnetic field from a sample was detected by a MR sensor. The inner and outer diameters of induction coil are 20 and 25 mm, respectively, and the turn number is 50. To reduce the effect of applied magnetic field and obtain a wide dynamic range, a compensation coil was wound outside of MR sensor. The compensation coil is directly connected to the input coil with reverse winding and the turn number is 10. The direction of the detected magnetic field is parallel to the applied magnetic field (z -axis direction) and the magnetic signal intensity and phase were measured using a lock-in amplifier. The measurement probe (induction coil and MR sensor with compensation coil) was fixed to an automatically controllable X-Y stage and the multipoint measurement was performed by controlling the X-Y stage using a PC.

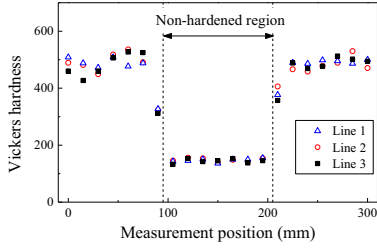


Figure 3. Vickers hardness of hardened and non-hardened regions.

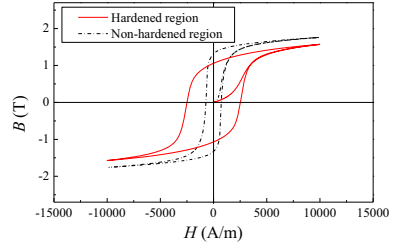


Figure 4. B - H curve of hardened and non-hardened regions

2.2. Sample and measurement conditions

A steel plate of 90 mm $\times$ 600 mm $\times$ 1.4 mm was used as a measurement sample, and the main chemical composition of this steel plate is as follow; C: 0.21 mass%, Si: 0.25 mass%, Mn: 1.27 mass%, Cr: 0.19 mass%. The hardened (martensite structure) and non-hardened (ferrite and perlite structure) regions were prepared in this steel plate by controlling the heat treatment temperature of each region in the air. The non-hardened region was prepared at the center of steel plate with the width of 110 mm and both sides of non-hardened region were hardened region (see Fig. 2). The magnetic field intensity and phase were measured at 3 lines with the interval of 10 mm which contain both regions, as shown in Fig. 2. Although the oxide film was generated on the surface of steel plate, this film was remained in the measurement.

To investigate the difference of electrical and magnetic properties between hardened and non-hardened regions caused by the heat treatment, the B - H curve was measured using a ring specimen and the electrical resistivity ρ was measured using the 4-terminal method.

3. Results and discussions

3.1. Material properties at hardened and non-hardened regions

To confirm that the prepared steel plate has a desired hardness at the desired position, the Vickers hardness of the hardened and non-hardened regions was measured. The sample for hardness measurement was prepared under the same conditions as the measurement sample for ECT and the hardness was measured at the center line and 15 mm above and below from the center line. The interval of measured hardness was 15 mm. The Vickers harnesses at hardened and no-hardened regions were approximately 500 and 150, respectively, and the tendency of hardness change at each line was almost the same, as shown in Fig. 3. Therefore, it was confirmed that the prepared sample had desired hardness at desired position.

The measured values of ρ at hardened and non-hardened regions were 2.56×10^{-7} and 2.32×10^{-7} Ω m, respectively. The B - H curves of each region were also different, as shown in Fig. 4. It was found from Fig. 4 that the permeability μ of the hardened region is lower than that of the non-hardened region in the range of small applied magnetic field used in ECT. Generally, the crystal structure changes depending on the

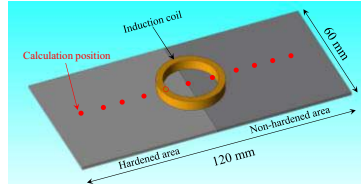


Figure 5. Model used for calculating the magnetic field response at hardened and non-hardened regions.

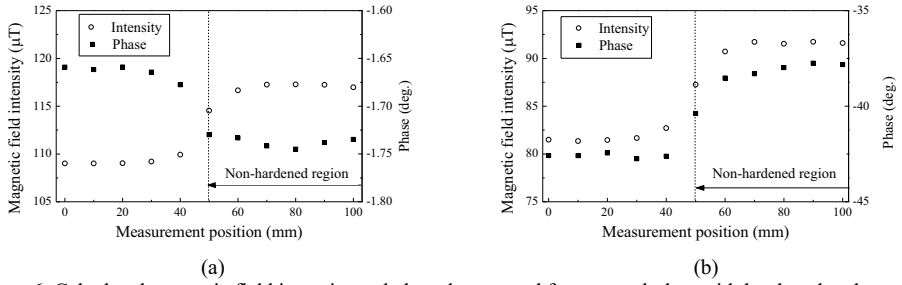


Figure 6. Calculated magnetic field intensity and phased generated from a steel plate with hardened and non-hardened regions. The frequencies of applied magnetic field are (a) 100 Hz and (b) 5 kHz.

heat treatment condition and it was found from the optical-microscopic observation that the crystal structures of the hardened and non-hardened regions were martensite and ferrite + perlite structure, respectively. Thus, the differences in ρ and μ at the hardened and no-hardened regions are considered to be caused by the difference of crystal structure at each region, i.e. the steel plate containing the martensite structure has higher ρ and lower μ than those of the steel plate containing the ferrite + perlite structure. As it is well known that the hardness is strongly correlated with the crystal structure of steel plate, ρ and μ are considered to be correlated with the hardness. This suggests that the hardness difference can be evaluated using ECT because the magnetic signal response of ECT strongly depends on the ρ and μ .

3.2. Calculated and measured magnetic signal at hardened and non-hardened regions

To investigate the difference of magnetic signal response from the hardened and non-hardened regions owing to the different ρ and μ at each region, the magnetic signal intensity and phase were calculated by FEM. For the calculation, a commercially available FEM software, JMAG, was used. This software uses an iterative method for the calculation of FEM. In the calculation, two plates of 60 mm $\times$ 60 mm $\times$ 1.4 mm were used and the magnetic field was applied to the plates at each calculation position. The calculation positions are 11 points at the center line of steel plate and the interval of each measurement point is 10 mm (Fig. 5). Tetrahedral meshes were generated for the calculation model of Fig. 5. The ρ and B - H curve of two plates used for the calculation were the measured values discussed in section 3.1 which correspond to the hardened and non-hardened regions. The dimension of coil was the same as that of measurement system and the material properties of coil was a copper.

In the calculation, the frequencies of applied magnetic field were 100 Hz and 5 kHz. The skin depth at these frequencies are 2.3 and 0.5 mm, respectively, which is

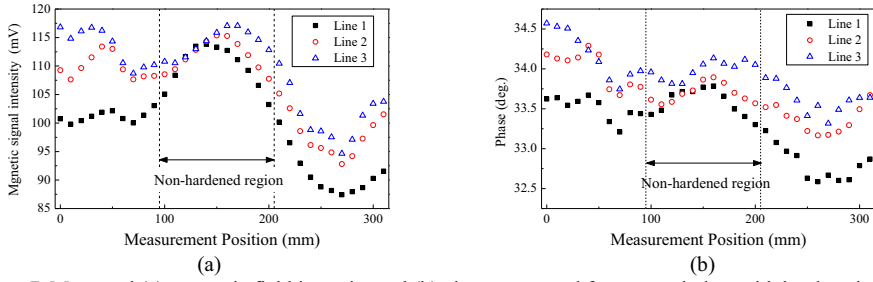


Figure 7. Measured (a) magnetic field intensity and (b) phase generated from a steel plate with hardened and non-hardened regions measured at 100 Hz.

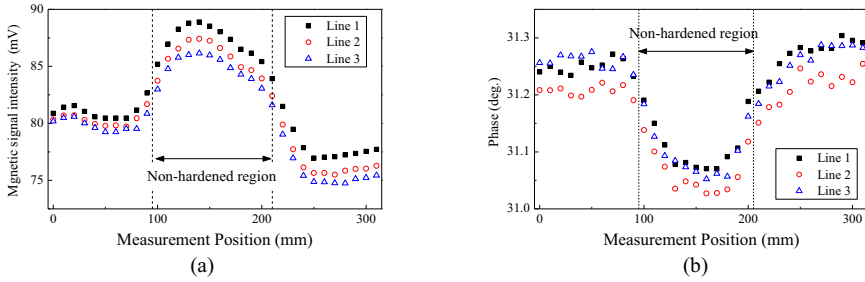


Figure 8. Measured (a) magnetic field intensity and (b) phase generated from a steel plate with hardened and non-hardened regions after demagnetizing the steel plate. The frequency of applied magnetic field is 100 Hz.

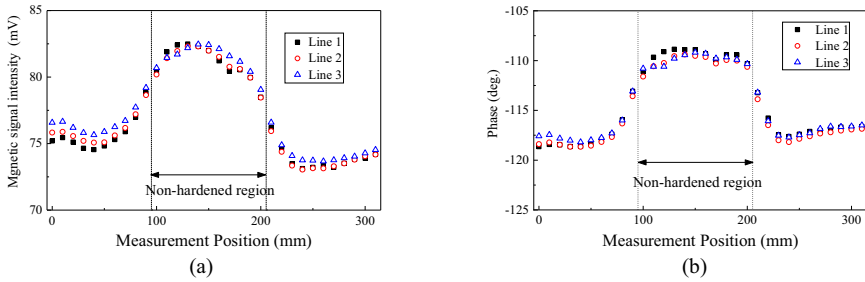


Figure 9. Measured (a) magnetic field intensity and (b) phase generated from a steel plate with hardened and non-hardened regions after demagnetizing the steel plate. The frequency of applied magnetic field is 5 kHz.

estimated from the measured ρ and μ . This is because the magnetic field of 100 Hz penetrates enough inside the sample and that of 5 kHz penetrates the quarter of sample thickness. Figure 6 shows the calculated magnetic field intensity and phase at each frequency. In both frequencies, the intensity and phase changed near the interface between hardened and non-hardened regions. This result suggests that the magnetic response depends on the hardened and non-hardened regions which have different ρ and μ values. Thus, the hardened and non-hardened regions can be detected using ECT. In addition, it was found from Fig. 6(b) that the differences in the magnetic field intensity and phase can be enough obtained even if the penetration depth of magnetic field is one quarter of plate thickness.

To confirm the validity of calculated results, the magnetic field intensity and phase were measured at 100 Hz using ECT and the measured results are shown in Fig. 7. Although the intensity and phase changed depending on the measured region, the

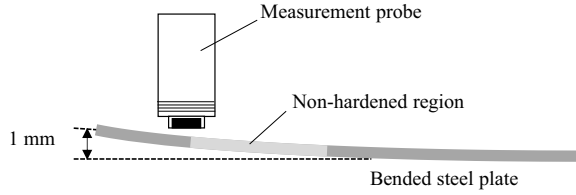


Figure 10. Schematic of cross sectional view of bended steel plate and measurement probe.

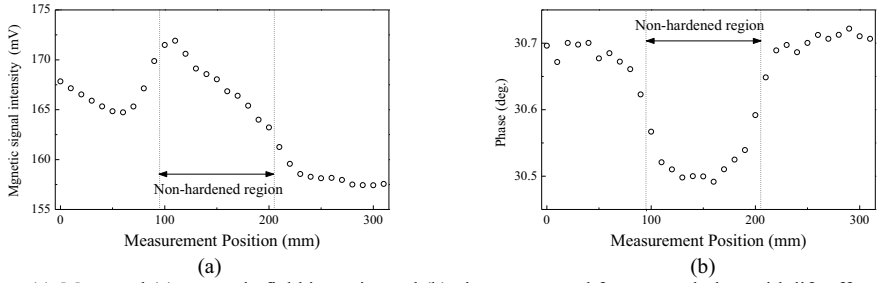


Figure 11. Measured (a) magnetic field intensity and (b) phase generated from a steel plate with lift-off change of approximately 1 mm. The frequency is 100 Hz.

difference between the hardened and non-hardened regions was not clear like the calculated values shown in Fig. 6(a). Moreover, the change of intensity and phase is different in each line and the intensity and phase are also different in both side of hardened region in spite of the same material properties. The reason for this fluctuation is estimated to be the local magnetization of the steel plate before or during the heat treatment, because the fluctuation due to localized magnetization differences is one of the problems in the magnetic nondestructive testing [8]. Therefore, the measurement sample was demagnetized using a commercially available demagnetizer. Figure 8 shows the magnetic field intensity and phase after demagnetization. The frequency of applied magnetic field is 100 Hz. When the sample was demagnetized, the intensity and phase showed different values only at the non-hardened region, and these changes corresponded to the calculate results shown in Fig. 6(a). In addition, when the sample was demagnetized and the magnetic field of 5 kHz was applied to the sample, the intensity and phase were also different between the hardened and non-hardened regions as shown in Fig. 9. These differences were almost the same as the calculated value (Fig. 6(b)). These results indicate that the signal change caused by the differences of ρ and μ can be detected by demagnetization after the heat treatment and the difference of hardness is evaluated using ECT.

3.3. Effect of lift-off and surface oxide film on the magnetic response

In the previous section, it was clarified that the hardness change can be detected using ECT. However, there are some phenomena which disturb the magnetic response. In this section, the effects of the distance between magnetic sensor and measurement sample (lift-off) and the oxide film on the magnetic field intensity and phase were investigated.

The steel plate for a practical use is not completely flat and the distance between the sample and magnetic sensor changes depending on the roughness or distortion of

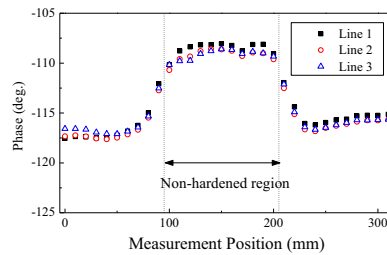


Figure 12. Measured magnetic field phase generated from a steel plate after demagnetizing the steel plate and removing the oxide film measured at 5 kHz.

steel plate. Thus, the steel plate which has the lift-off change approximately 1 mm caused by distortion of plate was measured (see Fig. 10). The measured magnetic field intensity and phase are shown in Fig. 11. The frequency was 100 Hz. When the lift-off changed owing to the distortion of steel plate, the magnetic field intensity had slope-like change which is related to the distortion of steel plate. This is because the magnetic field intensity sensitive to the change of lift-off. However, this slope-like change was not observed in the phase, as shown in Fig. 11(b). It has been reported that the effect of lift-off change on the phase is smaller than that on the intensity [9]. Because of this phenomenon, the difference of phase caused by the hardness change is detected even if the lift-off changed. However, the amount of phase change is small at 100 Hz and as predicted by the theoretical calculation, and the amount of phase change increases if the magnetic field of 5 kHz is applied (Fig. 6(b)). This was also confirmed from the measured result of phase at 5 kHz, as shown in Fig. 9(b). When the frequency is 5 kHz, the skin depth is approximately one quarter of sample thickness. The cross sectional view of the steel plate obtained using a microscope showed that the steel plate was heat-treated uniformly to the depth direction. Therefore, the skin depth of 5 kHz is enough to evaluate the hardened and non-hardened region. From the above discussions, the phase of magnetic field of which frequency is 5 kHz is suitable for evaluating the hardened and non-hardened regions prepared in one steel plate.

Next, to investigate the effect of oxide thin film, the phase was measured after removing the oxide film. In general, the oxide thin film composited of FeO, Fe₃O₄, and Fe₂O₃ layers are generated during a heat treatment in the air. Thus, the effect of the oxide thin film on the magnetic signal is considered because the Fe₂O₃ is a ferromagnetic material. However, as shown in Fig. 12, the measured phase at 5 kHz after removing the oxide film was almost the same as that measured before removing the oxide film (Fig. 9(b)). Therefore, it was found that the oxide film does not affect the magnetic response of ECT and hardness evaluation using ECT can be performed without removing the oxide film. This result leads to the reduction of inspection time.

Finally, the phase was measured at the center area of 40 mm × 310 mm with 10 mm intervals and a phase mapping was created to visualize the position of non-hardened region, as shown in Fig. 13. The frequency of applied field was 5 kHz and the phase was measured without removing the oxide film. When a suitable threshold which distinguishes the hardened and non-hardened regions was selected, the non-hardened region was clearly visualized, and the position of non-hardened region was corresponded to that confirmed by the hardness measurement. From this result, it was found that the position of non-hardened region can be visualized with short inspection time using ECT.

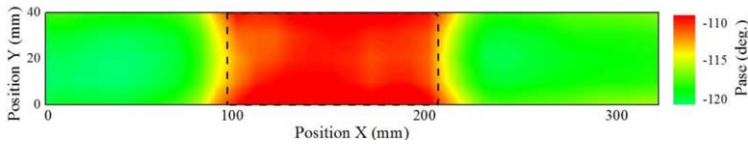


Figure 13. Phase mapping of measured at the center of steel plate containing the hardened and non-hardened regions. The frequency of applied magnetic field is 5 kHz.

4. Conclusions

A fast non-destructive evaluation for the partially non-hardened region prepared in one hardened steel plate was investigated using ECT. The measured values of ρ and μ at hardened and non-hardened regions were different. From the calculation of magnetic signal intensity and phase using the measured ρ and μ , it was confirmed that the signal intensity and phase are different at hardened and non-hardened regions and the evaluation of these regions using ECT was suggested. Although the measured magnetic signal and phase were affected by the local magnetization of steel plate, this was removed by the demagnetization and the signal change caused by only the difference of ρ and μ was obtained. Moreover, the phase was not affected by the lift-off change and clear phase change was observed when the frequency of applied magnetic field was 5 kHz. Furthermore, it was also found that the phase change between the hardened and non-hardened regions can be measured without removing the surface oxide film generated during heat treatment process and the non-hardened region can be visualized by the multipoint measurement of phase. If a certain calibration is performed, a quantitative evaluation of hardness difference is expected. These results indicate that a practical evaluation of the partially non-hardened steel plate is possible using ECT.

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High-Speed MFL Method Based on Multistage Magnetization Structure

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Abstract. In the process of magnetic flux leakage (MFL) testing, the leakage magnetic field decreases as the testing velocity increases, which leads to the severe detection omission. To address this issue, factors affecting high-speed magnetic flux leakage testing are analyzed, multistage magnetization structure is then proposed. By comparing magnetic flux leakage signals characteristics on the surface of steel pipe with high speed and low speed, effectiveness and feasibility of multistage magnetization structure are analyzed. Results show that multistage magnetization structure can effectively extend magnetization time. Compared with single magnetization, the magnetic flux leakage signal of the inner wall of the steel pipe can be detected effectively with multistage magnetization structure at high speed, and the magnetic flux leakage testing precision is improved.

Keywords. High-speed MFL testing, Multistage magnetization, Helmholtz, Eddy current

1. Introduction

With the rapid development of modern industrial technology, the ferromagnetic metal materials have been widely used in construction, aerospace, energy, transportation and other fields. In the use of ferromagnetic material components, the damage produced by external forces, corrosion and other factors will make the health of metal components gradually deteriorated, which leads to great potential dangers. As an efficient nondestructive testing method, it realizes nondestructive detection of components by identifying magnetic flux leakage field signals at defects. It can not only detect many types of defects, but also determine the geometric shape of defects. Magnetic flux leakage testing technology has been widely used in many industrial fields, magnetic flux leakage testing technology can realize the on-line nondestructive testing of the pressure pipeline, steel pipe, rail and other pressure equipment. However, in various applications, the magnetic flux leakage testing technology will be affected by the same factor, that is the detection speed. The pipeline magnetic flux leakage detector is usually running at 5m/s, and the defect will be seriously missed if the speed exceeds 5m/s. And in other applications, the same phenomenon occurs. The on-line detection speed of steel pipe is about 3m/s, and the running speed of rail inspection vehicle is about 10m/s.

However, in the process of magnetic flux leakage testing, the components must be firstly magnetized to saturation and then the defects can be detected effectively, which

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severely limits the testing speed and testing effect of the magnetic flux leakage testing technology^[1-3].

In this paper, a high-speed dynamic experimental platform of magnetic flux leakage detection is built, and the multistage magnetization structure is proposed firstly. The multistage DC (direct current) magnetization coil can make steel pipe arrive saturated, and three-dimensional high speed magnetic flux leakage detection sensors are used to detect magnetic flux leakage signal of surface defects. By comparing the characteristics of magnetic flux leakage field signals of the defects in same steel pipe at high speed and low speed operation, the influences of magnetizing coil with different magnetization are analyzed.

2. Theoretical analysis of high speed dynamic magnetization

High speed magnetic flux leakage testing schematic diagram as shown in Figure 1, the principle of magnetic leakage detection technology based on the high magnetic permeability of the ferromagnetic material. In high speed magnetic flux leakage detection process, detection device will move at a certain speed along the specified direction of the measured component^[4]. In Figure 1, L represents the distance between the magnetization device and the detection sensor, and the v represents the forward speed of the detector.

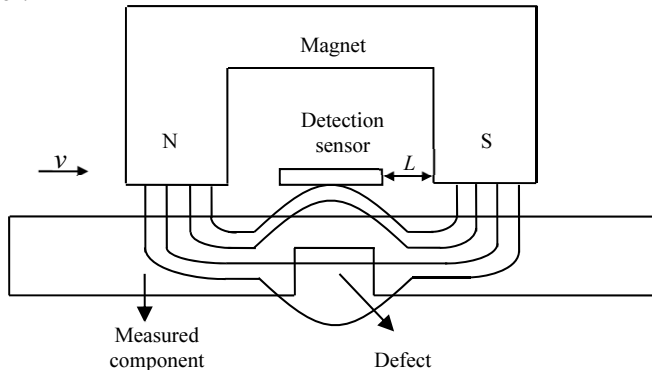


Figure 1. Schematic diagram of high speed magnetic flux leakage testing

From the magnetization to the leakage signal detected by detector, the effective magnetization time of the component can be expressed as:

$$t = \frac{L}{v} \quad (1)$$

In static or low-speed magnetization process, $v=0$ or v is very small, the magnetization time is enough and the measured component is easy to achieve magnetic saturation. With the speed increase of detection, the effective magnetization time is reduced greatly. And when the magnetic flux leakage detector through the defect area, the measured components have not yet reached the magnetic saturation, which lead to severe detection omission even serious missed detection at high speed.

According to the Stoner criterion, the magnetization of the measured component can be expressed as^[5]:

$$B = B_0 + B_1 \quad (2)$$

$$B_0 = \mu_0 H \quad (3)$$

$$B_1 = \mu_0 M \quad (4)$$

Where B_0 represents magnetic induction intensity produced by the external magnetic field, and B_1 represents magnetic induction intensity produced by solid itself. The magnetic signal detected by the magnetic leakage detection sensor can be expressed as:

$$B = \mu_0 (H + M) = \mu_0 \left(H + \frac{d \sum_i \mu_i}{dV} \right) \quad (5)$$

Where μ_0 represents magnetic permeability in vacuum, M represents magnetization intensity of the component, μ_i represents the magnetic moment of the atom, H represents the applied field strength.

3. Multistage coils magnetization principle

In high speed testing process, the magnetization time is very short, and magnetization will take a certain time to reach a stable state. The "delay phenomenon" caused by the rapid relative movement between the detection device and component will seriously affect the detection effect^[6]. In addition, the eddy current effect caused by the relative movement will also prevent the detection of defects at high speed testing process. In order to solve the problem that the component is not easy to be saturated, the multistage magnetization structure is proposed, which uses multistage direct current magnetization coils to magnetize the component.

The multistage magnetization structure is based on Helmholtz coil principle, as shown in Figure 2, the two coils are coaxially arranged, the distance is equal to the radius of the coil and the same current I is put into the coil.

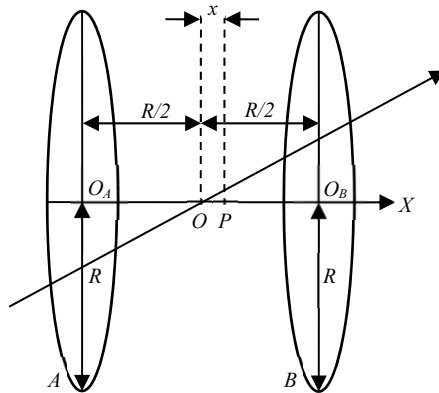


Figure 2. Schematic diagram of Helmholtz coil

Where R represents the radius of coil, x represents the distance from any point on the axis of coil to the center.

The magnetic induction intensity B of any point on the coil axis can be expressed as:

$$B = \frac{\mu_0 N R^2 I}{2 \left[R^2 + \left(\frac{R}{2} + x \right)^2 \right]^{\frac{3}{2}}} + \frac{\mu_0 N R^2 I}{2 \left[R^2 + \left(\frac{R}{2} - x \right)^2 \right]^{\frac{3}{2}}} \quad (6)$$

Where N represents the number of coil turns, and μ_0 represents magnetic permeability in vacuum.

By Eq. (6), magnetic field variation curve of two-stage coils is shown in Figure 3.

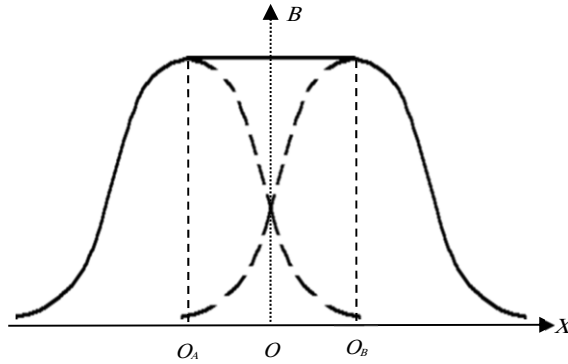


Figure 3. The magnetic field variation curve of two-stage coils

Where O_A and O_B respectively represent the two-stage coils center point. As shown in Figure 3, the magnetic field produced by the superposition of two coils is approximately uniform magnetic field, and the magnetic field intensity is the maximum value of the whole magnetic field.

The eddy current intensity in the steel pipe is related to the change rate of the axial component of the external magnetic field when the steel pipe is magnetized at high speed^[7]. The eddy current intensity is stronger when the change rate of the external magnetic field is larger. The multistage magnetization coils which produce a uniform magnetic field will reduce the eddy current effect and prolong the excitation time effectively at the same detection speed.

4. Experiment and results analysis

4.1 Experiment

In order to verify the effectiveness of the multistage structure, the high-speed magnetic flux leakage detection platform is built.

The schematic diagram of multistage magnetization structure is shown in Figure 4.

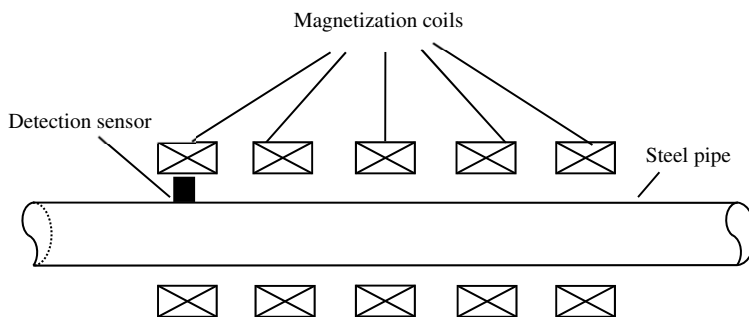


Figure 4. Schematic diagram of multistage magnetization structure

The experiment adopts 20# seamless steel pipe, the length of steel pipe is 1200mm, the diameter is 102mm, and the wall thickness is 10mm. And in order to verify the magnetization effect of the multistage magnetization structure, the defects are made inside the steel pipe and the sensors are placed on the outer wall of the pipe. The depth of the defect is 0.2mm, the width is 10mm and the two defects are axially spaced 1m. As shown in Figure 5, the multistage magnetization structure adopts the form of 5 direct current coils arranged coaxially, and the number of coil turns is consistent and the current with same direction and size is passed into the coils. As shown in Figure 6, in order to improve the reliability of the detection, four detection sensors which can detect three axis magnetic leakage signals are placed at the end of each coil.

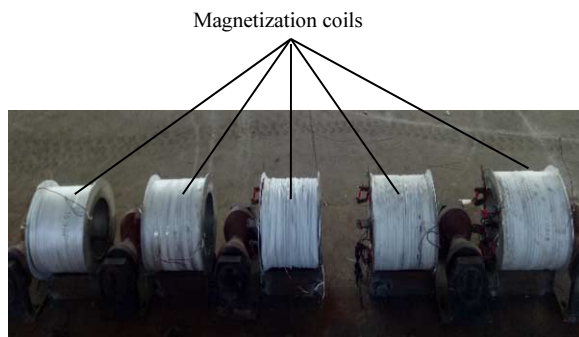


Figure 5. Multistage magnetization structure

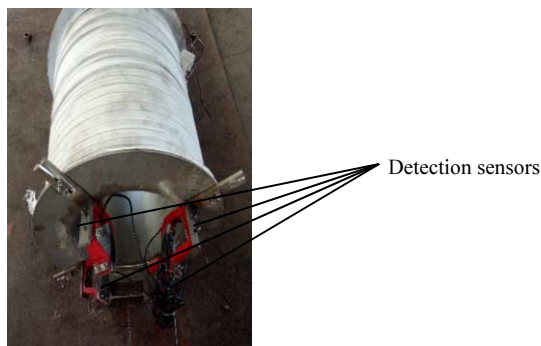


Figure 6. Detection sensors

Five sets of circuit device are designed to supply power to the coil, each device include voltage regulator, rectifier bridge, capacitor, current meter etc. The voltage regulator is used to adjust voltage to change the coil current slowly. And at the same time, the rectifier bridge will convert alternating current to direct current to the magnetization coil.



(a) Voltage regulator



(b) Circuit connection

Figure 7. Power supply system

The steel pipe can be gradually saturated by the magnetic field produced by each coil and magnetization time can be extended. With multistage magnetization structure, steel pipe can be quickly saturated and achieve defects detection with high speed.

4.2 Results analysis

In order to verify the validity of the multistage magnetization structure, the coils are passed into different current firstly. The change trend of magnetic flux leakage magnetic field signal and current is shown in Figure 8.

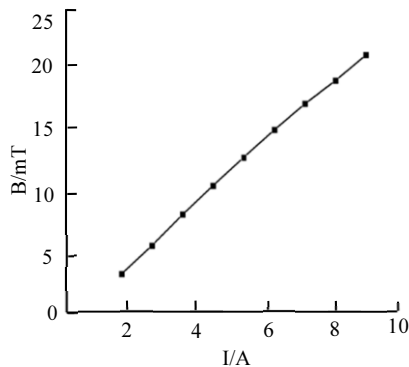
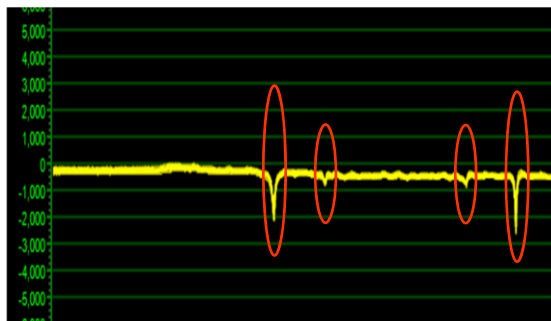


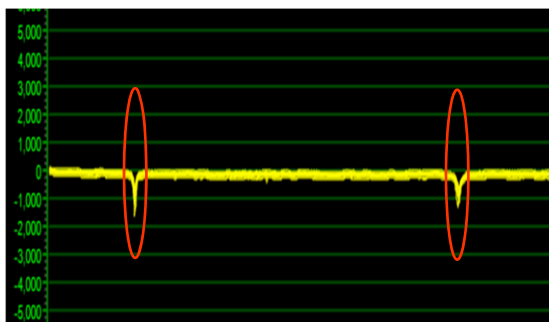
Figure 8. Experimental curve of magnetic flux leakage field signal

As can be seen in Figure 8, the magnetic flux leakage field signal is larger with the increase of current. If the current is greater, then the magnetic field generated by the coil becomes stronger, and the magnetic flux leakage field signal is more likely to be detected.

When current is consistent, the axial magnetic flux leakage at different speeds is shown by software in figure 9. The abscissa axis and ordinate axis respectively represent the working time of detection sensor and the digital quantity detected by detection sensors, and the leakage magnetic field can be obtained by the corresponding conversion of digital quantity and chip resolution.



(a) $v=0.5\text{m/s}$



(b) $v=5\text{m/s}$

Figure 9. Magnetic flux leakage signal at different speed

As can be seen from Figure 9, when steel pipe operating speed is 0.5m/s, the detection sensor can detect defects in the inner wall of the steel pipe with single magnetization coil, and the two larger signals are produced by weld line, the two smaller signals are produced by defects. However, the detection sensor can't detect signal produced by defects when operating speed is 5m/s.

The experimental results show that the magnetization process has a certain hysteresis, magnetization is not instantaneous. When the multistage coils are used for magnetization, the defects signals detected by sensor are more obvious at high speed.

5. Conclusion

In the process of high-speed magnetic flux leakage detection, the effective magnetization time will be reduced, and the component will not realize magnetic saturation rapidly. Multistage magnetization structure can effectively extend magnetization time. When operating speed of steel pipe is 5m/s, it has remarkable magnetization performance. Compared with the single-stage magnetization structure, multistage magnetization structure can improve testing precision, which provides a feasible solution for the comprehensive detection of defects with high speed. The magnetization effect is good and the magnetic leakage signal of the inner wall of the steel pipe can be effectively detected. Besides, the multistage magnetization structure can improve the accuracy of the magnetic flux leakage nondestructive testing at high speed and provide a feasible scheme for realizing high-speed magnetic flux leakage nondestructive testing.

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A New Probe for Velocity Induced Eddy Current Inspection

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Abstract. Velocity induced eddy current inspection is a technique to inspect conductive materials. It has a good detection performance and is particularly well suited to situations where the speed is a factor to be considered because its sensibility increases with the moving speed. This paper assesses the field produced by several probe geometries in the detection of linear machined defects. A more detailed analysis is carried out for a probe that includes an array of cubic magnets to induce the eddy currents inside the sample and Hall sensors to measure the resultant magnetic field.

Keywords. Nondestructive testing, eddy current testing, velocity-induced eddy current testing, Hall sensor.

1. Introduction

Preventive maintenance is a critical issue in safety and cost of several industrial applications [1, 2]. Due to its high sensitivity eddy current testing is one of the most used techniques to inspect conductive materials [3, 4]. A modified version of this method was proposed [5, 6] to fit out situations where there is a relative motion between the material to be inspected and the equipment that performs the measurement. The basic physical phenomena of this method, designated velocity induced eddy current is based on the interaction of a stationary magnetic field and a specimen which moves with a constant speed in relation to that field. It is a very sensitive method and reduced power is needed to apply the technique as no excitation current is required.

The literature [7-9] mentions that the performance of the method depends greatly on the configuration of the magnets included in the probe and also in the assessment of the magnetic field. Enhancements can be achieved not only by increasing the magnetic flux density in the material to be inspected, but also, by choosing the best position for the magnetic sensors. This paper reports results obtained experimentally for several probe geometries. It studies data measured with different probes when a sample with machined cracks is inspected. The study is presented step by step, meaning that, incremental

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enhancements are added to the probe and results are observed and compared. A compromise between weight and volume was also considered. It was concluded that a probe consisting of an array of cubic magnets that have alternating vertical magnetization and a set of small Hall effect sensors placed in the interstices showed the best performance to evaluate machined cracks in a plate using velocity induced eddy current inspection technique.

The remainder of the paper is structured into four sections. In the following section the experimental setup is shown. In Section 3 after a brief introduction to the NDT method using velocity induced eddy currents, enhancements to previous designs of the magnet source and of the detection sensors are proposed. Section 4 contains the experimental validation that confirms the improvement of the method with the use of a new probe by detecting four defects in an aluminium plate. Finally the conclusions are summarized in Section 5.

2. Experimental Setup

Using the experimental setup shown in Fig. 1. a sample of 4 mm thick 1050 aluminium plate with four machined defects was inspected. All defects are 50 mm long, 0.5 mm wide and depth varies from 0.5 mm to 2 mm with intervals of 0.5 mm. It was scanned at a fixed speed of 4 m/s.

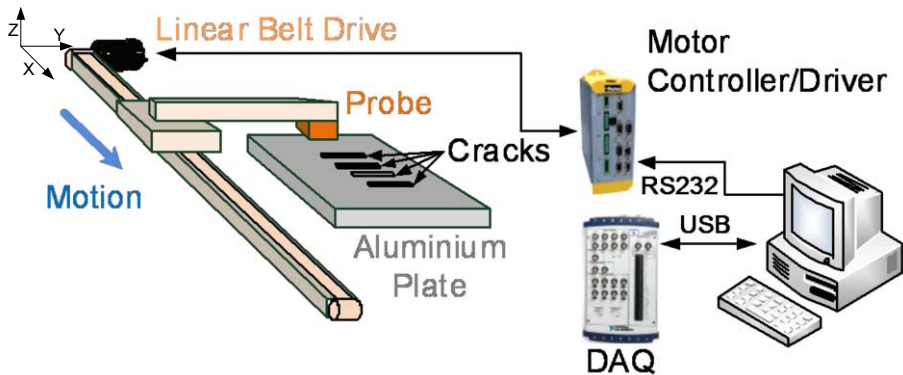


Fig. 1. Block diagram of the experimental setup.

The probe includes magnets to produce eddy currents and magnetic sensors to measure the magnetic field in the vicinity of the sample. Details about the experimental setup are described in [5].

3. Velocity Induced Eddy Current Method

In the velocity induced eddy current inspection method used in this work, it is a constant magnetic field produced by a permanent magnet that generates induced eddy currents, when the magnet moves at a constant speed over a metallic plate. This permanent magnet source is included in the probe represented in the setup depicted in Fig. 1.

The perturbations in the total magnetic field generated in the vicinity of the sample are assessed using Hall sensors also included in the probe.

3.1. Magnetic Field Source

The magnetic field generated by a coil with a constant current is hardly matched by the one that is generated by neodymium magnets (NdFeB). The remanence field property of a NdFeB magnet can be greater than 1 T, which makes them the strongest type of permanent magnets commercially available to date. For this reason this type of magnet was chosen to be implemented in the velocity probes developed within this work.

The shape and orientation of the permanent magnet is a key factor to induce the desired current distribution. The initial probe designed to perform velocity ECT inspection was based in previous experience that the sensitivity to defects is increased if all currents are perpendicular to the defect, because it is the case where most current perturbation occurs.

In [8] different moving magnetic probes were used to detect surface defects in electrically conductive metal. Two types of probe configurations were considered: (1) ‘Sleeping magnet probe’ which includes a single permanent magnet with its magnetization perpendicular to the motion axis and parallel to the surface under inspection; (2) ‘Standing magnet probe’ composed by two magnets having opposing magnetizations perpendicular to the surface under inspection. It was concluded [4] that the configuration (2) creates two smaller zones of uniform current density with the current direction perpendicular to the cracks when the probe moves along the sample. The value of eddy surface current density amplitude in these zones of uniform current is larger than the maximum current density amplitude obtained with the “sleeping magnet” simulation.

Since higher current density results in higher perturbations during the interaction with a crack, then the probe sensitivity to cracks is increased in ‘standing magnet probes’, due to the higher current densities that are induced by the two magnets. This configuration was generalized to multiple “standing magnets”, meaning that multiple magnets are placed next to each other with alternating magnetization. By placing each loop next to another loop with reversed current direction, a focused zone with the current along x can be obtained. Also it improves the current density by positioning the magnet vertically (z magnetization). The magnetic flux that penetrates into the metal is increased, thus increasing eddy current density.

After that, a probe with a matrix of permanent magnets, as the one schematically represented in Fig. 2.(a), that have alternating vertical magnetizations was developed and built. By placing the additional row of magnets, with opposing magnetization direction, next to the previous one, the derivative of the magnetic flux will increase as the probe moves above the metallic plate. A numerical simulation was made using Comsol software for the configuration represented in Fig.2 (a) containing two rows of alternating magnets. To constrain the span of the induced current loops (thus increasing the density of current eddies), cubic neodymium magnets with 5 mm of edge were used in both simulations and experimentally. These magnets were chosen not only because they are widely availability but also because the interstitial space between magnets can be used to position the Hall sensor. In this case, a 6 mm resolution is achievable (the spatial span along y axis of each loop is restrained to the length of the cube plus the 1 mm spacing of the Hall sensor, which equals to 6 mm). The top view of the probe’s finite element model moving at 4 m/s, with the outline of the six magnets and eddy current pattern are depicted in Fig. 2 (b).

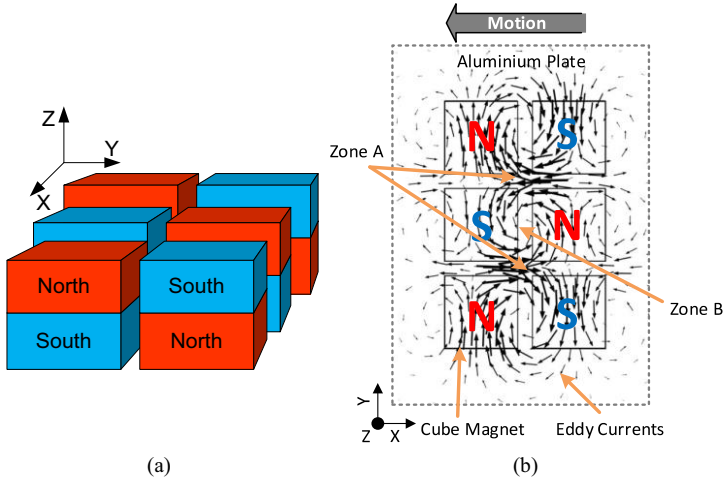


Fig. 2. (a) Magnet Probe assembly; (b) Eddy current distribution induced in the moving Al plate surface.

The maximum value of current density located in the surface in the single magnet model the maximum value obtained of eddy current density was 2.0×10^7 A/m², and it is located behind the magnet. In the improved probe this value was increased to 3.4×10^7 A/m² and the location is in the two zones A depicted in Fig. 2 (b).

3.2. Magnetic Sensor

[9] reports and analyses in detail data measured with three different magnetic sensors when a sample with machined cracks perpendicular to the eddy currents induced in the material are inspected using velocity induced eddy current method. The sensors used are a differential pick-up coil, a commercial magnetic sensor based on giant magneto-resistors (GMR) and a Hall sensor.

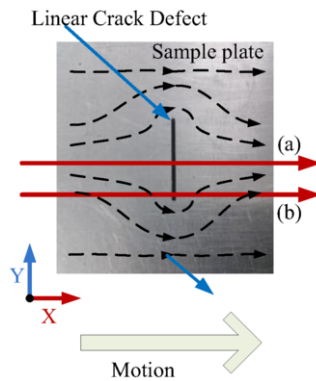


Fig. 3. Permanent magnet positioning and current flow perturbation due to the presence of a defect; paths covered by the magnetic sensors.

Due to the different nature, position and orientation of each of the magnetic sensors, a different optimal crossing point between the sensor scan path and the linear defect are obtained for each sensor, in order to maximize its detection sensitivity.

Fig. 3 shows a pictorial representation of the eddy current perturbations in the metallic sample when a linear crack is inspected if all the induced currents were considered parallel and in the direction of the motion. This is a very raw simplification, as these currents only temporarily appear in front and behind the permanent magnet as it moves above the plate. Nevertheless, it allows drawing conclusions about the best scanning path of each sensor in order to increase defect sensitivity.

The two paths along the movement direction, (a) and (b), will be travelled by the probe to obtain the magnetic field values that may inform about the existence of the crack. Thus, one may expect that when crossing the defect along the path (a) (in the middle of the crack), if the magnetic sensor sensing axis is in the y-direction, then the magnetic field has a minimum due to the conductivity discontinuity caused by the crack. For the same reason, if the magnetic component to be detected is along x then the path to be traversed must be (b) and two peaks (one negative and another positive) are observed. Taking these observations into account and because each magnetic sensor measures a different component of the magnetic field density, in [5] it was concluded that Hall effect sensor is more advantageous than either the GMR sensor or the differential coil. In fact, the GMR sensor can only detect the defect if the measurement is taken along the path crossing the edge of the defect and its polarization is rather difficult due to the magnet vicinity. In the cases of the pick-up coil and Hall sensor, the best crossing point between the scanning path and the defect should be in the center of the defect (path (a)). As these sensors are sensitive to x oriented currents, the maximum perturbations in the x oriented eddy currents occur in the intersection point between path (a) and the defect. The waveform for the Hall sensor contains one peak and the pick-up coil probe contains a derivative of the Hall sensor signal. The Hall sensor measures magnetic flux density and differential coils its derivative, therefore, more difficult to interpret.

3.3. Probe Design

The probe used to inspect the 4 mm thick 1050 aluminium plate with four machined defects using velocity induced eddy currents includes an array of 3x2 cubic neodymium magnets with 5 mm of edge to create the source magnetic field and Hall effect sensors to measure the resultant magnetic density field. All defects are 50 mm long, 0.5 mm wide and depth varies from 0.5 mm to 2 mm with intervals of 0.5 mm.

The placement of each Hall effect magnetic sensor is of great importance, as it must be able to sense the magnetic field perturbations caused when crossing a defect. From Fig. 2, one concludes which is best component of the magnetic field to sense current perturbations can be predicted. B_x is not a good choice, as the only possible location for the sensor is between the two magnets in zone B, where the contribution to B_x is self-nulling due to the curling nature of the induced currents. Also from Figs. 2 and 3, one can expect that measuring B_y in zone A and B_z in zone B will provide good results. In zone A the currents are oriented along x axis, which results in a B_y perturbation when they are interrupted. To be noted that the perturbations measured in zone B are mainly caused by interruptions to the loop of currents centered in zone B, however the perturbations in a zone A will be a contribution from both loops of current that crosses this zone.

4. Experimental Results

To validate the simulated results, an experimental setup was assembled with two probes, where each one differs on the placement of the magnetic sensor in zones A and B measuring B_y and B_z respectively. Two probes were created with the array of magnets with opposing magnetizations to maximize the induction of eddy currents, and they differ on the location of the magnetic sensor. The Hall sensor was placed in zones A and B as depicted in Fig. 2. The probe with the sensor in zone A is named Probe A and similarly Probe B is the one with the sensor above zone B. Fig. 4 depicts the profile of the two

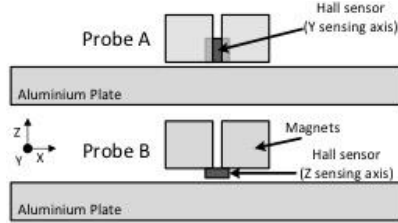


Fig. 4. Experimental results: (a) Representation of the two assembled probes.

To validate the simulated results, two probes were built with the array magnets with opposing magnetizations and with the Hall sensor placed in zones A (Probe A) and B (Probe B) as depicted in Fig. 4. Using the experimental setup shown in Fig. 1 the sample of aluminium was scanned.

To obtain the best possible experimental results, the lift-off should be minimized. Despite the 1 mm lift-off used in the simulation model, the lift-offs used experimentally differ for each probe as observed in Fig. 4. Probe A has the Hall sensor placed in the interstitial space between four magnets, so the magnets and Hall sensor have approximately 0.2 mm of lift-off. However, Probe B has the sensor laid flat between the plate and the magnets, which increases the distance between magnets and the surface of the aluminium plate to about 1.2 mm (1 mm due to the thickness of the Hall sensor and approximately 0.2 mm between sensor and aluminium).

The first experimental result was obtained by scanning the defects in the aluminium plate with both probes A and B. The probes scanning paths crossed the middle of the 50 mm long defects with 0.5, 1.0, 1.5 and 2.0 mm of depth at a fixed speed of 4 m/s. Figure 5 depicts the Hall sensor voltages acquired when scanning the aluminium plate.

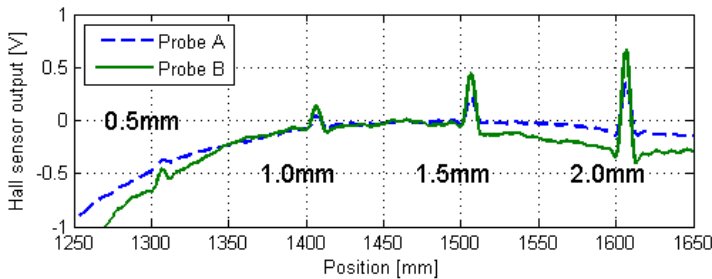


Fig. 5. Experimental results: (c) Differential output voltage from probe A and Probe B.

The characteristic waveform of a defect is composed by a positive peak preceded and followed by two smaller negative peaks. The positive peak is caused by the defect interrupting the loop centered on the zone B in Fig. 2. The other two smaller peaks appear

when the defect interrupts each loop in front and behind zone B. When this occurs, some current will switch from flowing in the perturbed loop to flow in to the zone B loop. Some noise due to slight variations in the lift-off distance can also be observed in Fig. 5. To remove this noise, the output voltage from the Hall sensor was differentiated in MATLAB and the obtained result is depicted in Fig. 6.

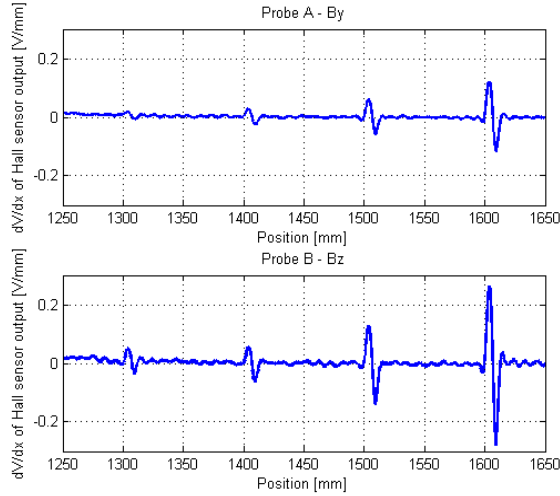


Fig. 6. Experimental results: (c) Differential output voltage from probe A and Probe B.

Observing the obtained signals in Fig. 6, it is possible to identify the four defects with both probes.

Despite having a larger lift-off than probe A, probe B shows approximately twice the signal amplitude in all defects, so it can be concluded that Probe B has greater sensitivity to defects.

5. Conclusions

The results obtained with the probe with two rows of permanent magnets showed that it is possible to constraint the loops of induced currents and thus increase their density, and in turn, the sensitivity to defects. The optimal location of the magnetic sensor was verified experimentally. This probe design was based on all the insights gained with the previous probe designs. It showed the best results and highest sensitivity out of all the probes previously studied and was able to detect a 0.5 mm deep superficial defect, which is a complex task to do even with conventional ECT.

The chosen design is capable of being scaled and also arrayed, which makes it versatile. Also, it is possible to detect defects with any orientation by using two probes, one in front of the other, and shifted by half the size of the cubic permanent magnet edge. The study made for aluminium specimens was successfully made, and the developed probe can be used to perform inspections in conductive non-ferromagnetic materials.

Acknowledgement

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Fast Simulation Tools Dedicated to Pulsed Eddy Current Applications

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Abstract. This paper presents an efficient strategy to simulate pulsed eddy current testing signals. It relies on adaptive sparse grids interpolation for the fast evaluation of Fourier harmonics composing the transient pulsed eddy current signal. In addition to the discussion on the performance obtained, some useful post-processing procedures enabling to retrieve relevant information from raw signals are described.

1. Introduction

Pulsed Eddy Current Testing (PECT) is a popular nondestructive testing (NDT) technique for some applications like corrosion monitoring in the oil and gas industry, or rivet inspection in aeronautics. The transient nature of the excitation signal enables to retrieve much more information about the material than the harmonic one commonly used in conventional ECT. This communication presents a complete modelling approach adopted at CEA for the fast simulation of PECT inspection, making use of existing ECT models implemented in harmonic regime [1] and exploiting the so-called Frequency Domain Summation (FDS) method [2] to derive time signals. In the classical FDS implementation, the excitation signal feeding the emitting part of the probe is decomposed in Fourier harmonics, then for the whole set of harmonic components a time-harmonic solver is evaluated and, finally, the time output signal is generated through an inverse Fourier transform. This strategy avoids the implementation of a complete time domain formalism [5] but depends on a loop over the frequency content of the excitation signal. therefore, it is efficient when dealing with narrow band excitations. However, it can turn to be cumbersome when dealing with broadband ones, which are often used in typical PECT applications. Improved FDS strategies have already been proposed to mitigate the computational burden by selecting a subset of representative harmonics within the whole spectra [2,3,4]. Following the path explored in [4], this work proposes to model an accurate representation of ECT signals over a wide frequency range by sampling it in a sparse and unsupervised manner. To this end, a mono-dimensional metamodel is constructed with respect to frequency using an Adaptive Sparse Grid (ASG) algorithm. Approximation over the complete frequency band is obtained through incremental interpolation [6].

The paper is structured as follows. In Section 2 an overview on the FDS strategy is presented. In Section 3, an introduction to hierarchical ASG algorithm is provided along

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with its application to the generation of PECT signals. In Section 4, the numerical results obtained by applying the proposed strategy will be presented and discussed. At the end of the paper conclusions and further perspectives of this work are highlighted.

2. Generation of PECT signals based on frequency domain summation

In ECT signal simulation, time harmonic semi-analytical solvers have shown to be a very effective and accurate simulation tool when canonical geometries are addressed [1]. This feature is highly desirable when one wants to model PECT signals based on the use of time-harmonic solver and Fourier transform. This approach, known under the name of frequency domain summation strategy, has been recently proposed in [2] for Finite Element Method (FEM) time harmonic solver and in [3] for the semi-analytical counterpart.

2.1. Calculation of ECT signal in time harmonic regime

Let us consider to work at a given frequency f with a time-harmonic dependance $e^{-i\omega t}$ (conveniently omitted in our notation), therefore the coil impedance variation due to the presence of S discontinuities within the inspected non-magnetic media can be written as for a given coil position as [7]

$$\Delta Z = \frac{1}{I^2} \sum_{s=1}^S \int_{V_s} \frac{1}{\sigma_s} \mathbf{J}_s^{inc}(\mathbf{r}) \cdot \mathbf{p}_s(\mathbf{r}) d\mathbf{r}, \quad \mathbf{r} \in V_s, \quad s \in 1, S.$$

where V_s is the volume of the s -th flaw present in the piece. $\mathbf{J}_s^{inc}$ is the current density induced by the coil in the region V_s without flaw and $\mathbf{p}_s(\mathbf{r})$ is the fictitious current dipole density representing the s -th flaw contribution. I is the coil driven current and σ_s is the conductivity associated to the layer where the s -th flaw is located. More details on the calculation of $\mathbf{p}_s(\mathbf{r})$ obtained with semi-analytical approaches can be found in [8,9,10]. In PECT, like in ECT, the crack signal is generally a tension across the receiving part of a probe, consisting of a set of coils or magnetic field sensors (like Hall effect, Giant Magneto Resistance (GMR) for instance). A popular design consists in placing a magnetic sensor just under the excitation coil.

2.2. Frequency domain summation schema

In PECT a periodic pulse wave excitation signal (Fig. 1(a)) is applied to the driven coil. The associated spectral components can be retrieved via the discrete Fourier transform (Fig. 1(b)) and the periodic signal can be decomposed as a sum of the k -th sinusoidal waves having different frequencies (f_k) and amplitudes ($\tilde{F}_k$) such that

$$curr_{inj}(t) = \sum_{k=1}^K \tilde{F}_k \exp(-i \cdot 2\pi f_k t), \quad (1)$$

Supposing to deal with linear media, one can retrieve the PECT response signal (S_{PECT}) by summing the response at every frequency weighted by the associated amplitude coefficient as [2]

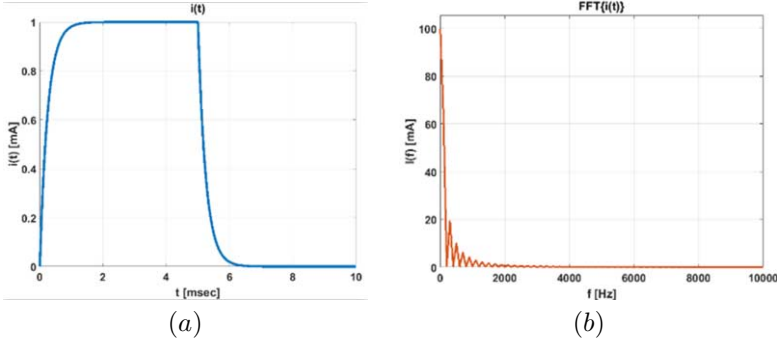


Figure 1. In (a), a typical PECT excitation signal is provided along with the associated spectra shown in (b).

$$S_{PECT}(t) = \sum_{k=1}^K \tilde{F}_k H(f_k) \exp(-i \cdot 2\pi f_k t), \quad (2)$$

where $H(f_k)$ represents the ECT output signal (i.e., the crack response) at the frequency f_k . Equations (1) and (2) can be computed efficiently through the Fast Fourier Transform (FFT) and Inverse FFT (IFFT) algorithms, respectively. From the computational point of view, the FDS strategy can easily turn to be a cumbersome approach. In fact, the efficiency depends on the number of frequencies to be simulated and/or the computational time associated to the numerical solver. This means that the calculation of $H(f_k)$ may represent the major bottle neck in terms of performance. Therefore it becomes crucial to minimize as much as possible the number of calls to the forward solver (i.e., the value of K). A way to handle this issue consists in calculating only a subset K^* harmonic responses $H(f_{1,\dots,K^*})$ and to interpolate on the remaining $K - K^*$ frequencies. This kind of strategy have been applied recently in different works with good results [2,3,4]. It is proposed in [2,3] to choose a set of log-spaced frequencies, from which the complete spectrum is interpolated. In [4] the choice of the frequencies was based on a dedicated adaptive algorithm based on kriging interpolator. For both strategies the accuracy on the interpolation results can be assessed *a-posteriori*. In this work, we shall employ an ASG interpolation strategy choosing adaptively the set of frequencies (e.g., sample the associated PECT signal band). Furthermore, it allows to estimate the interpolation accuracy during the sampling process. It is worth mentioning that another advantage of the sparse grid algorithm is its efficiency to manage high dimensional spaces, even if we do not exploit it here.

3. Adaptive sparse grid interpolation applied to fast generation of PECT signals

In the last two decades, the technique of Sparse Grid (SG) has been applied with success in the framework of numerical integration, resolution of partial differential equation and, more recently, in the context of metamodel generation. The use of sparse grid is motivated by the fact that it allows to partially overcome the curse of dimensionality that plagues the classical full-grid discretization[12]. The SG method is based on the Smolyak algorithm [11] which relies to the idea that only some elements of the full tensor grid are useful to interpolate the ground truth function. The set of points are known

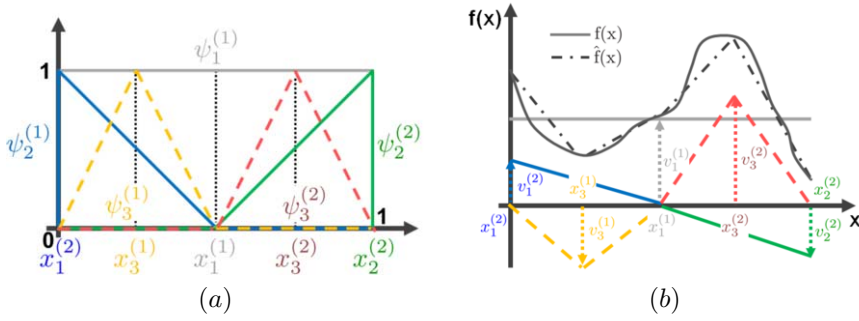


Figure 2. Schematic view of hierarchical sparse grid interpolation. In (a), hierarchical basis function $\psi_l^{(i)}(x_l^{(i)})$ with the associated support nodes $x_l^{(i)}$ until $l = 3$ are shown. In (b), hierarchical basis function interpolation process is given for a one dimensional toy-function, the hierarchical basis are employed until the third level.

a-priori and are arranged on a level-wise grid. Two main classes of basis (interpolation) functions are adopted in the literature. The first one is based on global polynomial function which are smooth and continuously differentiable. The second class, in its basic implementation, employs hierarchical surpluses and piecewise linear local functions leading to non-smooth and non-differentiable approximation (other hierarchical basis can be applied though [12,13]). In this work, we employ a hierarchical ASG [6,12] aiming at reducing the computational burden through a parsimonious selection of the SG points to be simulated.

3.1. Hierarchical adaptive sparse grid methods

Let us suppose to deal with a costly-to-evaluate function $f(\mathbf{x}) \in \mathbb{C}$ of a vector variable $\mathbf{x} \in \Omega \mid \Omega \in \mathbb{R}^{1 \times N}$ for which we want to find a suitable cheap-to-evaluate sparse grid interpolation $\hat{f}(\mathbf{x}) : \hat{f}(\mathbf{x}) \approx f(\mathbf{x}), \forall \mathbf{x} \in \Omega$. Thus, we define the hierarchical structure of basis function employed starting from the lower level ($l = 1$) up to the highest level $l = D$ for an one dimensional problem (i.e., $N = 1$). The hierarchical structure is defined through a set of level-wise basis functions $\psi_l^{(i)}(x)$ with $x \in [0, 1]$. To each basis function is associated a supported node(s) $x_l^{(i)}$ defined through a recursive nested algorithm. The evaluation of the basis function onto the collocation node(s) leads to $\psi_l^{(i)}(x_l^{(i)}) = \delta_{ij}$ where δ_{ij} is the Kronecker symbol. In this paper, we will employ a Newton-Cotes grid (sometimes known as Clenshaw-Curtis) with equidistant nodes and linear basis function [6,12]. The one-dimensional set of collocation points and the associated linear basis is given for the first three levels in Fig. 2(a). Starting from the level 3 basis functions are all of the same kind, whereas level 1 and level 2 basis behave differently. The sparse grid interpolant at the depth d is defined as

$$\hat{f}_d(x) = \hat{f}_{d-1}(x) + \sum_{i=1}^{n_d} \psi_d^{(i)}(x) v_d^{(i)}, \quad (3)$$

where n_d represents the number of basis functions at level d . $v_d^{(i)}$ is the so-called hierarchical surplus for the i -th node and it is defined as $v_d^{(i)} = [f(x_d^{(i)}) - \hat{f}_{d-1}(x_d^{(i)})]$. Look-

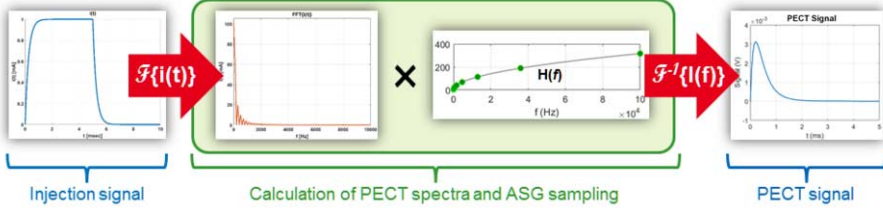


Figure 3. Simulation of PECT signal through FDS method with adaptive sparse grid.

ing at equation (3), we notice that the hierarchical surplus “weights” the contribution of the i -th basis $\psi_d^{(i)}(x)$ function on the interpolation process. This process is schematically shown in Fig. 2(b). Considering the definition of hierarchical surpluses, we can observe that it embeds the predicted variation of the accuracy interpolation between two adjacent sparse grid levels. This information can be seen as a measure of the interpolation accuracy and a kind of measure of the degree of smoothness of the underlying function landscape. Therefore, surpluses are exploited in order to adaptively refine the hierarchical sparse grid to obtain an even more “sparse” database with a criterion that considers both error and the tolerance associated. We define the estimation of the interpolation error as $\varepsilon_d(x) = \left\| \hat{f}_d(x) - f(x) \right\|_2$, where $\|\cdot\|_2$ stands for the L_2 norm. By comparing the above equation with the definition of hierarchical surplus, we have that the error at the depth $d-1$ (i.e., $\varepsilon_{d-1}(x)$), is equal to magnitude of the surplus $v_d^{(i)}$, thus it can be considered as local estimation of the error at the depth d . This adaptive refinement strategy has been applied and extended to N -dimensional case with success on electromagnetic problems, for which accurate databases and metamodels with many parameters have been generated and tested [6].

3.2. ASG algorithm applied to the generation of PECT signals

In this work, the simulated adaptive sparse grid samples, which constitute the system response $H(f)$, are represented by the coil impedance variation $\Delta Z(f)$ or the magnetic flux density $\mathbf{B}(f)$, where f stands for the set of simulated harmonics. Once the adaptive sampling strategy ended, the ASG interpolator is evaluated by employing equation (3) on the set of frequencies belonging to the $FFT\{curr_{inj}(t)\}$. Subsequently, the PECT signal is calculated through IFFT algorithm through $\hat{S}_{PECT}(t) = IFFT\{\hat{S}_{ECT}(f)\} = \sum_{n=1}^N \tilde{F}_n \hat{H}(f_n) \exp(i \cdot 2\pi f_n t)$, where the symbol “ $\wedge$ ” denotes the approximated values computed via the ASG interpolation. The PECT signal generation process is shown schematically in Fig. 3. An interesting point is the capacity of this method to generate directly PECT signals for many probe positions. This feature can be of great interest when a semi-analytical solver [4] is employed, since the computational time is almost independent to the number of coil positions.

4. Numerical results

Two configurations of interest in NdT community have been considered: a problem of narrow crack(s) detection in a planar medium and the detection of a corroded area in a

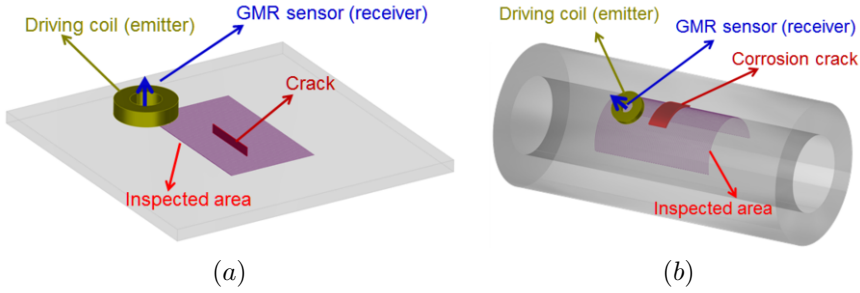


Figure 4. Considered PECT signal modeling test cases studied tube geometry with a PECT probe inspecting the pipe from the outside.

large pipe. Considering Fig. 4 (a), a planar medium of 4.0 mm thickness having conductivity equal to 18.76 MS/m is inspected with a probe made by an emitting coil having 10.0 mm and 20.0 mm as inner and outer coil, respectively. The coil high is equal to 5.0 mm, 625 turns and the its lift-off is 1.0 mm. The sensing probe is a Giant Magnetic Resistance (GMR) sensor located 0.5 mm beneath the driven coil and able to sense the z -component of the magnetic flux density $\mathbf{B}$. The inspected plate contained a slot having length, depth and width equal to 20.0 mm, 3.0 mm and 0.2 mm, respectively. The injected signal into the driven coil is a periodic pulse having period equal to 10.0 msec, duty-cycle of 50% and time constant τ equal to 250 msec. In Fig. 5 (a), we compared the results obtained with the proposed approach with the ones simulated with Finite Element Method (FEM) solver (i.e, Comsol v.5.2a) for which the time stepping approach has been adopted. We can notice that overall the results agree well on a meaningful and large part of the PECT signal. In Fig. 5 (b), we have shown how the extracted features corresponding to maximum of B_z varies over all the coil positions. Looking at this feature, we can notice that this kind of information can be quite appealing to be employed to crack localization and/or characterization since the signal variation appears clearly on the map. From the computational point of view, the ASG-based approach for generating PECT signal has shown good computational performance. In fact, by interpolating the signal spectrum only 35 frequencies were generated to obtain the final signal out of 100 frequencies of the complete spectrum. Furthermore, comparing to FEM approach, the whole map of PECT signals has been obtained, on a standard laptop, in about fifteen minutes instead of the ten minutes required by FEM solver for just one coil position.

The second problem addressed this paper deals with the issue of corrosion crack(s) detection in large pipes. As sketched in Fig. 4 (b) the inspection procedure is performed with a probe which is scanning the tube outer surface whereas the corrosion laying on the inner surface. The driven element considered is a coil having inner and outer radius equal to 10.0 mm and 25.0 mm, respectively. The coil height is 10.0 mm, the number of winding is equal to 5000 and the lift-off is equal to 1.0 mm. The response signal is picked-up by a GMR located 0.5 mm beneath the driven coil. The GMR sensor is sensitive to the r -component of the magnetic flux density (B_r). The probe (driven coil and GMR sensor) is moving in a zone made by 71×51 measured points centered over the crack region. The considered specimen is a thick pipe having inner radius equal to 60.0 mm and a thickness of 50.0 mm. The specimen conductivity is 1.0 MS/m and the relative permeability is equal to 100. The considered corrosion crack is modelled like a flat slot having length, depth and width equal to 73.3 mm, 30.0 mm and 3.0 mm, respec-

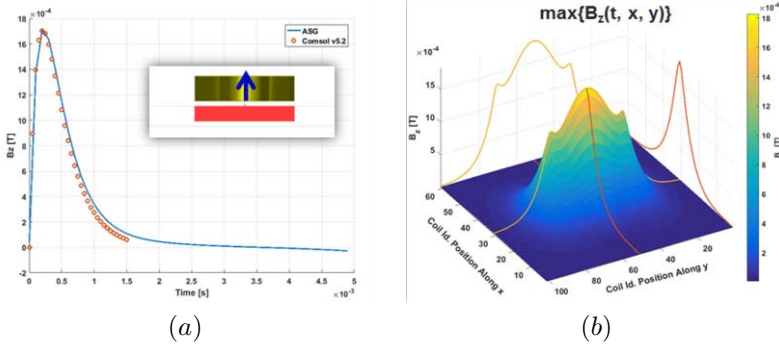


Figure 5. Considered PECT signal modeling of a plate affected by a slot. In (a), comparisons between FEM results and ASG approach are shown for the coil centered above the crack. In (b), we shown the extracted features corresponding to the maximum over the time of B_z for each coil position of the inspection map.

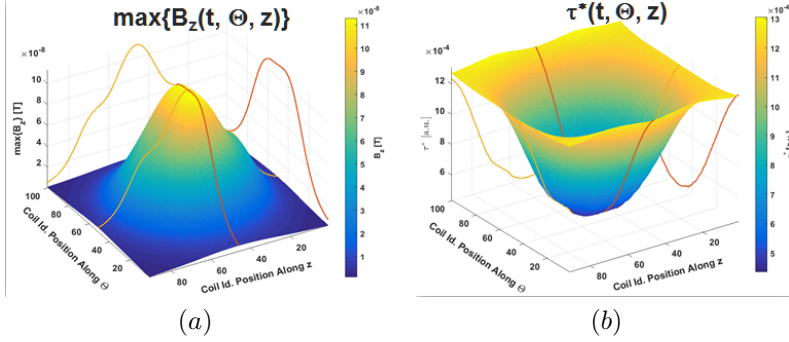


Figure 6. Considered PECT signal modelling of a tube affected by a corrosion crack. In (a), the maximum value of B_z over the time for each coil position of the inspection map is shown. In (b), again for each coil position, we shown the feature extraction associated to the coefficient τ^* .

tively. The injected signal into the driven coil is a periodic pulse having period equal to 10.0 msec, duty-cycle of 50% and time constant τ equal to 0.25 msec. Detection and characterization of corrosion crack(s) is often performed by considering derived quantities instead of the full time-domain PECT signal. The main two features used are the maximum amplitude of the PECT signal for each coil position (e.g., see Fig. 6 (a)) and the decay coefficient τ^* associated to an exponential decay fit on the PECT tail once injected signal is switched off. One can see in Fig. 6 (b) how τ^* behaves with respect to the probe scan. The database was generated based on the simulation of 9 frequencies instead of the 100 of the complete spectrum. Such a few number of simulated frequencies is due to the “smooth” frequency response $H(f)$ obtained for the studied pipe problem. In fact, the degree of “smoothness” is due to the fact that the problem involves a thick ferromagnetic tube having a corrosion crack lying on the inner surface while a the inspecting probe is placed on the outer surface. Therefore, eddy current decays rapidly nearby the crack when frequency increases.

5. Conclusion and perspectives

This paper proposes an efficient strategy for modelling PECT signals based on the FDS approach. In particular, the ASG algorithm has been employed to sampling, in an unsupervised way and successively interpolating the frequency spectrum associated to a pulse excitation signal. Results presented and discussed have shown that the implemented approach is a convenient alternative to other techniques already proposed in the literature [2,3,4]. The possibility to generate fast PECT signals paved the way for the generation and the exploitation of databases [6,16] for computationally expensive studies like model assisted probability of detection (MAPOD), sensitivity analysis and almost real-time parametric inversion.

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A Level Set Method for Magnetic Induction Tomography of 3D Boxes and Containers

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Abstract We present a level set technique for 3D Magnetic Induction Tomography with an emphasis on applications to the screening of small boxes up to cargo containers. A level set method is used for modeling a shape evolution when minimizing a given cost functional. Numerical results will be presented that illustrate the performance of our method in practical situations. A novel line-search technique is introduced that is suitable to control the shape evolution for this computationally demanding MIT inverse problem.

Keywords. Magnetic induction tomography; Shape evolution; Level sets; Nonlinear Kaczmarz; Line search;

1. Introduction

Magnetic Induction Tomography (MIT) has found various applications in the nondestructive testing of materials, see for example [1,4,10] and the references cited there. Recent progress in measurement and detector technology is opening up exciting new applications for this technology, for example in the security screening of small boxes up to cargo containers [6,11]. Mathematically 3D MIT constitutes a highly ill-posed nonlinear inverse problem where appropriate regularization and efficient reconstruction algorithms are essential when facing realistic scenarios. Linearized models have been tested (see for example [10]) but, even though being relatively fast, they quickly reach their limitations regarding resolution and accuracy. Nonlinear schemes seem more promising, even though they still are quite time-consuming and heavy on computer memory consumption. Tailor-made approaches for specific reconstruction, identification and classification tasks are needed for circumventing some of these difficulties. As part of this effort, shape-based techniques for identifying and characterizing hidden objects are developed here that make use of available prior knowledge or assumptions on the materials present in the region of interest. Mathematically this approach has been investigated in [1,7], amongst others. In our contribution we will follow up on this approach and investigate shape based methods for MIT of boxes or containers using a level-set framework [3].

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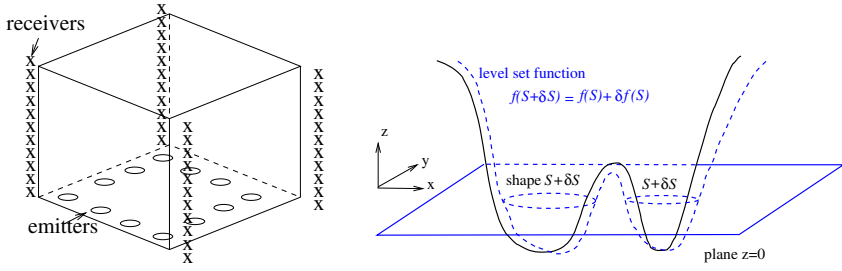


Figure 1. Left: geometric setup of emitters and receivers (here all horizontal coils) around a box-shaped region of interest. Right: cartoon in 2D showing evolution of a level set function for shape propagation.

2. The shape reconstruction problem

2.1. The forward problem

For modelling electromagnetic fields inside the box of interest we use the system of Maxwell's equations in frequency domain

$$\nabla \times \mathbf{E}_j(\mathbf{x}) - a(\mathbf{x})\mathbf{H}_j(\mathbf{x}) = \mathbf{M}_j(\mathbf{x}) \quad (1)$$

$$\nabla \times \mathbf{H}_j(\mathbf{x}) - b(\mathbf{x})\mathbf{E}_j(\mathbf{x}) = \mathbf{J}_j(\mathbf{x}) \quad (2)$$

in a box-shaped domain $\Omega \subset \mathbb{R}^3$, where we imply a time-dependence $e^{-i\omega t}$ for a given frequency $\omega = 2\pi f$. In this study we will assume that the parameter $a(\mathbf{x}) = i\omega\mu(\mathbf{x})$ is known and constant with typical (e.g. free space) values in real and imaginary part. Furthermore, the parameter of interest $b(\mathbf{x}) = \sigma(\mathbf{x}) - i\omega\epsilon(\mathbf{x})$ is 'shape-based' in the sense that it chooses at each location $\mathbf{x}$ in the domain between either an 'internal value' $b_i(\mathbf{x})$ or an 'external value' $b_e(\mathbf{x})$, both being given as known and fixed functions of $\mathbf{x}$. This way $b(\mathbf{x})$ contains discontinuities along closed interfaces $\Gamma \subset \Omega$ that become now the unknowns of our inverse problem. For example, b_i might indicate some 'threat' and b_e some background material of which the specific nature is of no particular interest.

The index j in (1), (2) indicates the different applied source patterns $\mathbf{q}_j = (\mathbf{M}_j, \mathbf{J}_j)^T$, $j = 1, \dots, p$, for creating the electromagnetic fields. Generalizations of this basic assumption will be addressed in our future research.

In the inverse problem of MIT one is interested in estimating the unknown parameter b from experimentally measured (or simulated noisy) data. In our case, this reduces to the estimation of the interface Γ between internal and external profiles. Practically, we introduce a sufficiently smooth level set function $\phi : \Omega \rightarrow \mathbb{R}$ which defines the shape S by

$$b(\mathbf{x}) = \Pi(\phi)(\mathbf{x}) = \begin{cases} b_i(\mathbf{x}) & \text{in } S \text{ where } \phi(\mathbf{x}) \leq 0, \\ b_e(\mathbf{x}) & \text{in } \Omega \setminus S \text{ where } \phi(\mathbf{x}) > 0. \end{cases} \quad (3)$$

The interface $\Gamma = \partial S$ is given by the zero level set $\partial S = \{x \in \Omega : \phi(x) = 0\}$. See the right image of Figure 1 for a visualization of the level set scheme. In our (proof-of-concept style) setup we assume for simplicity that we have p different sources $\mathbf{q}_j$, $j = 1, \dots, p$, modelled by reasonably sized wire loops or coils (see the left image of Figure 1) located at the bottom of the box or container. Those give rise to the probing fields $\mathbf{E}_j$ and $\mathbf{H}_j$. The

calculated data are $g_j(b) = \mathcal{M}_j \mathbf{E}_j$ where $\mathbf{E}_j, \mathbf{H}_j$ solve (1), (2) with parameter b and where $\mathcal{M}_j$ is a (linear) measurement operator which might depend on the source position. In our case, $\mathcal{M}_j \mathbf{E}_j$ are the line integrals of the electric fields along receiving closed wire loops or coils distributed along the four vertical edges of the box (see again Figure 1). We denote the actually measured (or 'true') data by $\tilde{g}_j$. Then, the residuals $\mathcal{R}_j(b)$ and $\mathcal{T}_j(\phi)$ are defined as $\mathcal{R}_j(b) = g_j(b) - \tilde{g}_j$ and $\mathcal{T}_j(\phi) = \mathcal{R}_j(\Pi(\phi))$.

2.2. Gradient calculations by an adjoint scheme

Defining the least squares cost functionals

$$\mathcal{J}_j(b) = \frac{1}{2} \|\mathcal{R}_j(b)\|_2^2, \quad \text{and} \quad \hat{\mathcal{J}}_j(\phi) = \frac{1}{2} \|\mathcal{T}_j(\phi)\|_2^2 \quad (4)$$

and the total cost by

$$\mathcal{J}(b) = \sum_{j=1}^p \mathcal{J}_j(b) \quad \text{and} \quad \hat{\mathcal{J}}(b) = \sum_{j=1}^p \hat{\mathcal{J}}_j(b) \quad (5)$$

we call $\mathbf{grad}_{\mathcal{J}_j, L_2}(b) = \mathcal{R}'_j(b)^* \mathcal{R}_j(b)$ and $\mathbf{grad}_{\hat{\mathcal{J}}_j, L_2}(\phi) = \mathcal{T}'_j(\phi)^* \mathcal{T}_j(\phi)$ the *gradient directions* of expressions in (4) and similar for those in (5). Notice that these gradient directions will depend on the choice of function spaces for the parameter functions b and level set functions ϕ , as indicated in the notation. This will be used when selecting our regularization scheme for the inversion. We have the following *adjoint representation* of $\mathbf{grad}_{\mathcal{J}_j, L_2}(b)$ as shown for example in [2].

Let $\mathbf{E}_j^a$ and $\mathbf{H}_j^a$ be the solution of the following (adjoint) system of Maxwell's equations:

$$\nabla \times \mathbf{E}_j^a(\mathbf{x}) - a(\mathbf{x}) \mathbf{H}_j^a(\mathbf{x}) = 0 \quad (6)$$

$$\nabla \times \mathbf{H}_j^a(\mathbf{x}) - b(\mathbf{x}) \mathbf{E}_j^a(\mathbf{x}) = \mathcal{M}_j^* \overline{\mathcal{R}_j(b)} \quad (7)$$

where $\mathcal{M}_j^*$ denotes the formal adjoint of the measurement operator $\mathcal{M}_j$. Its application amounts to putting the argument (here $\overline{\mathcal{R}_j(b)}$) as artificial 'adjoint sources' at the receiver locations. The overline denotes 'complex conjugate'. Then we have

$$\left[\mathcal{R}'_j(b)^* \mathcal{R}_j(b) \right](\mathbf{x}) = \overline{\mathbf{E}_j(\mathbf{x})} \cdot \mathbf{E}_j^a(\mathbf{x}) \quad (8)$$

where $\mathbf{E}_j$ and $\mathbf{H}_j$ are the solution of (1), (2) with b .

Formal differentiation by the chain rule yields $\mathcal{T}'_j(\phi) = \mathcal{R}'_j(\Pi_j(\phi)) \Pi'_j(\phi)$. We have $\Pi'_j(\phi) = (b_i - b_e) \delta(\phi)$. However, the Dirac delta distribution $\delta(\phi)$ will be considered to be approximated by a suitable L_2 -function. In our numerical implementations, we will use the narrowband function of thickness d for that purpose. In other words, $\delta(\phi) \approx \chi_{B_d(\Gamma)}$ with $B_d(\Gamma) = \{\mathbf{x} : \text{dist}(\mathbf{x}, \Gamma) \leq d/2\}$ and χ_D denoting the characteristic function of the set D . We have $\mathcal{T}'_j(\phi)^* = \Pi'_j(\phi)^* \mathcal{R}'_j(\Pi(\phi))^*$.

Notice that so far no regularization scheme is applied in order to stabilize the inversion. We will use a regularization scheme explained in more details in [3] which uses suit-

able Sobolev spaces for the gradient calculation (sometimes called ‘Sobolev gradients’). This has the effect of replacing the adjoint operator $\mathcal{T}'(\phi)^*$ by a new adjoint operator $\mathcal{T}'_j(\phi)^\circ$ with

$$\mathcal{T}'_j(\phi)^\circ = (\alpha I - \beta \Delta)^{-1} \mathcal{T}'_j(\phi)^*, \quad \mathbf{grad}_{\mathcal{J}_j, W_1}(\phi) = \mathcal{T}'_j(\phi)^\circ \mathcal{T}_j(\phi) \quad (9)$$

and suitably chosen parameters α, β controlling the degree of regularization. The positive definite operator $(\alpha I - \beta \Delta)^{-1}$ has the effect of ‘projecting’ the gradient $\mathcal{T}'_j(\phi)^* \mathcal{T}_j(\phi)$ from $L_2(\Omega)$ towards a Sobolev space of smoother functions [3].

2.3. A Kaczmarz-type shape reconstruction algorithm

We have presented above gradient directions with respect to the individual cost functionals $\mathcal{J}_j(b)$ which only take into account the part of the complete data set which corresponds to source $\mathbf{q}_j$. The gradient direction which corresponds to the full data set would be a vector whose components are the individual gradient directions $\mathbf{grad}_{\mathcal{J}_j, W_1}(\phi)$. However, calculating this combined gradient vector in each step of an iterative reconstruction technique is quite expensive in the case of 3D Maxwell’s equations. Therefore, we have adopted in our inversion a single-step (‘Kaczmarz-type’) reconstruction scheme which follows the general single-step idea of the nonlinear Algebraic Reconstruction Technique (ART) as described in [8]. We perform so-called ‘sweeps’ over the source positions and only consider the information of one source at a time while calculating an update by an efficient adjoint scheme.

2.4. A novel line search for the Kaczmarz-type shape reconstruction

We have equipped our algorithm with a novel line search strategy which automatically adjusts step sizes at each sweep. First we need to mention that a strict line search as in standard gradient based schemes (using for example a backtracking scheme with a Wolfe or Armijo condition [9]) usually does not work well in a Kaczmarz type algorithm since in each step we are only calculating gradients with respect to one of the p subsets of the data (corresponding to one source position). Searching for a step size that *minimizes* (4) just for the chosen j is not only computationally expensive but also might eradicate some of the progress made when working on the equivalent expressions (4) in previous steps (addressing the other $p - 1$ data subsets). A way out is to adopt a step size criterion that guarantees a smooth overall evolution of the shape where each individual update necessarily needs to be sufficiently small in order to preserve previous updates from the other data subsets. In particular, with step sizes chosen to be small, any additional forward solves need to be avoided in the corresponding step size selection process in order not to increase further the already demanding overall computational cost.

Notice that the avoidance of additional forward solves makes it technically difficult to monitor the evolution of (5) accurately. We obtain an *approximation* to (5) by recording values of (4) in each individual step as a by-product of gradient calculation (without adding any forward calculation) and plugging those values into the summation (5) at the end of each sweep. Thereby all forward solves actually contribute to an update, and no forward solve is ‘wasted’ just for monitoring the evolution of (5). This also means that we do not *enforce* a reduction of the cost (4) (or (5)) in a given step.

Instead, in order to maintain control of convergence of our algorithm, we record the number N_j of voxels that change value (from $b_i(\mathbf{x})$ to $b_e(\mathbf{x})$ or vice versa) in each update. Our aim is to control this number by an inexpensive backtracking line search. Following general ideas of a standard backtracking line search, our goal is to define an interval $I_{target} = [N_{inf}, N_{sup}]$ of values such that $N_j \in I_{target}$ in each update. N_{inf} is chosen to guarantee that sufficient progress is made in each step, and N_{sup} is chosen to make sure that we are not ‘overshooting’ in the above described sense. We choose the step size γ_j (see the pseudo-code further below) by a traditional backtracking line search, starting with γ_{start} sufficiently large, where however no forward solves are required. For each of the probed γ we neither have to calculate (4) nor (5), but only the corresponding N_j , which is done ‘on the fly’.

Implementing this backtracking line search is not difficult, but there remains the question of how to choose γ_{start} . Maybe surprisingly, in our experience the performance of the line search in this particular application of 3D MIT actually depends critically on a clever and dynamic choice of γ_{start} in each sweep! If it is chosen fixed and very large then the backtracking scheme will always stop just below N_{sup} , giving low sensitivity source-sensor pairs the same impact on shape evolution as high-sensitivity pairs. This can lead to a diverging algorithm overall. If it is chosen fixed and small, then the shape evolution might become either extremely slow at certain stages of the algorithm or again (choosing now always values N_j just above N_{inf}) insensitive to the actual sensitivity pattern of the experimental setup.

Therefore we have introduced a dynamic correction of γ_{start} after each sweep, enforcing that the *average* number of voxels $\bar{N} = (\sum_{j=1}^p N_j)/p$ that change value per step in each sweep remains inside a smaller interval $[N_{low}, N_{high}]$ with $N_{inf} < N_{low} < N_{high} < N_{sup}$. If $\bar{N} > N_{high}$ in some sweep, γ_{start} is reduced by multiplying it with a value $\tau_1 \in [0.5, 1.0[$. If, on the other hand, $\bar{N} < N_{low}$ in some sweep, then γ_{start} is increased by multiplying it with a value $\tau_2 \in]1.0, 2.0]$. This way we typically obtain a good spread of N_j in each sweep over the entire interval $[N_{inf}, N_{sup}]$ (after a short ‘burn-in phase’), making sure that more weight is given to high sensitivity source-receiver pairs (with respect to the current shape boundary) than for low sensitivity source-receiver pairs.

A pseudo-code of the algorithm including the line search is presented in the table.

3. Numerical experiments

We will show two numerical experiments which demonstrate the performance of the general shape-based 3D reconstruction algorithm as well as the novel line search technique proposed in this paper.

The general experimental and computational setup is in both cases the same as displayed in Figure 1. However, the two experiments differ in the dimensions and probing frequency chosen for the screening of the boxes. In Experiment 1 we aim at the screening of small to medium sized boxes where the size of the domain shown in Figure 1 is $2[\text{m}] \times 2[\text{m}] \times 2[\text{m}]$ and the frequency is $f = 10$ MHz. In Experiment 2 we want to screen bigger sized boxes such as cargo containers, where the domain of interest is chosen to be $10[\text{m}] \times 10[\text{m}] \times 10[\text{m}]$ and the frequency is $f = 0.2$ MHz. Notice that the computational domain is actually bigger than the domain of interest shown in Figure 1 since it is surrounded by some outside material (e.g. air). Both, b_e and b_i use constant free space

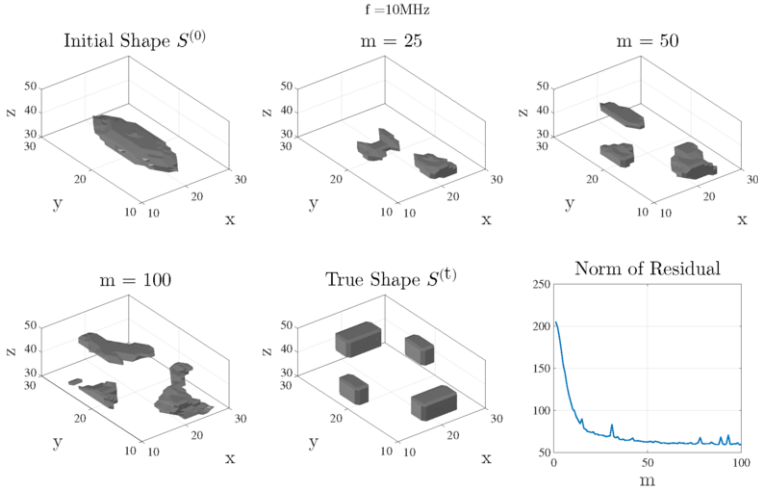
Algorithm: Single step level set shape reconstruction technique*Initialization:*Specify initial level set function as a signed distance function $\phi^{(0)}$;Choose $[N_{inf}, N_{sup}]$, $[N_{low}, N_{high}]$, γ_{start} , τ_1 , τ_2 and put $n = 0$.*Reconstruction:*FOR $m = 1 : M$ (loop over sweeps)FOR $j = 1 : p$ (loop over source positions)Calculate $\mathbf{grad}_{\mathcal{J}_j, W_1}(\phi^{(n)}) = \mathcal{T}'_j(\phi^{(n)}) \circ \mathcal{T}_j(\phi^{(n)})$ by (6)-(9)Perform backtracking line search for step-size γ_j Update level set function: $\phi^{(n+1)} = \phi^{(n)} - \gamma_j \mathbf{grad}_{\mathcal{J}_j, W_1}(\phi^{(n)})$ Rescale $\phi^{(n+1)} \rightarrow \zeta \phi^{(n+1)}$ with scaling parameter $\zeta \in \mathbf{R}^+$.

END

Verify optional stopping criterion.

Adjust, if necessary, γ_{start} for next sweep.

END

**Figure 2.** Shape and cost evolution vs sweep number in Numerical Experiment 1 (screening of small to medium sized boxes).

values for ϵ and μ in (1), (2). The conductivity profiles of b_e and b_i inside the box domain are constant with values 0.1 Sm^{-1} and 0.5 Sm^{-1} , respectively.

In our numerical experiments we are using a finite volume (FV) technique for modelling (1), (2) similar to the one described in [5], but we are also experimenting with an alternative approach using a Finite Differences Frequency Domain (FDFD) technique similar to the one described in [2]. All data are simulated with the same forward model-

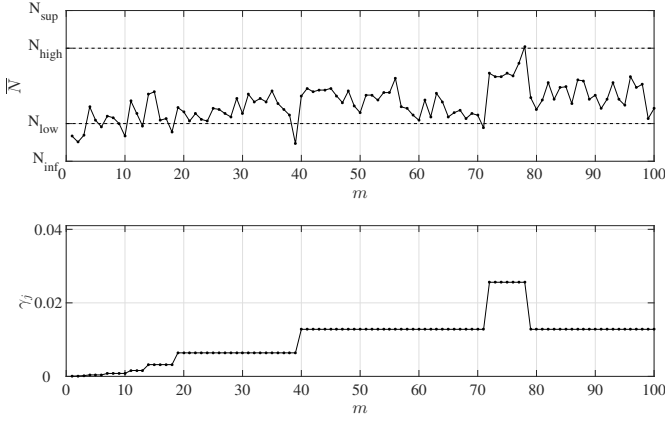


Figure 3. Evolution of $\bar{N}$ (upper image) and γ_j (lower image) vs sweep number in Numerical Experiment 1.

ling code as used in the reconstruction but then 1% Gaussian noise is added to reduce the so-called ‘inverse crime’. For the future we plan to generate data with one technique (e.g. FDFD) and do the reconstructions with an alternative technique (e.g. FV).

We have used $[N_{inf}, N_{sup}] = [0, 15]$, $[N_{low}, N_{high}] = [3.75, 11.25]$, $\tau_1 = 0.5$, $\tau_2 = 2.0$ in both numerical experiments. Figure 2 shows the shape evolution and reconstructions of Experiment 1, whereas Figure 4 shows the one of Experiment 2. Figures 3 and 5 show the evolution of the average voxel change per sweep for both experiments, and as a result the evolution of the step size γ_j due to our line search strategy. Both N_{low} and N_{high} are indicated by dashed lines in the upper graph of Figures 3 and 5.

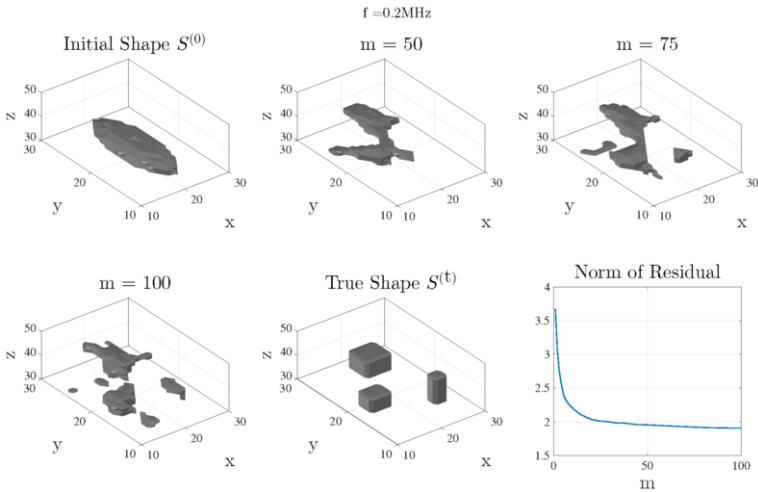


Figure 4. Shape and cost evolution vs sweep number in Numerical Experiment 2 (cargo container screening).

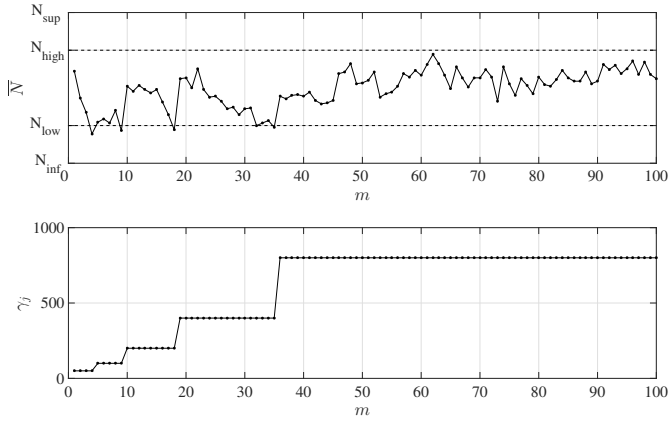


Figure 5. Evolution of $\bar{N}$ (upper image) and γ_j (lower image) vs sweep number in Numerical Experiment 2.

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Influence of the Plastic Deformation on Magnetic Properties of Austenitic Steels

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Abstract. Austenitic stainless biomaterials are worldwide used in biomedical practice. Electromagnetic nondestructive evaluation allows their contactless investigation in terms of magnetic field mapping. Intrinsic magnetic field directly reflects changes in their mechanical properties. This article presents investigation of concrete austenitic biomaterials by use of fluxgate sensor. Pre-defined plastic deformation levels of the cylindrical specimens are evaluated. Three-dimensional scanning procedure has been performed.

Keywords. Austenitic stainless steel, magnetic field mapping, nondestructive evaluation.

1. Introduction

Since society first realized the fallibility of people and their machines, they have recognized a need to inspect these machines in order to prevent failures. A wide variety of test schemes exist both destructive and the nondestructive. The practical benefits of nondestructive inspection are obvious, as long as the results are reliable and the inspection cost-effective. Load conditions, material properties and flaw size will determine if a flaw is likely to affect the serviceability of the part. NDE, however, is essential in locating and sizing of the flaw. NDE evaluation can be also beneficial to reducing the frequency of unscheduled maintenance which usually is more expensive than regularly scheduled maintenance, [1].

Inspection of conductive materials is however extremely important to the field of biomedical engineering. Biomaterial implants that have the same function as the bones (implants which replace failed hard tissue, artificial hip joints, bone plates and dental implants) are usually used under severe cyclic loading conditions. Therefore, metallic materials that typically exhibit high strength, ductility and toughness are the main candidates for the structural biomaterials of these implants. These implants must exhibit high biocompatibility, high performance and reliability for long-term use. The reliability of the implants after implantations is determined by their failure after the

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critical period of infection. Therefore, the mechanical properties of structural biomaterials in a living body environment (fatigue, toughness, wear resistance) need to be evaluated in order to confidently use the implants for a long period of time. Some metals and alloys that combine high strength with reasonable corrosion resistance are favorite biomaterials for the fabrication of orthopedic implants, which are subjected to severe mechanical loading into the human body. The metallic biomaterials most commonly used for orthopedic applications are the austenitic stainless steels [2], [3], [4], [5].

This article will deal with the three-dimensional analysis of the conductive austenitic biomaterials after a previous defined plastic deformation. An intrinsic magnetic field of the specimens is analyzed, displayed and discussed. Because the magnetic field strength is comparable with the Earth's magnetic field strength a commercial sensor (fluxgate type) is used for the measurements. The main aim is to reveal a correlation between the deformation structure and its influence on resulting magnetic field of the biomaterials. The originality of this study lies in 3D visualization of the intrinsic magnetic field of inspected biomaterials. Individual components of the field are sensed to bring new information about the behaviour of the field after plastic deformation.

2. Austenitic biomaterials properties

Biomaterials are the materials that usefully interact with biological systems. They are often used or adapted for a medical application, and thus comprise whole or part of a living structure or biomedical device which performs, augments, or replaces a natural function. One of the primary reasons for which biomaterials are used is to replace hard or soft tissues that have become damaged or destroyed through some pathological processes. Although the tissues and structures of the body which operate for an extended period of time in most people, they do suffer from a variety of destructive processes, including fracture, infection, and cancer that cause pain, disfigurement, or loss of function. Under these circumstances, it may be possible to remove the diseased tissue and replace it with some suitable synthetic material. One of the main application areas for biomaterials is for orthopedic implant devices. Both osteoarthritis and rheumatoid arthritis affect the structure of freely movable (synovial) joints, such as the hip, knee, shoulder, ankle, and elbow. The conductive biomaterials are the most widely used for load-bearing implants. For instance, some of the most common orthopedic surgeries involve the implantation of metallic implants. These range from simple wires and screws to fracture fixation plates and total joint prostheses (artificial joints) for hips, knees, shoulders, ankles etc., [5], [6], [7], [8]. In addition to orthopedics, the metallic implants are used in maxillofacial surgery, cardiovascular surgery, and as dental materials. Although many metals and alloys are used for medical device applications, the most commonly employed are austenitic stainless steels, commercially pure titanium and titanium alloys, and cobalt-base alloys. In general, these materials are paramagnetic. Magnetic properties of the stainless steels may be affected by their alloying elements, atomic grain structure and amount of the cold-working during fabrication. In their basic forms the stainless steels have a ferrite grain structure, similar to carbon steel, and they are ferromagnetic, Table.1. The addition of nickel in the 300-series stainless steels modifies the crystal grain structure to austenitic structure, which is paramagnetic. The austenitic grades are mostly non-magnetic in the unworked state due to their nickel content.

Table 1. Spectral analysis of the stainless steels chemical composition

Material	C	Si	Mn	Cr	Mo	Ni	N	Ti	Fe
AISI 304	0.0509	0.401	1.33	19.88	0.235	8.19	0.15	0.0048	rest
AISI 316L	0.0457	0.557	1.40	17.92	2.26	10.32	0.04	0.0095	rest
AISI 316Ti	0.0537	0.716	1.39	17.70	2.28	11.11	0.02	0.145	rest

When 300-series stainless steels are cold-worked, straining of the atomic lattice structure in the areas of cold-working forms the ferromagnetic grain martensitic structure. Generally speaking, the higher the nickel content in the steel the more stable the austenitic structure and less magnetic response from cold-working. Consequently 316 stainless steel containing higher amounts of nickel, exhibits virtually no magnetism after cold-working in most cases, while 304, with lower nickel content, may become partially ferromagnetic. Austenitic (300-series) stainless steels that have become magnetic due to work hardening can be returned to a paramagnetic state through annealing or stress-relieving. Brief heating at elevated temperatures revert the affected grain structure from the martensitic state to the austenitic one. Since 400-series stainless steels are entirely ferrite or martensitic, their magnetic properties cannot be reduced through annealing. There is no plating or finishing processes, such as passivation, that can reduce or eliminate work hardening induced magnetism. They are merely superficial and do not change the affected grain structure.

3. Experimental set-up

Experimental material specimens – austenitic stainless steel having a quasi-cylindrical shape, Fig.1, is inspected in this study.

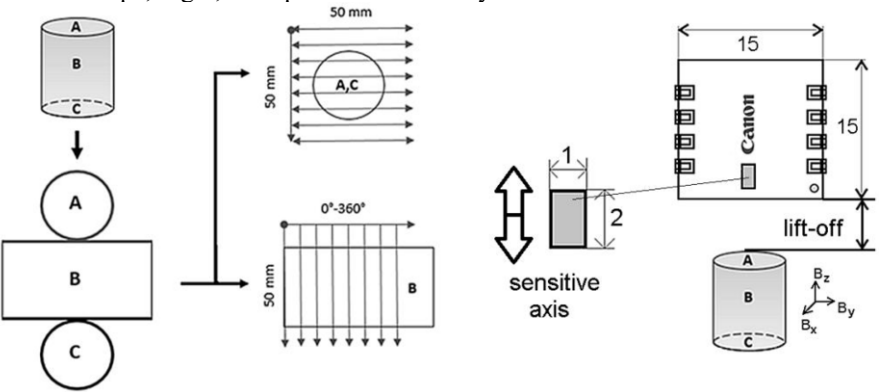


Figure 1. Scanning procedure detail (left) and configuration of the fluxgate sensor (right).

Three different biomaterial grades are inspected, respectively: chrome-nickel stainless steel AISI 304, chrome-nickel-molybdenum stainless steel AISI 316L and chrome-nickel-molybdenum-titan stainless steel AISI 316Ti. Initially, all the inspected specimens are prepared from one piece of material, by the cutting at the defined dimensions. Thus, all the specimens have the cylindrical shape, with an initial height (measured from the base to the top) of $h = 14$ mm and with a radius of $r = 6$ mm. It is very important to be all the specimens dissolution-annealed after the mechanical cutting. The samples underwent the recrystallization annealing process at the

temperature of $T = 850\text{ }^{\circ}\text{C}$ for time of $t = 15\text{ min}$, followed by air cooling to reduce the strength applied during cold-working. Thus, the residual magnetic field caused by the cutting is eliminated. This regime is defined as the reference state. Then the process of the controlled plastic deformation may be applied. The deformation is performed by mechanical pressing of the opposite sides of the cylindrical specimens, using the ripper system. Seven different values of the applied plastic deformation are realized: $\text{PD}_1 = 1\%$, $\text{PD}_2 = 2\%$, $\text{PD}_3 = 5\%$, $\text{PD}_4 = 10\%$, $\text{PD}_5 = 20\%$, $\text{PD}_6 = 30\%$ and $\text{PD}_7 = 40\%$. These percentage values represent shortening of the specimen after the applied deformation, in comparison with the reference (non-deformed) specimen, which has the length of $P_{\text{REF}} = 100\%$. The fluxgate magnetometer, shown in Fig. 2, is used as the sensing element for measuring of the intrinsic magnetic field. This sensor picks-up the response signal in the sensitive axis. Its voltage output signal is converted to the magnetic field strength value, using the appropriate calibration curve of the sensor. The commercial type is used (Canon Inc.). The sensor allows to measure magnetic field in the frequency range of $f \in (0; 3.4\text{ kHz})$. Individual magnetic field component is sensed, respectively. Thus, three-dimensional scanning is achieved for each specimen. Each specimen is inspected as follows: a classical raster 2D scan is performed above the specimen. Thus, the inspected area has a square shape, with dimensions of $\text{dims} = 50\text{ mm} \times 50\text{ mm}$, while the resolution between individual scanning lines is set to $\text{res} = 0.25\text{ mm}$. It gives in results of $n = 100$ scanning lines per each specimen. Further, the sensor is positioned in constant distance from the top side of each specimen, at lift-off distance of $l_o = 2\text{ mm}$. This distance between the surface of the specimen to be inspected and the center of the sensing element is kept at constant value. Thus, all the specimens are inspected under the same conditions. The gained signals are acquired with the resolution of $r_{\text{DAQ}} = 16\text{ bits/channel}$ and with sampling frequency of $f_s = 2\text{ kS/sec}$. User interface for data manipulation, controlling the stage and processing the data are created and designed using the LabVIEW (Virtual Instrumentation).

4. Experimental results and discussion

The obtained results of the experimental measurements are introduced in this section. Figures 2 - 5 display visualization of the intrinsic magnetic field in terms of the individual sensed components. Images present the all parts of individual specimens: the base, the top and the shell, respectively. The dotted line in the graphs represents the profile of the specimens. Obtained signals for the rotation of the specimens were obtained as follows: the fluxgate sensor was kept at the constant lift-off position and the specimens were rotated with step of ten degrees. Each picture represents magnetic field distribution above the specimen. Only the selected results are presents in this section. As can be seen, when plastic deformation of 10% was applied, the strongest response arise AISI 304 biomaterial. Further, the AISI 316 Ti magnetic field is also detectable, although with lower intensity. Finally, the AISI 316 L magnetic fields was too weak to be detected via fluxgate sensor.

Almost the same conclusion can be stated for the highest plastic deformation level of $\text{PD}_7 = 40\%$. The gained results present stronger field intensity in comparison with all the previous states. Moreover, these specimens were successfully inspected from higher lift-off distance of $l_o = 10\text{ mm}$. (This fact is needed mainly in biomedical practice: contactless detection ability from distances of several centimeters.)

All of the obtained results and discussed facts are in direct correlation with the theoretical background, from the material engineering point of view. It means that structure of this experimental material has strong influence on its magnetic properties after previous plastic deformation.

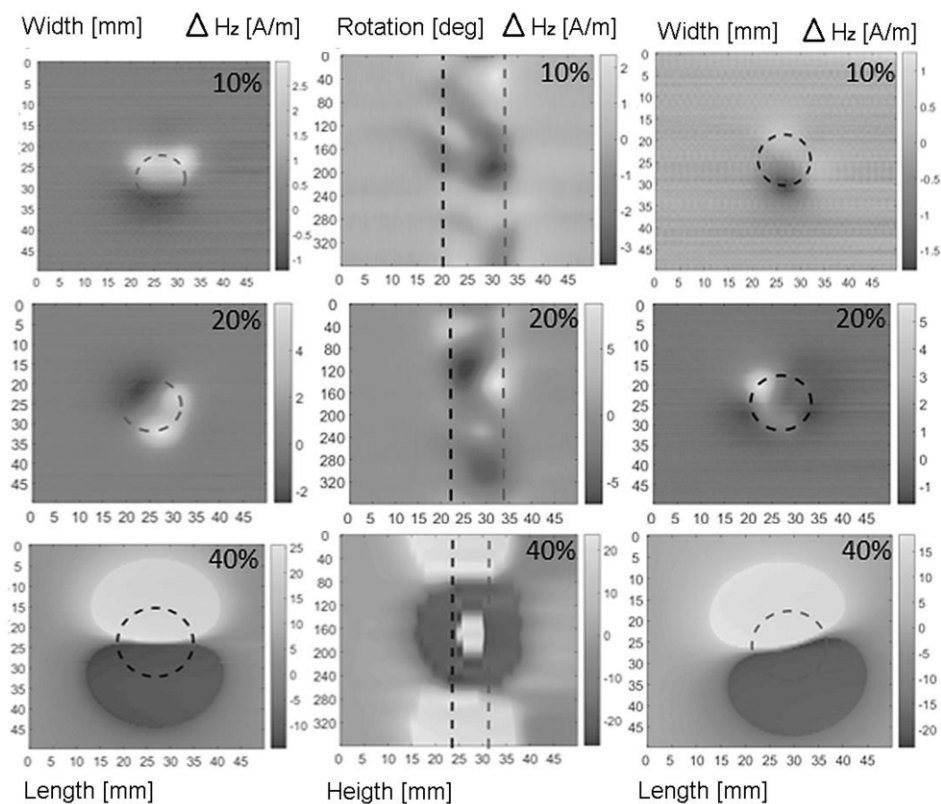


Figure 2. Experimental results: Z-component differential responses, lift-off = 2 mm, AISI 304, (dotted line = profile of the specimen)

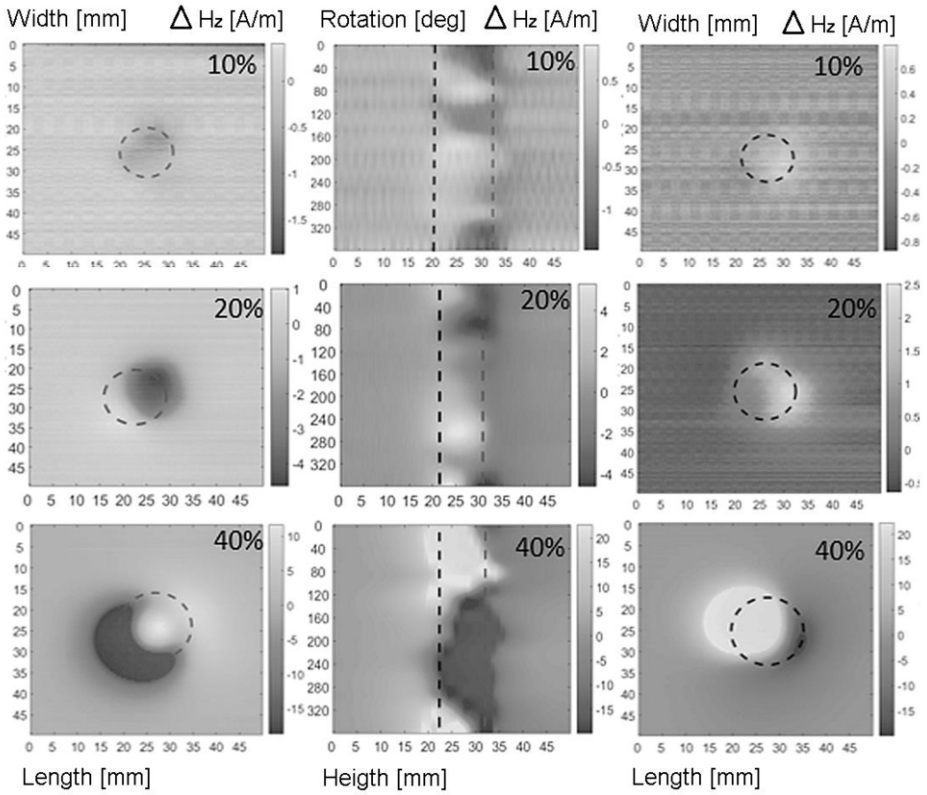


Figure 3. Experimental results: Z-component, lift-off = 2 mm, differential responses, AISI 316Ti, (dotted line = profile of the specimen).

Figure 4 displays the 3D magnetic field reconstruction obtained for the inspected specimens. This comparison is introduced to be visible the differences among individual steel grades. The dependence of the individual components of ΔH over individual axes is shown. The full-line represents profile of the sample. As can be seen from these results, the most-uniform shape of the field occurs for AISI 304 specimens. Towards to weaker magnetic properties of the material, magnetic field becomes more non-uniform. This information is useful for the inspection process: directional properties of these cylindrical specimens may vary when different grade of material is used.

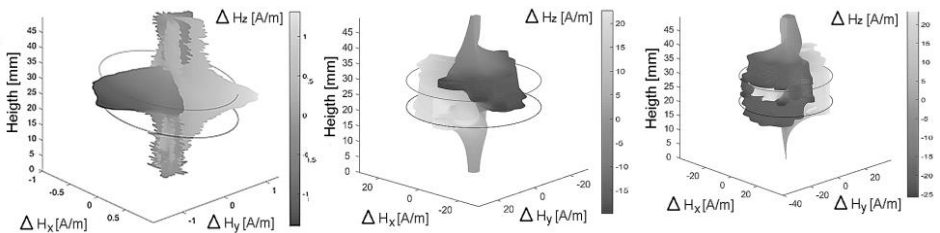


Figure 4. Experimental results: $PD_7 = 40\%$, 3D visualization of the intrinsic magnetic field, AISI 316L (left), AISI 316Ti (center) and AISI 304 (right).

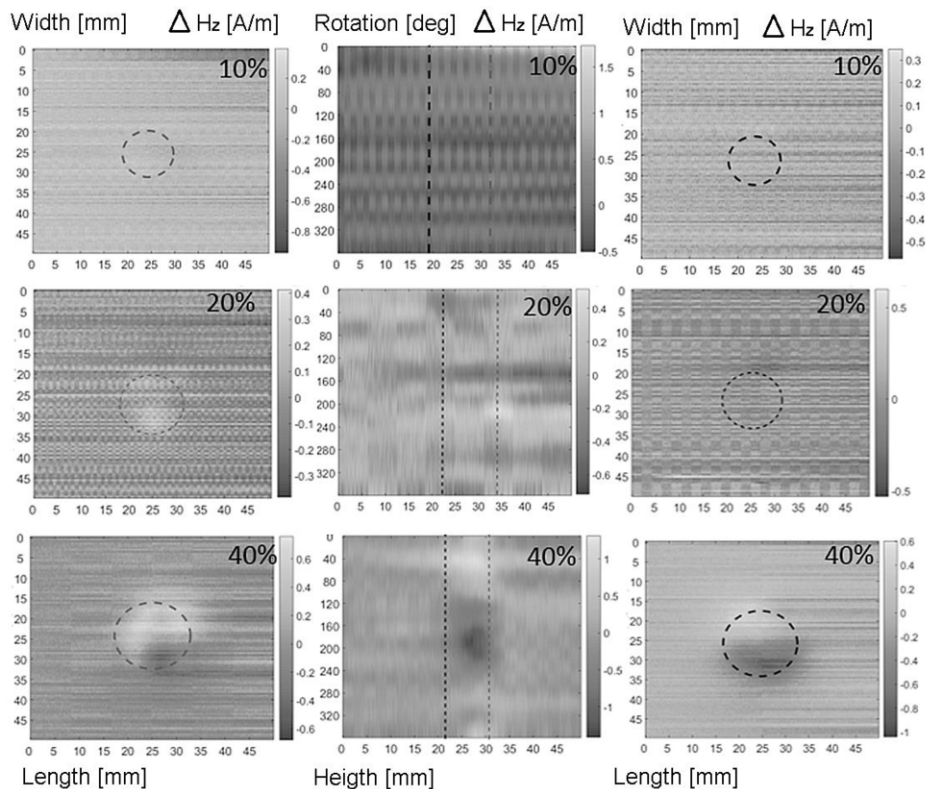


Figure 5. Experimental results: Z-component, lift-off = 2 mm, differential responses, AISI 316L, (dotted line = profile of the specimen).

5. Conclusion

The article dealt with investigation of the austenitic stainless steels after various level of plastic deformation. All inspected materials are intensively used in biomedicine as for implant or replacement purposes. The presented experimental work showed that these materials become partially ferromagnetic due to the austenitic to martensitic phase transition, which can be sensed by use of sensitive magnetic field sensor. Intrinsic magnetic field sensing at individual three-components brought new information about characterizing of these materials. Plastic deformation is non-homogeneous and non-linear process. Thus, by measuring of electromagnetic properties the deformation level can be indirectly investigated.

The presented results showed that the plastic deformation level is in direct correlation with magnetic field strength, measured by the fluxgate sensor. The strongest response were sensed for the AISI 304 specimens and on the other hand very weak signal was detected for the AISI 316L one. This biomaterial grade is limited in evaluation for this kind of inspection. Further, the AISI 316Ti material reveals stronger magnetic field than of the same grade of 316L. This is caused by the addition of titanium into the composition. Differences in the microstructure of the inspected biomaterials after the previous heat treatment observed by light microscope was experimentally confirmed by

means of performed measurement. This means that magnetic properties of the materials are to way to determine its mechanical properties. This fact is very important to be these materials evaluated nondestructively. This finds many applications on the field of biomedicine as well as other technical applications.

It can be concluded that proposed procedure offers relevant results for these biomaterials investigation. Further work will focus on quantification of the deformation martensitic component presented in these biomaterials. Main challenge is to show low deformation levels at higher lift-off distances with commercial sensors.

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EM Sensor System for Characterisation of Advanced High Strength Strip Steels

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Abstract. In order to obtain accurate quality control of steel products, it is desirable to be able to monitor the mechanical properties non-destructively. This study proposes using an electromagnetic (EM) sensor, suitable for use on strip samples, as a tool for non-destructive steel characterisation, in particular phase balance and grain size in dual phase (DP) steels and hence strength. It is known that the low frequency inductance measured using an EM sensor depends on the relative permeability of the sample and that the permeability is affected by microstructural features (i.e. phase fraction / distribution and, to a lesser extent, grain size are the important features in DP steel). EM sensors can be used to characterise austenite and ferrite fraction in hot strip mills (EMspecTM system) and for statistical correlations to mechanical properties (IMPOC and HACOM systems) in cold strip mills. In this paper an EM sensor system and a finite element based sensor – microstructure model have been used to characterize DP microstructures (using commercial DP600, DP800 and DP1000 grades and a heat treated DP600 grade) taking into account the effect of strip thickness on the signals. This paper also reports on the relationship between the ferrite fraction and tensile strength, which follows the expected relationship from the literature, and the magnetic permeability (determined from the EM sensor signal which is directly related to ferrite fraction and grain size) and tensile strength. The ability of the EM sensor to determine the tensile strength is therefore illustrated taking into account strip thickness and microstructure.

Keywords. Electromagnetic sensor, permeability, mechanical property, DP steel

1. Introduction

The increasing requirements of passenger safety, vehicle performance and fuel efficiency in the automotive industry means that improved steel grades are required. There has been growing attention to the manufacturing of dual phase (DP) steels for their high strength allowing light-weighting of automotive components. DP steels are characterised by a combination of continuous yielding behaviour, good formability and high tensile strength [1]. The microstructure of DP steels generally consists of a soft matrix of ferrite with typically 20-80% dispersion of second phase (e.g. martensite, tempered martensite and/or bainite) [2]. The mechanical properties of DP steel are strongly influenced by its microstructural parameters, in particular phase balance and grain size [3-5].

A variety of techniques can be used to characterise the microstructure, the most frequently used are optical microscopy and scanning electron microscopy (SEM) on

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polished and etched samples. These conventional methods to obtain phase information are destructive, and time consuming, as they require a small piece of material to be removed from the strip / component.

Multi-frequency electromagnetic (EM) sensors are sensitive to both changes in relative permeability and resistivity of steel, with the low frequency inductance values being directly related to the permeability. The EM sensor measurements are non-destructive, have a fast response, can operate in a non-contact manner, are unaffected by dust, water and are relatively inexpensive [6]. In recent years, the use of magnetic techniques based on EM sensors, for non-destructive testing have increased [6-9]. Previously EM sensors have been shown to be able to monitor the transformation from austenite to ferrite, below the Curie temperature, and to distinguish between samples with mixed microstructures (ferrite + austenite; ferrite + pearlite; ferrite + martensite) across the whole range of ferrite percentages [10, 11]. Hence there is potential for using a U-shaped EM sensor to characterise the microstructure in DP steels, with verification previously shown for DP structures within the range of ferrite fraction from 35-72% using samples of constant thickness. As there are known relationships between the microstructure of DP steels with tensile strength [3-5,12], the EM signals can therefore be used to estimate strength and/or hardness.

Previous work has established finite element (FE) based models to relate the microstructure (phase balance) to the magnetic property of low field permeability, allowing the microstructure (ferrite fraction) to be determined from the EM sensor signal [13]. In that work only single sheet thickness DP steel samples were considered and a full sensor model, which could take into account sample geometry, was not available. This paper reports on the use of an EM sensor for characterising DP steels (microstructure and tensile strength) and the development of a calibration procedure, using a full sensor-sample FE model, to account for strip thickness.

2. Experimental Procedure

A number of commercial DP600, 800 and 1000 grade steels of different thicknesses (1 – 4 mm), supplied by Tata Steel Europe and a set of heat treated DP600 grade of 1.4 mm thickness (heat treated to generate different ferrite-martensite fractions from 35% to 72% ferrite) have been investigated. Metallographic samples were prepared by mechanical polishing using successive polishing steps (to a 0.05 μm final surface finish). The polished surfaces of the specimens were etched in Nital 2%. Up to eight optical microscopy images from different locations for each DP sample were taken (from the transverse section) and ferrite percentages were calculated using Image J software.

Vickers hardness testing was carried out using a 500g load and each sample hardness value was determined by taking the average of ten measurements. Flat tensile specimens of 80mm gauge length and 20mm gauge width were prepared as per ASTM E8M standard; tensile tests were conducted using a Static Instron 100kN testing machine at a strain rate of 0.002s⁻¹ and stress-strain plots were obtained for each sample. Four repeated tensile tests were carried out for each DP steel. The EM measurements were carried out using a U shaped multi-frequency EM sensor on sheet samples. A minimum sheet size of 300 x 80mm was required to avoid edge effects during testing (although this could be accounted for using the FE sensor model if required). The sensor consists of one excitation coil with 100 turns and two sensing coils with 86 turns with a bridge of 100mm, leg lengths and thickness of 56mm and 25mm respectively. The EM sensor's

excitation coils were driven using an impedance analyser Solartron 1260 with an AC voltage of 3V at a frequency of 10Hz and measurements of inductance in the sensing coil were taken for the specimens.

The samples for incremental permeability measurements (strip shape with 50mm length and 5mm width) were prepared by EDM wire cutting from the as-received DP steels. A BH measurement system, developed at the University of Manchester [14] and shown in Figure 1, was used to measure the permeability of the samples. A current with low frequency time varying signal was applied to the excitation coils wrapped around the silicon steel core. The strip samples were fitted into the slot in the core to maximise coupling between the core and the sample. The axial applied field (H) and the flux density of the induced field (B) were measured. A sinusoidal excitation of 10 Hz was used to generate minor loops (deviation from the initial magnetisation curve). The sample was firstly demagnetized then the applied field (H value) gradually increased and minor loop cycles recorded. The incremental permeability was calculated according to Eq. (1);

$$\mu_{\Delta} = \Delta B / (\Delta H \cdot \mu_0) \quad \text{Eq. (1)}$$

where μ_0 is the permeability of free space.

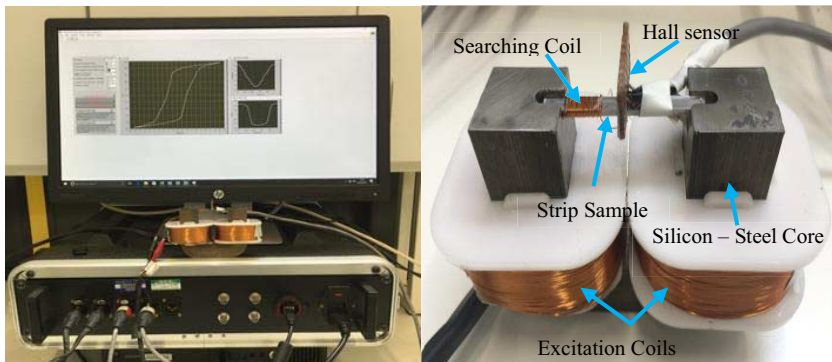


Figure 1. The BH measurement system, developed at the University of Manchester, to measure BH hysteresis loops and permeability of strip sample.

3. Results and Discussion

3.1. Microstructure and mechanical properties

Typical SEM images of the DP600, DP800 and DP1000 steels are shown in Figure 2. As can be seen, the microstructure of dual phase steels consists of a ferrite matrix containing a second phase in the form of islands (martensite, bainite and/or tempered martensite) [15]. Different volume fractions of ferrite and martensite / bainite are present in the different grades of DP steels, used to give the different strength levels [3,4,15].

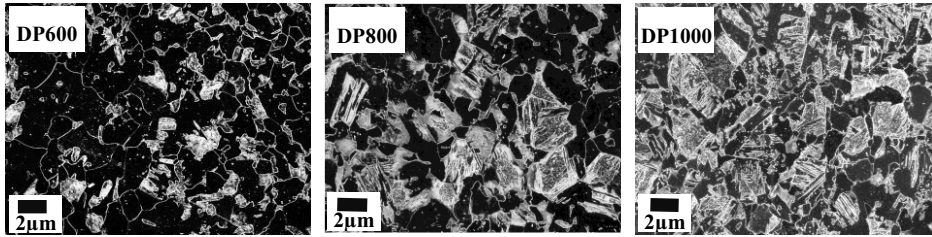


Figure 2. SEM images of commercial DP600, DP800 and DP1000 steels. The microstructures consist of a ferrite matrix (black) containing a second phase in the form of islands (martensite, bainite and/or tempered martensite).

To determine the relationship between the phase fraction and strength, the ferrite fraction and tensile strength were quantified. The decrease in tensile strength with increasing volume fraction of ferrite for the DP steel is well known and widely documented [2,4,5,16]. This is in agreement with the results shown in Figure 3. There is a clear decrease in the tensile strength with an increase in ferrite fraction for DP grades, which is related to the lower martensite (or bainite) fraction.

The trend follows an approximately linear relationship for the DP samples ($R^2 = 0.86$). This trend stems mostly from the composite effect due to the presence of hard particles in a soft matrix. From Figure 3 it can be seen that there is a lot more scatter for the DP800 grades. To determine the cause of the scatter, a closer examination of their microstructures was carried out and in particular, the ferrite grain size was investigated. It was found that the scatter can be related to the difference in grain size between the samples (as the DP800 strips came from different mills and hence processing conditions, which can give different grain sizes), with the samples above the best fit line having a finer grain size than those below the best fit line. The approximately linear relationship between strength and ferrite fraction holds well if the ferrite grain size does not alter between samples, however if ferrite grain size varies then this also needs to be taken into account. It should be noted that for DP1000 material the effect of ferrite grain size variation will be less significant on strength as the ferrite fraction is low (approximately 40%).

In term of mechanical properties, grain boundaries may act as obstacles to dislocation slip. At room temperature, yield stress σ_y depends on the inverse square root of the grain size, i.e. the Hall–Petch relationship [15–19] and grain size affects tensile strength [20]. It is known that the ferrite grain size affects the magnetic properties in low carbon steel as the grain boundaries act as effective pinning points to magnetic domain movement [21–23]. Therefore it might be expected that not only ferrite fraction but also ferrite grain size will affect the magnetic properties in DP steels. The results in Figure 4 show a strong effect of ferrite fraction on the low field permeability, as has been seen previously [24]. In addition, the two DP800 samples with the smaller ferrite grain size show lower permeability values than the other DP800 samples indicating that grain size has a significant effect on the magnetic property (permeability) in these steels as well as ferrite fraction.

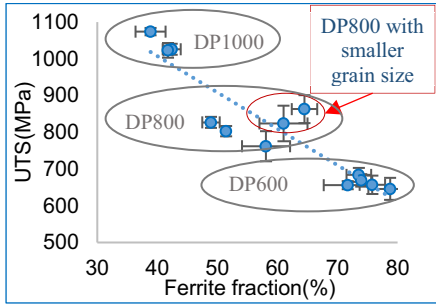


Figure 3. Ultimate tensile strength as a function of ferrite fraction for commercial DP600, DP800 and DP1000 steel.

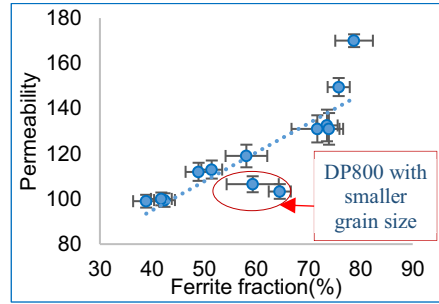


Figure 4. The permeability as a function of ferrite fraction for DP steel, the marked samples showing lower than expected permeability due to the smaller grain size.

Since both tensile strength and permeability are affected by ferrite fraction and grain size it might be expected that the two properties will show a strong correlation. The incremental permeability (determined from the BH hysteresis loops) is plotted against tensile strength in Figure 5. The results indicate that there is a general correlation between permeability and tensile strength for the commercial DP steels. A DP steel with higher volume fraction of ferrite (and/or larger grain size) showing a higher relative permeability and lower tensile strength. The relative permeability value increases with an increase in ferrite fraction due to the effect of decreased dislocation density (from martensite formation), easier reverse domain formation and domain wall motions take place at a lower applied field in ferrite [23]. The correlation in Figure 5 is not linear but there is less scatter around a line of best fit than in Figure 4.

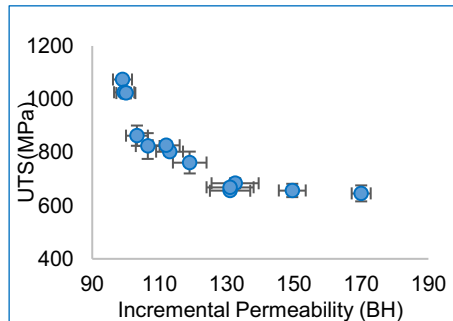


Figure 5. Measured incremental permeability values (from the BH hysteresis loops) for the commercial DP600, 800 and 1000 steels plotted against tensile strength.

3.2. EM sensor measurements and FE Modelling

The real inductance versus frequency plot for the U-shaped EM sensor shows a plateau in inductance value at low frequency (10-100Hz) in the region where the signal is independent of the electrical resistivity but dependent on the relative permeability of the sample. It can be seen in Figure 6 that the EM sensor signal of low frequency inductance increases almost linearly with ferrite fraction in the range of 35-73% for the uniform

thickness (1.4mm) heat treated DP600 steels, which all have a similar grain size. On the other hand, from Figure 7, it can be observed that for the commercial DP steel grades with different thicknesses (and similar ferrite fraction) the inductance is strongly affected by the thickness of the sample.

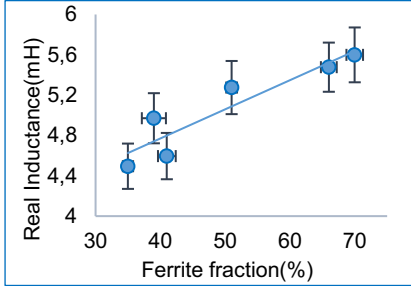


Figure 6. The real inductance measurements (using U shaped EM sensor) of heat treated DP600 steel at a frequency of 10Hz versus ferrite fraction.

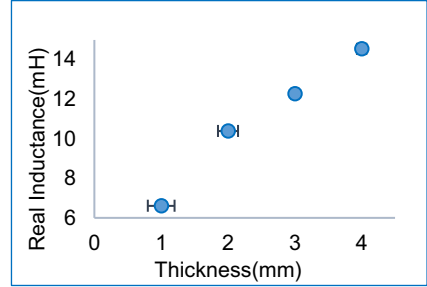


Figure 7. The real inductance measurements (using U shaped EM sensor at a frequency of 10Hz) of DP600 samples (with 79% ferrite, grain size $10 \pm 4 \mu\text{m}$) steels.

The thickness of material affecting the sensor signal can be estimated by the skin depth equation, Eq.(2);

$$\Delta s = \sqrt{\frac{\rho}{\pi f \mu_0 \mu_r}} \quad \text{Eq. (2)}$$

where ρ is the resistivity of the conductor, f is the frequency, μ is absolute magnetic permeability ($\mu = \mu_0 \mu_r$). By assuming that resistivity, $\rho = 2.1 \times 10^{-7} \Omega \cdot \text{m}$ and permeability for DP steel is around 100 to 170, the value of skin depth for the current set up and operation frequency of 10 Hz, was calculated to be approximately 6mm, which is larger than the sample thickness. Therefore, a thicker sample shows a higher mutual inductance sensor response as effectively more material is being measured. This means that the sensor signal cannot be correlated to tensile strength directly if different thickness samples are assessed unless a calibration curve to account for thickness is generated.

A calibration curve to consider thickness, for different DP samples, was constructed using a 3D finite element model developed in COMSOL Multiphysics, Figure 8.

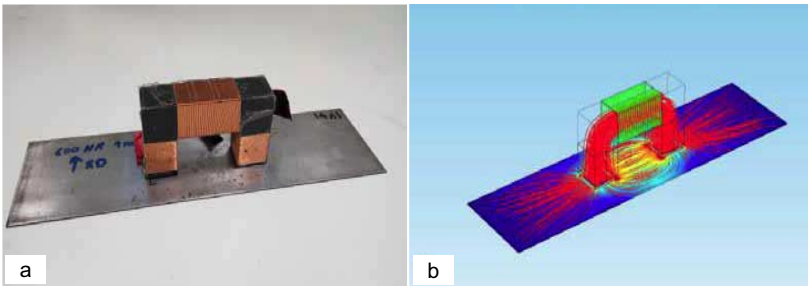


Figure 8. U-shaped EM sensor on the strip sample (a) and 3D U-shaped FE model to estimate the permeability of specimens (b).

The 3D sensor model was constructed based on the measured geometry of the sensor and number of windings for the excitation and sensing coils. The model was validated using measured signals for samples of known permeability (determined using the BH system) and thickness. It is important to note that the low field permeability value is required as the EM sensors used in this work apply a field in the range of 500A/m and the measured value of permeability is a function of magnetic field.

Figure 9 illustrates the calibration curves to obtain the permeability from the EM sensor mutual inductance value and measured sample thickness. Both model generated curves and validation points from the DP steels used in this work are shown. Good agreement between the model and measured values can be seen. The results from Figure 5 show that there is a link between permeability and tensile strength, which indicates that measuring the permeability using an EM sensor on steel strip samples will allow the tensile strength to be predicted. In addition information about the ferrite fraction in the steel can be obtained from the permeability (Figure 4) if the grain size remains similar. The calibration curve is required so that any DP sample thickness can be tested with the material permeability being determined, which can then be used to determine the strength. Using the calibration curve and the measured inductance from the EM sensor the permeability of the commercial DP samples can be determined, hence the tensile strength can be accurately predicted using Figure 5.

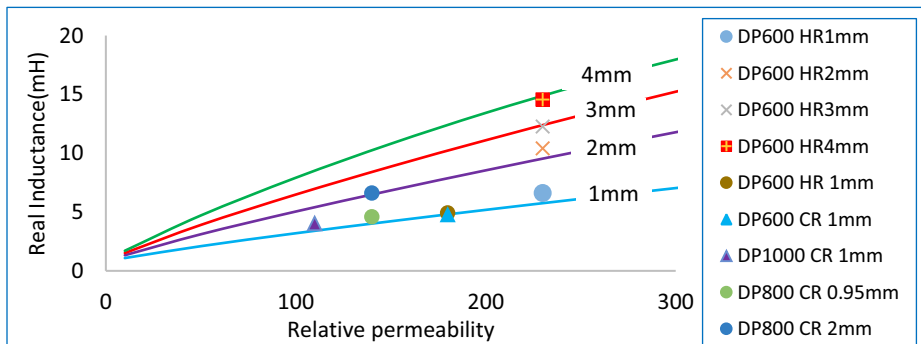


Figure 9. Calibration curves relating low frequency (10Hz) real inductance with permeability for different thickness samples. The lines represent modelling results and experimental data for samples are indicated by different points.

4. Conclusions

Previously EM sensors have been shown to be sensitive to changes in ferrite fraction in DP steel sheet of constant thickness with little variance in grain size. In this paper it is shown that the permeability can be determined from the low frequency mutual inductance measured using a U-shaped electromagnetic sensor for any sheet thickness using a calibration curve to account for the effect of thickness. The calibration curve was determined using a FE model for the sensor and sample geometry and verified with experimental results. The permeability is affected by ferrite fraction and ferrite grain size for DP steels, both of which affect the tensile strength, therefore a single relationship between permeability and tensile strength results. The permeability can then be used to estimate the tensile strength for DP steels.

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Linear Sampling Method for Localizing Small Anomalies in Microwave Imaging

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Abstract. We design a linear sampling method to retrieve the shape of an anomaly in microwave imaging. It is based on the structure of nonzero singular vectors associated with the nonzero singular values of a complex-symmetric matrix, whose elements are measured S -parameters. Simulation results show the effectiveness and limitations of the linear sampling method in real-world microwave imaging.

Keywords. Linear sampling method, small anomaly, microwave imaging, S -parameter, simulation result

Introduction

Herein, we consider an inverse scattering problem to retrieve the shape of a small anomaly whose permittivity and conductivity differ from those of the embedding medium using S -parameters that are collected at a single fixed angular frequency. The underlying application involves the imaging of early-stage tumor-like anomalies from the measured scattered field or S -parameters. Related research can be found in [1,2,3,4,5,6] and references therein. This problem is known to be difficult because of its intrinsic nonlinearity and ill-posedness; however, it is a considerably interesting and important problem.

Motivated by this, various reconstruction techniques have been investigated for example level-set method [7,8,9,10], MULTiple Signal Classification (MUSIC) [11,12,13,14,15], subspace and Kirchhoff migrations [16,17,18,19,20], topological derivative [21,22,23,24,25], and direct sampling method [26,27,28,29]. Among them, linear sampling method has been remarkably developed and effectively applied to various inverse scattering problems in order to locate small scatterers and image volumetric objects or crack-like defects [30,31,32,33,34,35,36,37]. However, to the best of our knowledge, the linear sampling method has not yet been considered as a method for identifying the shape of unknown anomalies from the collected S -parameters.

In this research, we design a linear sampling method to retrieve the shape of an anomaly whose dielectric permittivity and conductivity are different from those of the embedding medium using the collected S -parameters. To demonstrate the effectiveness of the proposed method and confirm the theoretical results, we present the results of simulations using synthetic data generated through the CST STUDIO SUITE.

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This study is organized as follows. In Section 1, we briefly survey the forward problem and introduce the concept of the S -parameter. In Section 2, we design a linear sampling method to identify the unknown shape of an anomaly from the collected S -parameter data. In Section 3, we present the simulation results for the small and extended anomalies. Section 4 presents our conclusions.

Furthermore, it is well known that the linear sampling method and the Multiple Signal Classification (MUSIC) algorithm are somewhat similar (see [31] for instance). We refer to a recent study [40] that focuses on the application of the MUSIC algorithm by employing S -parameters.

1. Forward problem and S -parameter

We briefly introduce the forward problem to detect the shape of an anomaly D . For the purpose of simplicity, we assume that D is a small ball with radius α , which is located at $\mathbf{r}_D$ such that

$$D = \mathbf{r}_D + \alpha\mathbf{B},$$

where $\mathbf{B}$ denotes a simply connected domain. Again, for the purpose of simplicity, we assume that $\mathbf{B}$ is a unit sphere. Throughout this study, we assume that D is surrounded by N different dipole antennas located at $\mathbf{a}_n$, $n = 1, 2, \dots, N$, and that all materials are characterized by their dielectric permittivity and electrical conductivity at a given angular frequency $\omega = 2\pi f$, i.e., the permeability $\mu \equiv \mu_B$ is the same for all materials. Based on assumption, we set ε_B and σ_B , respectively, as the background permittivity and conductivity and ε_D and σ_D , respectively, to those of D . We can then specify the following piecewise constant permittivity $\varepsilon(\mathbf{r})$ and the conductivity $\sigma(\mathbf{r})$:

$$\varepsilon(\mathbf{r}) = \begin{cases} \varepsilon_D, & \mathbf{r} \in D, \\ \varepsilon_B, & \mathbf{r} \in \mathbb{R}^3 \setminus \overline{D}, \end{cases} \quad \text{and} \quad \sigma(\mathbf{r}) = \begin{cases} \sigma_D, & \mathbf{r} \in D, \\ \sigma_B, & \mathbf{r} \in \mathbb{R}^3 \setminus \overline{D}, \end{cases}$$

respectively.

Let $\mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r})$ be the incident electric field in a homogeneous medium due to a point current density $\mathbf{J}(\mathbf{a}_n, \mathbf{r})$ at $\mathbf{a}_n$. Then, following from Maxwell equations, $\mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r})$ satisfies

$$\begin{aligned} \nabla \times \mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r}) &= -i\omega\mu_B \mathbf{H}(\mathbf{a}_n, \mathbf{r}) \\ \nabla \times \mathbf{H}(\mathbf{a}_n, \mathbf{r}) &= (\sigma_B + i\omega\varepsilon_B) \mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r}). \end{aligned}$$

Let $\mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n)$ be the total field in the presence of D . Similarly, $\mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n)$ satisfies

$$\begin{aligned} \nabla \times \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n) &= -i\omega\mu_B \mathbf{H}(\mathbf{r}, \mathbf{a}_n) \\ \nabla \times \mathbf{H}(\mathbf{r}, \mathbf{a}_n) &= (\sigma(\mathbf{r}) + i\omega\varepsilon(\mathbf{r})) \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n) \end{aligned}$$

with transmission condition on the boundary of D and radiation (or open boundary) condition

$$\lim_{|\mathbf{r}| \rightarrow \infty} \mathbf{r} \left(\nabla \times \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n) - ik\hat{\mathbf{r}} \times \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_n) \right) = 0.$$

Here, $\hat{\mathbf{r}} = \mathbf{r}/|\mathbf{r}|$ and k denotes the wavenumber

$$k^2 = \omega^2 \mu \left(\varepsilon_B + i \frac{\sigma_B}{\omega} \right) = \frac{2\pi}{\lambda},$$

where λ is the given wavelength. For a detailed discussion, please refer to [38]. In this research, we assume that the size of D is sufficiently small such that $\alpha \ll \lambda/2$. Let $S(m, n)$ be the S -parameter (or scattering parameter), which is the ratio of the reflected waves at the m th antenna to the incident waves at the n th antenna. Herein, $S_{\text{scat}}(m, n)$ denotes the scattered field S -parameter obtained by subtracting the S -parameters from the total and incident fields. Then, based on [1], $S_{\text{scat}}(m, n)$ can be expressed as follows:

$$S_{\text{scat}}(m, n) \approx -\frac{k^2}{4i\omega\mu_B} \int_D \left(\frac{\varepsilon(\mathbf{r}) - \varepsilon_B}{\varepsilon_B} + i \frac{\sigma(\mathbf{r}) - \sigma_B}{\omega\sigma_B} \right) \mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r}) \cdot \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_m) d\mathbf{r}. \quad (1)$$

2. Design of a linear sampling method

In this section, we design a linear sampling method to identify the shape of anomaly D from collected $S_{\text{scat}}(m, n)$. To this end, we generate a Multi-Static Response matrix, $\mathbb{K} \in \mathbb{C}^{N \times N}$, whose elements are $S_{\text{scat}}(m, n)$:

$$\begin{aligned} \mathbb{K} &= \left[S_{\text{scat}}(m, n) \right]_{m, n=1}^N \\ &= \left[-\frac{k^2}{4i\omega\mu_B} \int_D \left(\frac{\varepsilon(\mathbf{r}) - \varepsilon_B}{\varepsilon_B} + i \frac{\sigma(\mathbf{r}) - \sigma_B}{\omega\sigma_B} \right) \mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r}) \cdot \mathbf{E}_{\text{tot}}(\mathbf{r}, \mathbf{a}_m) d\mathbf{r} \right]_{m, n=1}^N. \end{aligned}$$

Note that because $\alpha \ll \lambda/2$, we can apply the Born approximation. Correspondingly, $S_{\text{scat}}(m, n)$ can be approximated as follows:

$$S_{\text{scat}}(m, n) \approx \alpha^3 \frac{ik^2 |\mathbf{B}|}{4\omega\mu_B} \left(\frac{\varepsilon_D - \varepsilon_B}{\varepsilon_B} + i \frac{\sigma_D - \sigma_B}{\omega\sigma_B} \right) \mathbf{E}_{\text{inc}}(\mathbf{a}_n, \mathbf{r}_D) \cdot \mathbf{E}_{\text{inc}}(\mathbf{r}_D, \mathbf{a}_m) + o(\alpha^3). \quad (2)$$

Based on approximation (2), we can design a linear sampling method to detect the shape of D as follows. Following [31, 39], the range of $(\mathbb{K}\mathbb{K}^*)^{1/4} = \mathbb{A}^{1/4}$ can be obtained as follows:

$$\text{Range}(\mathbb{A}^{1/4}) = \left\{ \mathbf{W} \in \mathbb{C}^{N \times 1} : \sum_{n=1}^N \frac{|\langle \mathbf{U}_n, \mathbf{W} \rangle|^2}{\sqrt{|\tau_n|}} < +\infty \right\}, \quad (3)$$

where $\langle \mathbf{U}_n, \mathbf{W} \rangle := \overline{\mathbf{U}_n} \cdot \mathbf{W}$, τ_n are the singular values, and $\mathbf{U}_n$ are the left singular vectors of $\mathbb{K}$. Based on the structure of (2), we introduce a unit test vector, $\mathbf{W}(\mathbf{r})$, expressed as follows:

$$\mathbf{W}(\mathbf{r}) := \frac{\mathbf{V}(\mathbf{r})}{|\mathbf{V}(\mathbf{r})|},$$

where

$$\mathbf{V}(\mathbf{r}) = \left[\mathbf{E}_{\text{inc}}(\mathbf{a}_1, \mathbf{r}), \mathbf{E}_{\text{inc}}(\mathbf{a}_2, \mathbf{r}), \dots, \mathbf{E}_{\text{inc}}(\mathbf{a}_N, \mathbf{r}) \right]^T.$$

Then, based on (3), the imaging function of the linear sampling method can be introduced as follows:

$$\mathfrak{F}_{\text{LSM}}(\mathbf{r}) = \left(\sum_{n=1}^N \frac{|\langle \mathbf{U}_n, \mathbf{W}(\mathbf{r}) \rangle|^2}{\sqrt{|\tau_n|}} \right)^{-1}.$$

Because $\mathbf{W}(\mathbf{r}) \in \text{Range}(\mathbb{A}^{1/4})$ when $\mathbf{r} = \mathbf{r}_D$, the value of $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$ will be close to zero when $\mathbf{r} \notin D$ and nonzero when $\mathbf{r} = \mathbf{r}_D$. Hence, the location of D can be identified by plotting $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$.

3. Simulation results

A number of simulation results is shown as a demonstration of the concept of the linear sampling method. For this, we use 16 dipole antennas and apply a $f = 1$ GHz frequency. All S -parameters $S_{\text{scat}}(m, n)$ are generated using the CST STUDIO SUITE. To perform the linear sampling method, we set the search domain Ω to $\Omega = [-0.1, 0.1] \times [-0.1, 0.1] \text{m}^2$. For each $\mathbf{r} \in \Omega$, the step size of $\mathbf{r}$ is considered to be of the order of 0.001m. Material properties, locations, and sizes of anomalies as well as backgrounds are summarized in Table 1.

Table 1. Electromagnetic properties, locations, and sizes of anomalies as well as backgrounds

Target	Relative permittivity	Conductivity (S/m)	Location	Size (diameter)
B: background	20	0.2	—	—
D: anomaly	55	1.2	(0.01, 0.03)	20 mm
S: skin	35	0.3	—	2-mm thick
C: cylinder	15	0.5	(0.01, 0.02)	96 mm

Example 3.1 (Detection of a small anomaly). Figure 1 shows the test configurations and the map of $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$ for detecting a small anomaly, D , where S -parameters for the incident and total fields are generated by the settings shown in Figures 1(a) and 1(b), respectively. From the map of $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$, we can identify the exact location of D and determine whether the linear sampling method is applicable for the detection of small anomalies.

Example 3.2 (Detection of an extended anomaly). Now, we employ the linear sampling method for imaging the extended anomaly $C \cup S$. The test configurations and the map of $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$ are shown in Figure 2. In this example, S -parameters for the incident and total

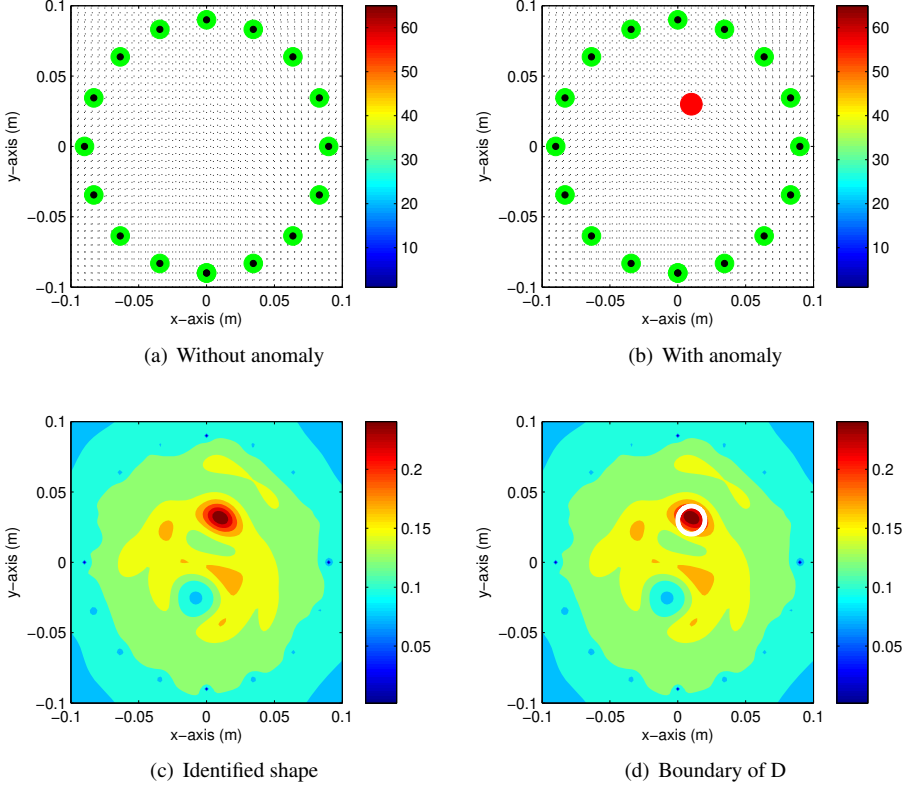


Figure 1. Test configurations with and without an anomaly and a map of $\mathfrak{F}_{LSM}(\mathbf{r})$ at $f = 1$ GHz. The white circle represents the boundary of D .

fields are generated by the settings shown in Figures 2(a) and 2(b), respectively. On the basis of the map of $\mathfrak{F}_{LSM}(\mathbf{r})$, we can observe that although the exact shape of S cannot be retrieved, we can recognize an outline of the shape for $C \cup S$ and verify whether the linear sampling method is an effective algorithm for detecting an extended anomaly.

4. Conclusion

In this study, we designed a linear sampling method for rapid identification of small anomalies with a small number of dipole antennas from collected S -parameter data. Simulation results with synthetic data are shown to demonstrate the effectiveness of the linear sampling method.

In this paper, we focused on the imaging of circle-like small and extended anomalies. Extension to the imaging of crack-like defects will be considered in future research. Finally, it is not possible to extract an anomaly completely concealed within a larger anomaly with a homogeneous background. Hence, an improvement of the current imaging technique is required.

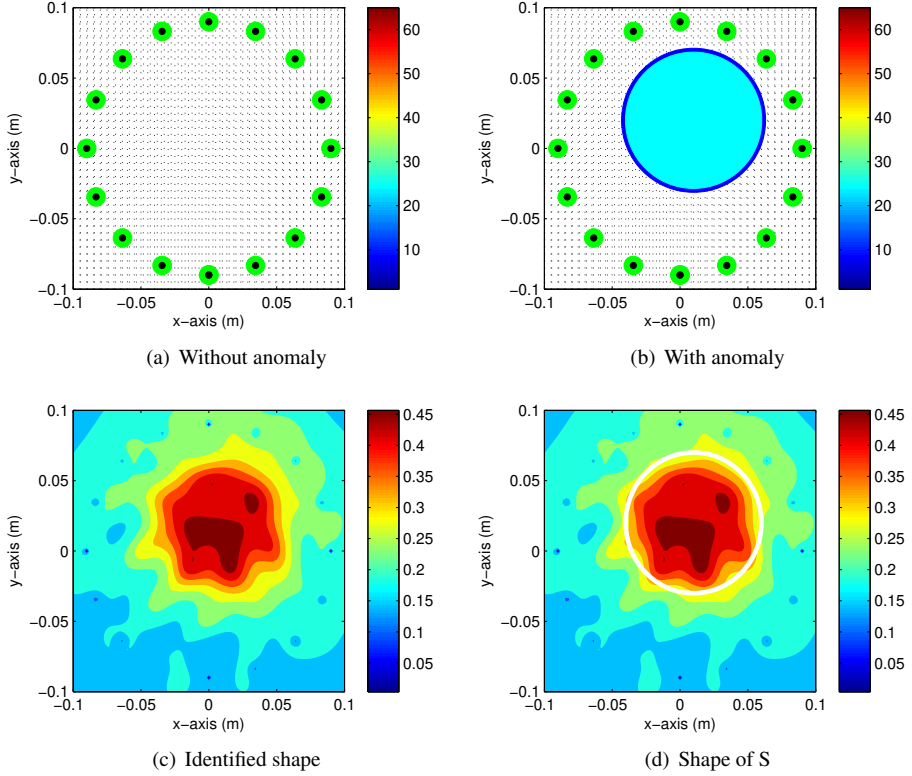


Figure 2. Test configurations with and without the anomaly and a map of $\mathfrak{F}_{\text{LSM}}(\mathbf{r})$ at $f = 1$ GHz. The white circle represents the shape of S.

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Nondestructive Evaluation of Hardening Degree of Epoxy Resin in CFRP with Eddy Current Testing

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Abstract. Chemiluminescence analysis and high-frequency eddy current testing (HF-ECT) were examined as nondestructive evaluation techniques to assess the hardening degree of epoxy resin in carbon fiber-reinforced plastic. Both techniques showed a correlation with the hardening degree of epoxy resin measured by differential scanning calorimetry. The results also indicated that the application of HF-ECT at frequencies of >50 MHz is suited for hardening degree determination.

Keywords. CFRP, high frequency eddy current testing, chemiluminescence analysis, hardening degree, differential scanning calorimetry

1. Introduction

The matrix resin, such as epoxy, play an important role in the mechanical properties of carbon fiber-reinforced plastics (CFRPs). However, complete curing of the matrix is not always achieved. The curing conditions employed thus determine the final mechanical properties [1]. "Uncured regions" may be seen during the manufacturing of rather large or long products. At the heating process in resin transfer molding method or pultrusion method, which are the processes not using prepreg, the uncured regions tend to be also occurred. To avoid the formation of uncured region, manufacturers often perform prolonged heating and/or post curing of the products. However, such processes increase the production cost and cycles.

The hardening degree of resin is typically evaluated by differential scanning calorimetry (DSC). In such a process, a few milligrams of the test specimen are heated, and the transition enthalpy of the specimen is measured. When the specimen is heated under nitrogen atmosphere, the hardening degree α can be determined by Eq. (1);

$$\alpha(t) = \frac{\int_0^t q(t)dt}{\int_0^{t_e} q(t)dt} = \frac{H(t)}{H_T} \quad (1)$$

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where, q is the heat flow (mW), $H(t)$ (mJ) is the enthalpy of the reaction up to elapsed time t (s), t_c is the time at which curing is complete, and H_T is the total enthalpy of the curing reaction [2]. From Eq. (1), the hardening degree of resin varies from 0 to 1, which can be used as a measure of the degree of completeness of the curing process, with 0 indicating that no curing had occurred and 1 indicating that curing is complete. Accordingly, the hardening degree can be used to indirectly evaluate the quality of the product.

However, evaluating the hardening degree via DSC is unsuitable for a comprehensive assessment of the quality of a product because only small portions of the product (rather than the entire product) can be tested. Additionally, DSC is a destructive evaluation technique as the product is lost upon heating. Thus, developing alternative nondestructive evaluation technique for the hardening degree of resin is desirable.

Potential candidates include chemiluminescence analysis (CLA), which involves the measurement of chemiluminescence (CL) photons that is generated by resin upon curing with photo multipliers. As the amount of CL photons generated is linearly correlated to the hardening degree of epoxy resin [3], CLA can be used to assess the hardening degree of CFRPs. In another method, high-frequency eddy current testing (HF-ECT), an electromotive force v is generated on a pickup coil when eddy current is generated on the CFRP using an exciting coil; v can be described by Eq. (2):

$$v = - \int_S \frac{\partial \mathbf{B}}{\partial t} \cdot d\mathbf{S} = -\mu \int_S \frac{\partial \mathbf{H}}{\partial t} \cdot d\mathbf{S} \quad (2)$$

where $\mathbf{B}$ is the magnetic field intensity (Wb/m²), S is the cross section of the pickup coil through which the magnetic flux passes, μ is the magnetic permeability (H/m), and $\mathbf{H}$ is the magnetic field strength (A/m). According to Maxwell's equations, $\mathbf{H}$ and the electric field $\mathbf{E}$ (V) are related as described in Eq. (3):

$$\nabla \times \mathbf{H} = \sigma \mathbf{E} + j\omega\epsilon \mathbf{E} \quad (3)$$

where σ is the electric conductivity (S/m), ω is angular frequency, and ϵ is the permittivity (F/m). Furthermore, $\sigma \mathbf{E}$ is effectively the conduction current $\mathbf{J}_C$ and $j\omega\epsilon \mathbf{E}$ is the displacement current $\mathbf{J}_D$. When discussing eddy current in metallic materials, $\mathbf{J}_D$ is often ignored because it is considerably smaller than $\mathbf{J}_C$. However, when considering insulators such as epoxy, the contribution of $\mathbf{J}_D$ is important in particular in high-frequency regions.

For the case of CFRPs, the epoxy matrix region accounts for more than 40 vol.% of the total volume of the CFRP. Therefore, the contribution of $\mathbf{J}_D$ is considered for interpretation of the eddy current signal. In a high-frequency region, e.g., > 2 MHz, ϵ of epoxy resin can influence the electromagnetic field based on Eqs. (2) and (3). For instance, it is important to consider $|\mathbf{J}_D|/|\mathbf{J}_C|$ ($= \omega\epsilon/\sigma = \Gamma$) values when either carbon fiber or epoxy resin is dominant in determination of Γ . When carbon fiber is dominant, Γ_{CF} is calculated by using values of σ and ϵ_r of carbon fiber, whereas Γ_{EP} is calculated with those of epoxy resin when epoxy resin is dominant.

We briefly assessed the influence of $\mathbf{J}_D$ by calculating Γ_{CF} , Γ_{EP} and a product of them ($\Gamma_{CF} \cdot \Gamma_{EP}$) by assuming that σ values of carbon and epoxy are 1.5×10^4 S/m [4] and 1.0×10^{-12} S/m, respectively. The relative permittivity ϵ_r values ($\epsilon_r = \epsilon/\epsilon_0$, where ϵ_0 is permittivity of vacuum) are 1 and 2, respectively. Table 1 shows the values of Γ_{CF} , Γ_{EP} and $\Gamma_{CF} \cdot \Gamma_{EP}$ at frequencies ranging from 0.5 to 50 MHz. As observed, generally, at lower frequencies (< 2 MHz), the Γ_{CF} value of carbon fiber is comparable with Γ_{EP} the value of epoxy resin. At frequencies ≥ 2 MHz, the $\Gamma_{CF} \cdot \Gamma_{EP}$ value is larger than 1.0, which indicates that $\mathbf{J}_D$ of epoxy resin is important in eddy current. Considering that the actual

electromagnetic properties of CFRPs are more complex, these values can only be used as guidelines for establishing criteria surrounding frequency for eddy current testing.

Table 1. $|J_D|/|J_C|$ ($=I$) values when either carbon fiber (I_{CF}) or epoxy resin (I_{EP}) is dominant and a product of I_{CF} and I_{EP} ($I_{CF} \cdot I_{EP}$).

	0.5 MHz	1 MHz	2 MHz	5 MHz	10 MHz	50 MHz
I_{CF}	2.8×10^{-9}	5.6×10^{-9}	1.1×10^{-8}	2.8×10^{-8}	5.6×10^{-8}	2.8×10^{-7}
I_{EP}	2.8×10^7	5.6×10^7	1.1×10^8	2.8×10^8	5.6×10^8	2.8×10^9
$I_{CF} \cdot I_{EP}$	7.7×10^{-2}	3.1×10^{-1}	1.2×10^0	7.7×10^0	3.1×10^1	7.7×10^2

As curing of the epoxy progresses, the molecular weight of the resin increases and the dielectric constant of the resin decreases [4]. Variations in permittivity can change the electromotive force and coil impedance. Therefore, HF-ECT could be used to evaluate the extent of curing and hardening degree based on variations in the dielectric constant of the cured epoxy.

Gäbler et al. reported the use of HF-ECT at high frequencies (6 MHz) for evaluating the hardening degree of epoxy resin [6]. Measuring variations in the hardening degree of epoxy resin was possible using HF-ECT and a coil with a diameter of 2 mm. Their results indicated that HF-ECT could be used to evaluate the hardening degree of CFRPs. However, it is yet to be determined whether the HF-ECT signal of CFRPs is correlated to the hardening degree.

In the present study, we attempt to evaluate the HF-ECT signal amplitude of CFRPs for comparison with the hardening degree obtained by DSC at high frequencies of ≥ 2 MHz. Additionally, we discuss the applicability of HF-ECT and CLA as nondestructive evaluation methods for assessing the hardening degree of CFRPs.

2. Method

2.1. Preparation and DSC of CFRP

One ply of prepreg (P3252S-25, Toray Industries Inc., Tokyo, Japan) was cut into a certain shape and cured by heating in a furnace at 100°C for 30, 60, and 120 min and at 150°C for 120 min under air atmosphere. The heated specimens were then sealed and stored in a freezer until analysis.

The curing progress of epoxy was evaluated by DSC. A few milligrams from various parts of the test specimen were sampled and subjected to DSC analysis (DSC-6100, Seiko Instruments Inc., Chiba, Japan). The samples were heated at a rate of 5°C/min from 50°C to 200°C, and the exothermic heat flow (mW) was measured. If the epoxy is not perfectly cured, the curing reaction heat is generated. The transition enthalpy ΔH (mJ/mg) of the CFRP can then be measured using the heat flow and weight of the sample. In the present paper, the hardening degree α of epoxy resin is described as follows:

$$\alpha = \frac{\Delta H_p - \Delta H_i}{\Delta H_p} \quad (4)$$

where, ΔH_p is the total reaction enthalpy of the prepreg (uncured CFRP) and ΔH_i is the enthalpy of the total reaction of the heated CFRP.

2.2. Chemiluminescence Analysis of CFRP

The CL intensity (photon counts per second, cps) of the specimens was measured using a chemiluminescence analyzer (CLA-FS4, Tohoku Electronic Industrial Co., Ltd., Sendai, Japan). The specimens ($20 \times 20 \text{ mm}^2$) were placed in a sample chamber filled with nitrogen gas at 150°C , and the CL intensity was measured using photomultiplier tubes. The CL intensity was calculated using the area of the CL curve.

2.3. High Frequency Eddy Current Analysis of CFRP

For HF-ECT, a pancake coil comprising a ferrite core was used. The inner diameter and outer diameter of coil were 1.4 mm and 3.0 mm respectively. The diameter of the wire was 0.2 mm. The number of turns was 38 and the height of the coil was 2.4 mm. The coil liftoff was 0.3 mm. The ferrite core was a cylinder with a diameter of 0.8 mm and a height of 3.0 mm. The surface of the core was arranged on the surface of the probe. In the present study, two resonance peaks were observed on a prepreg and CFRPs using an impedance analyzer (4294A, Agilent Technologies, Inc., Santa Clara, California, USA) (Fig. 1). The first resonance (5.9 MHz) is self-resonance and the second resonance (56.9 MHz) is due to coupling between coil and CFRP. CFRP has inductance and capacitance components when eddy current is generated. Especially, the coil impedance at higher resonance frequency is sensitive to the variation of the permittivity of target material [6]. The probe was placed on the surface of the CFRP specimen, and the real part and imaginary part of the impedance were measured at a constant current of 20 mA. The impedance was calibrated by the impedance value measured in the air.

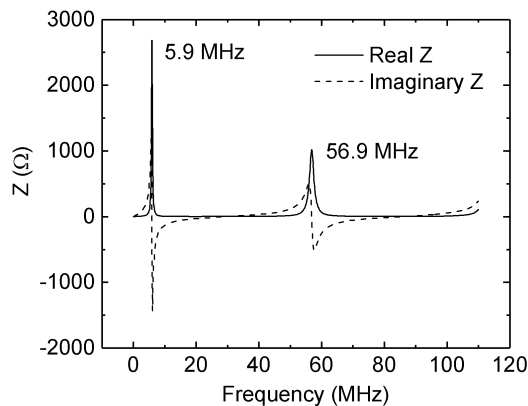


Figure 1. Impedance profile of the pancake coil comprising a ferrite core on a prepreg at 23°C .

3. Results and Discussion

The DSC curves of the heated CFRP specimens are shown in Fig. 2. As observed, prepreg, which was not heated, exhibited a large heat flow during DSC measurement. In

contrast, the specimens that were heated for a long time exhibited a lower heat flow. Specifically, the specimen that was heated for 120 min at 150°C hardly exhibited a heat flow, indicating that the CFRP specimen was perfectly cured under this heating condition. The transition enthalpy values of the specimens, calculated from the DSC curves, are shown in Fig. 3a. The relationship between the hardening degree and the transition enthalpy of the specimens is shown in Fig. 3b. From the latter graph, the relationship between CL intensity and HF-ECT signal and the hardening degree of the CFRPs can be evaluated.

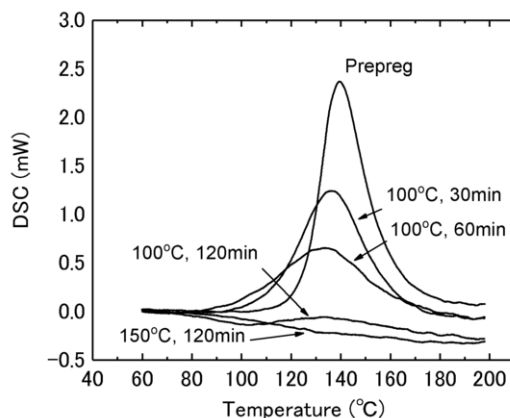


Figure 2. DSC curves of prepreg (not heated) and the heated CFRPs.

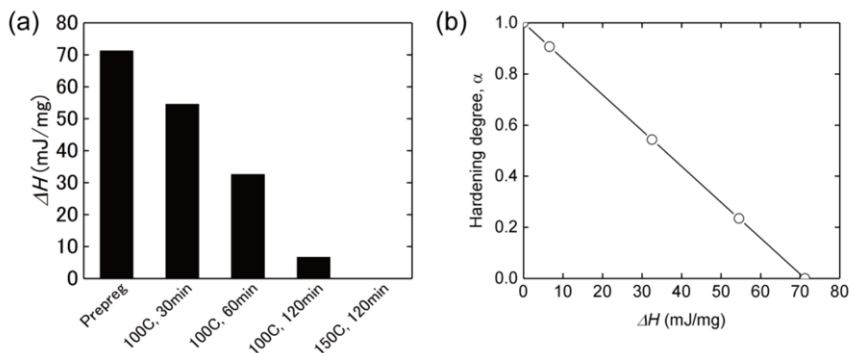


Figure 3. (a) Transition enthalpy ΔH of prepreg and heated CFRPs. (b) Relationship between the hardening degree α and transition enthalpy ΔH of CFRP specimens measured by DSC.

Fig. 4 shows the variations in CL intensity as a function of time. The high CL counts obtained indicate that epoxy is not cured upon heating at 100°C. As heating was prolonged, the CL counts decreased, indicating the progress of curing. Fig. 5 shows the relationship between the average CL intensity obtained over 8 min of measurement and the hardening degree obtained by DSC. The obtained correlation indicates that CLA can be used as a nondestructively evaluating technique for assessing the hardening degree of CFRPs.

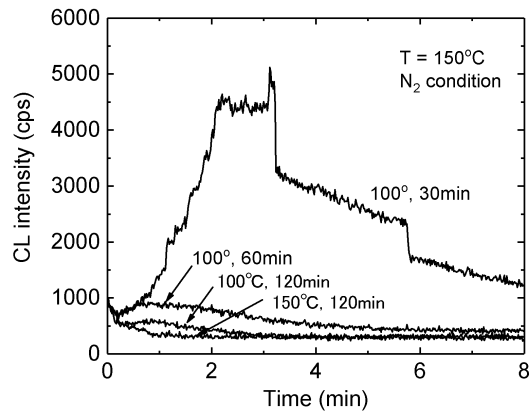


Figure 4. Variations in CL intensity as a function of time of the heated CFRPs in N₂ atmosphere.

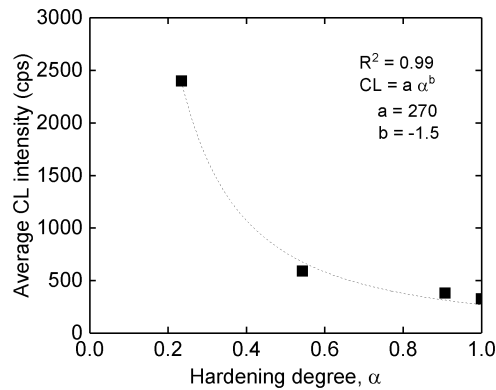


Figure 5. Relationship between the average CL intensity obtained over 8 min of measurement and hardening degree.

Fig. 6 shows the coil impedance diagram of the CFRPs at varying frequencies from 54 to 60 MHz obtained by HF-ECT. The phase of the impedance plot changed according to the frequency. Variations in the impedance Z ($Z = \{(\text{Real } Z)^2 + (\text{Imaginary } Z)^2\}^{1/2}$) became larger as the frequency approached the resonance and reaches a maximum at 56.9 MHz before becoming smaller as the frequency increased above the resonance.

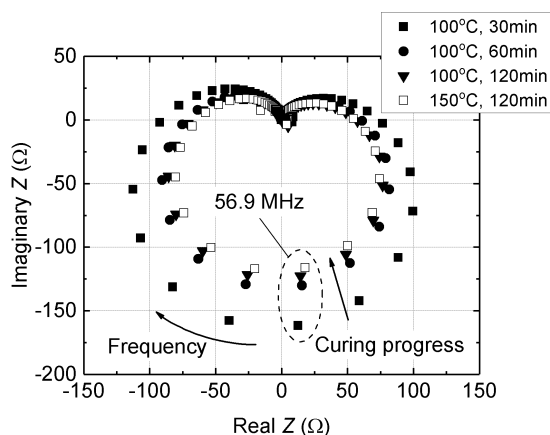


Figure 6. Impedance diagram of the heated CFRPs at frequencies ranging from 54 to 60 MHz.

Fig. 7 shows the relationship between the coil impedance Z and the hardening degree. As observed, the coil impedance of CFRP decreased as the hardening degree (which is a measure of the extent of curing) increased. The same relationship was reported for epoxy resin by Gäbler et al. [6]. Thus, HF-ECT can also be used as a nondestructive evaluation method for assessing the hardening degree of CFRP matrices.

The relationship between coil impedance and hardening degree was similar to that obtained between CL intensity. Both relationships could be well fitted with the allometric function $y = ax^b$. However, it is yet to be determined why the relationships approximate to an allometric function. Regardless, the two methods examined herein can be used as nondestructive evaluation tools. CLA can be conducted in the absence of probes, which are typically in close proximity to the test specimens in conventional methods. However, CLA should be carried out in a dark room for precise detection of CL. Therefore, evaluation of the hardening degree with CLA has some constraints. In contrast, HF-ECT should be performed at a constant temperature and liftoff of the probe should be well controlled to avoid interference with the impedance measurement. Thus, though CLA and HF-ECT can be used as nondestructive techniques for evaluating the hardening degree of CFRPs, the conditions required for the measurements differ. Selection of the technique is therefore dependent on the user's needs. Additionally, results obtained from CLA are only representative of the surface of the test specimen. In contrast, in HF-ECT, it is unclear how representative the extracted information is in relation to the depth of the specimens examined because eddy current including displacement current in CFRPs are complicated and not well understood to date. However, numerical analysis and precise experiment can be useful for the interpretation and quantitative calculation of the effect. Understanding the mechanism of eddy current in CFRPs is one of the main fields of research in electromagnetics.

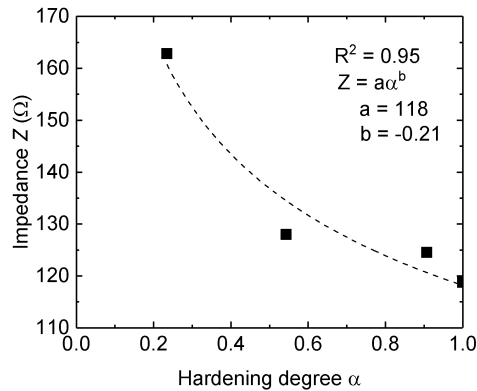


Figure 7. Relationship between coil impedance obtained at 56.9 MHz and hardening degree.

4. Conclusion

Both chemiluminescence analysis and high-frequency eddy current testing techniques show the correlation with the hardening degree of CFRP measured by DSC. Similar variations in the coil impedance of the CFRPs examined and epoxy resin were observed at the resonance frequency. By measuring the coil impedance at the resonance, it was possible to assess the hardening degree of the CFRPs by eddy current testing. The present results indicate that both chemiluminescence analysis and high-frequency eddy current testing techniques can be applied as nondestructive evaluation techniques for assessing of the hardening degree of CFRPs.

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Eddy Current System for Clad Pipe Inspection

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Abstract. Orthogonal eddy current sensors operating in differential mode was applied to evaluate fatigue cracks in clad pipelines circumferential welds. A dedicated electronic hardware was developed to drive the sensors and measure the electrical impedance. In the preliminary experiments, an automated inspection was performed with the goal to evaluate sensors detectability and different scanning speed was tested to reproduce in service situation. The results have shown that the presented eddy current transducer is a potential solution for a fatigue crack detection on a clad circumferential weld bead.

Keywords. Clad material, fatigue crack, eddy current testing

1. Introduction

The application of clad material to subsea pipelines is gaining ground in deep water oil exploration. Its bimetallic configuration presents an attractive combination of mechanical strength and corrosion resistance, ensuring the safety and integrity of pipelines that connect the reservoir to oil rig. The clad material for oil exploration consists of a base material, usually carbon steel, inner coated with a thin layer of corrosion resistance alloy (CRA), turning into an attractive economical solution for deep water exploration since only a small portion of the noble anti-corrosive alloy is required. Clad material has a metallurgical bond between the CRA and the base material attained by the carbon diffusion during the hot rolling process [1].

The potential for fatigue cracks to occur in pipeline structures due to cycling loads inherent of offshore oil production, such as, tide variation, waves, ocean current, platform movements, etc., makes necessary have an inspection tool to carry out periodic non destructive inspection in the inner pipe surface. In case of clad material, it is crucial to detect fatigue crack on its initial stage because if the crack propagates through the layer of the CRA and reaches the carbon steel a strong galvanic couple is completed

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accelerating exponentially the fatigue corrosion process [2]. The most critical point of pipeline structures is the circumferential weld [3] and demands special attention during inspection. Figure 1 presents a section of clad pipeline with a base material of carbon steel API X65 coated with Inconel 625 and highlighted the inspection region with the crack position.

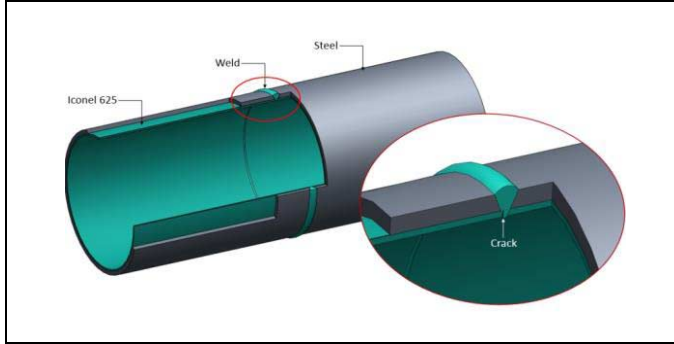


Figure 1. Clad pipe with the inspection region highlighted.

This scenario, detection of fatigue crack in Inconel, encouraged the development of an eddy current (EC) system with the aim to inner inspect clad pipelines. The main challenges are the circumferential weld geometry and the fact that the tool must operate in a speed around one meter per second. The goal of the present work is evaluate the capability of EC transducer to detect fatigue cracks on circumferential weld bead root of Clad pipelines. An orthogonal EC transducer was tested and an electronic hardware was built to drive the transducer and measure the testing coils electrical impedance.

2. Experimental set-up

Two clad plates were manufactured with a 45 degree bevel to receive a weld bead from GTAW weld process. EDM notches with 10.0x1.5x0.2mm were machined in the Inconel side between the weld root and the Inconel base material. Figure 2 presents a photo of the testing sample and a metallographic image of the weld cross section after mechanical grinding, polishing and etching with chromium nitride solution.

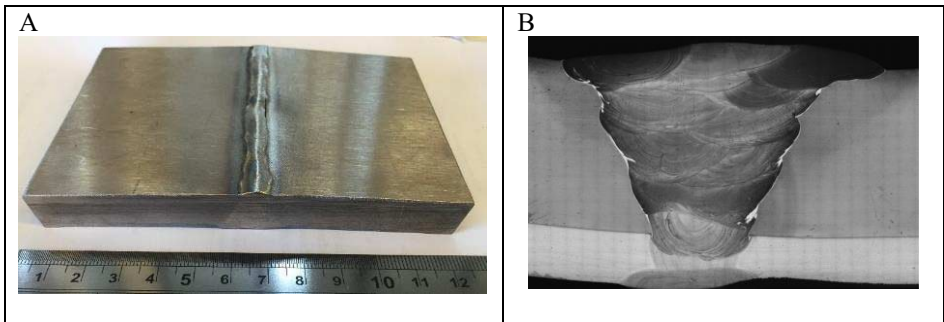


Figure 2. A) Clad sample with the weld bead and EDM notch and B) metallographic image of the weld cross section.

The testing coils manufactured to inspect the notch present an orthogonal configuration with the layers interweaved as presented by figure 3A. The coils are differentially connected thereby reducing spurious signals caused by lift-off variation [4, 5]. When compared with EC pencil probes, orthogonal coils present low lift-off sensitivity, reducing lift-off variation from 40 dB/mm to 8 dB/mm. For weld inspection orthogonal coils configuration present significant results because spurious signals arising from some specific materials characteristics or from some physical structures that are common to both coils are annihilated thereby providing no undesirable response. The transducer testing frequency was 400 kHz and to assist the inspection a KUKA robotic arm model Hollow Wrist with KRC4 controller was used. With a payload of five kilograms, the robotic arm carry the sensors and tests different inspection speed from 0.05 m/s to 1.0 m/s.

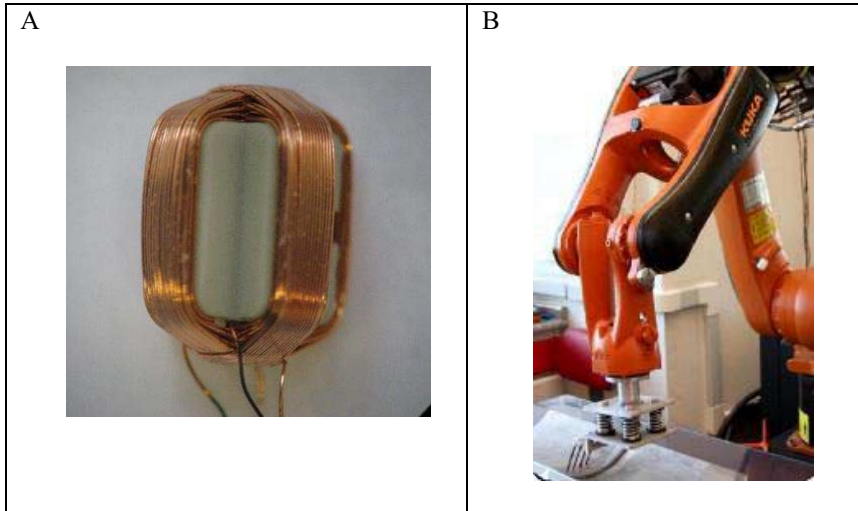


Figure 3. A) Orthogonal coils with the layers interweaved and B) KUKA robotic arm for automated inspection.

The electronic hardware drives the sensors and measure the electrical impedance. Figure 4 presents the block scheme of the measuring system. Initially the algorithm implemented in the microprocessor to calculate the real and imaginary part of the coil impedance was the Fast Fourier Transform (FFT) but later in order to improve the hardware time processing Goertzel algorithm replaced FFT. Goertzel algorithm is an efficient evaluation of individual terms of the Discrete Fourier Transform (DFT) [6]. Equation 1 presents the computed DFT term for the input sequence $x[n]$ in the frequencies chosen ω_0 for the Goertzel analysis. The index k indicates the frequency bin of the DFT.

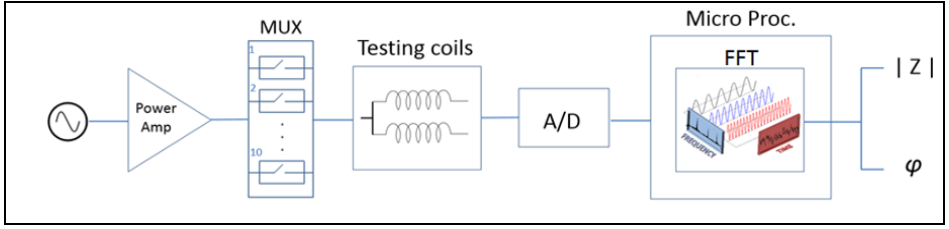


Figure 4. Schematic block of the hardware to drive the eddy current coils and evaluate the impedance variation.

$$y[N] = \sum_{n=0}^N x[n]e^{-j\omega_0 n} \quad (1)$$

where: $\omega_0 = 2\pi \frac{k}{N}$ and $k \in \{0, 1, 2, \dots, N-1\}$.

The use of Goertzel algorithm presented a significant improvement in the hardware data streaming. When compared with FFT, Goertzel presented a calculation speed six times faster, however once it calculates only a single term of the DFT some relevant information is lost, such as, the total harmonic distortion (THD) which gives evidences about the behavior of the coil input excitation signal. If the input signal saturates the A/D converter it can be easily noticed by the distortion caused in the FFT spectrum. Finally, with the faster calculation and less data in the streaming package (THD data is only calculated in the FFT version) the total data rate of the electronic hardware increased from 50Hz to 100Hz with Goertzel. Figure 5 presents two oscilloscope images of the impedance calculation and data transmission time of both algorithms.

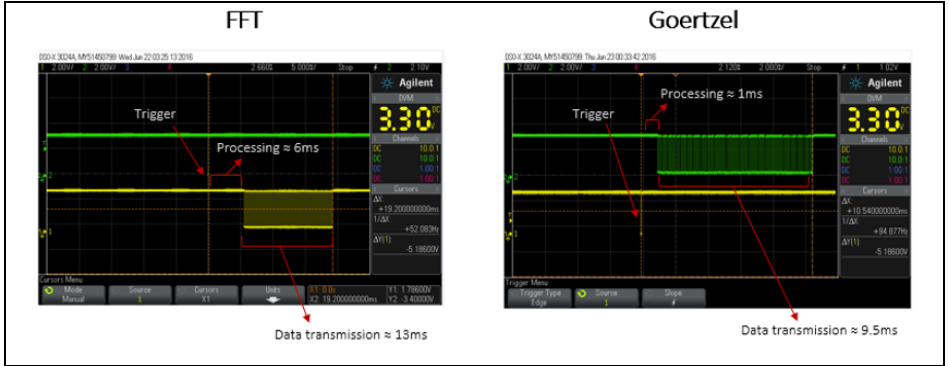


Figure 5. Oscilloscope images showing the time processing achieved by FFT and Goertzel algorithm.

3. Results

The testing sample was scanned with orthogonal EC transducer and different scan speed. Figure 6 presents the automated tests with EC orthogonal transducer on a clad sample and the inspection results of the resistance and inductive reactance. One can observe that

the notch indication was clearly separated from the weld bead on both impedance components, presenting a SNR of 14 dB.

The inspection speed is an important parameter for the tests because once the inspection tool carrying the sensors navigates inside the pipe it is propelled by the oil and the oil flow determines the tool speed in field situation. Therefore, a fast inspection tool results in a shorter inspection process, which means higher oil production during the inspection and a shorter interference in the production original course. Figure 7 presents different inspection speed of a single transducer passing over the weld and the notch. The presence of the notch amplifies the signal resistive component and in case of the inductive reactance a second peak in the weld signal is verified, being very prominent in the low speed tests of: 0.05, 0.1 and 0.15 m/s. However when the inspection velocity is increased such variation is not so notable. For high speed tests, 0.5 and 1.0m/s, it is possible to notice a different behavior in the resistance and reactance signal. The resistance signal is attenuated mainly because of the intense physical impact of the transducer in the weld root penetration, which is negligible in the lower speed tests, besides that, the notch peak in the inductive reactance signal start to merge in the weld signal due the fact of less inspection data. It is important mention that the hardware data rate was the same for all speed condition, 100 Hertz, but as the speed is increased the test longitudinal resolution decreases. At 1.0m/s, the red arrow indicates a smooth deformation in the weld signal caused by the EDM notch.

The hardware data rate is a relevant parameter for high speed tests, the implementation of Goertzel algorithm enhanced the processing time of the impedance evaluation, however the data streaming still need to be improved to achieve a better notch identification at 1.0 m/s.

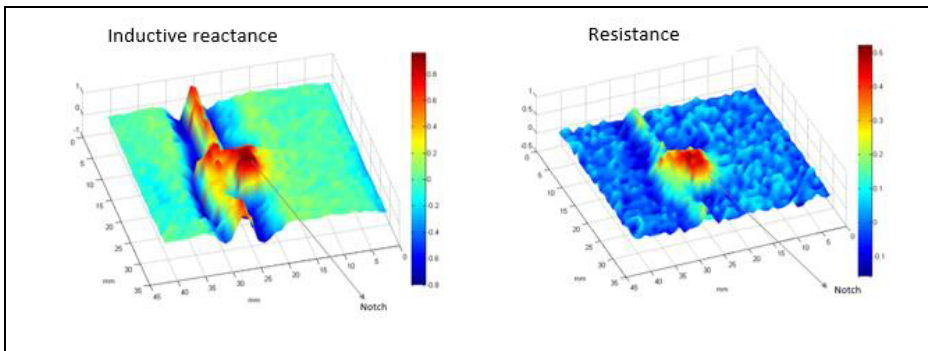


Figure 6. Inspection result of the clad sample with the notch besides the weld bead.

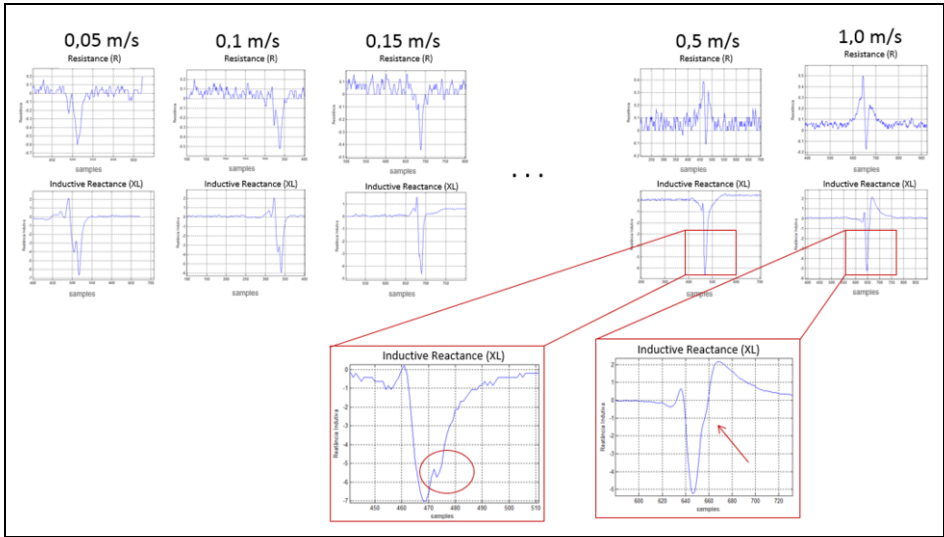


Figure 7. Different speed tests inspection. In red is highlighted the notch indication in velocities of 0.5 and 1.0 m/s.

4. Conclusions

The weld root of clad samples were inspected with orthogonal EC coils and was possible to clear identify the EDM notch.

A dedicated hardware was developed and the Goertzel algorithm was implemented to improve the hardware data rate, and, seems to be a good alternative for high speed inspection tools where only a single frequency can be evaluated. The hardware limitation is the data streaming and tests with Ethernet instead USB will be performed.

Different speed inspection was tested assisted by a robotic arm and was noticed that when the speed is increased the notch identification starts to merge with the weld signal requiring special signal processing.

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Fabrication of Imitative Stress-Corrosion Cracks for Electromagnetic NDE Using Synchrotron Deep X-Ray Lithography and Powder Metallurgy Processing

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Abstract. This paper presents advanced methods using the deep X-ray lithography (X-ray LIGA) and the powder metallurgy processing to fabricate imitative Stress Corrosion Crack (SCC) for electromagnetic NDE as needed to be a reference specimen. A pattern of the imitative SCC was formed by using X-ray LIGA with controllable shape and size with precise details of SU-8 material. The prepared pattern was then inserted into mixed-stainless steel and Sn powders in a soft mold by using less pressure filling method. Afterward, the specimens were sintered at 1100°C. The results showed that these promising combined techniques can be utilized to fabricate micro sizes of the imitative SCCs. In addition, an inversion scheme was utilized to evaluate the equivalent conductivity of the specimens. The analysis results revealed that they were appropriated for desired equivalent conductivities of actual SCC detection by an eddy current testing.

Keywords. Imitative SCC, synchrotron, lithography, eddy current testing

1. Introduction

An eddy current testing (ECT) of stress corrosion cracks (SCCs) is often utilized to ensure the structural integrity of key structures in the nuclear power plants (NPPs), oil & gas and petrochemical industries. Lots of SCC reference specimens are necessary for examination as well as related training for NDT personnel. In general, artificial cracks prepared by Electrical Discharge Machining (EDM) are used to be standards of testing. Unfortunately, it is not always possible for the real complicated SCCs. An artificial

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SCC can also be introduced by loading the specimen in a corrosive solution but, this fabrication is difficult to control the profile of the SCC crack. Especially in terms of quantitative evaluation of SCCs, the specified reference SCC specimen become to a very important part of ECT for quantitative evaluation of SCCs.

The SCC has been researched that it exhibits like a conductive crack in viewpoint of ECT for both of simulation and evaluation of SCCs [1, 2]. Therefore, there are a number of studies for development of SCC references by diffusion bonding and laser 3D printing to fabricate ‘imitative’ SCC which is not exactly identical to actual SCCs but provides the same response to ECT signal [3-5]. However, the welding effect and controlling the imitative SCC sizes are the problems for ECT when the specimen is fabricated by using bonding techniques. The metal 3D printing is also needed the high technology, and it is not a cost-effective technique at the present.

The aim of this paper is to develop an alternative cost-effective technique for fabrication of imitative SCCs that can eliminate the welding effect of bonding area, and also can control the sizes of the imitative SCCs. Synchrotron deep X-ray lithography (X-ray LIGA) based on photolithography technique, has been proposed to produce the pattern of the imitative SCCs. The X-ray LIGA has an advantage for fabrication of very small and complex shape specimens with low cost and rapid replication [6]. The imitative SCC pattern was embedded into the mixed metal powders using less pressure filling method and sintering process. Finally, the imitative SCC specimen has been fabricated with controllable sizes and without welding effect zone. The consequences are helpful for the inversion problem to reconstruct SCCs profile using ECT, and contribute to design and reproduce the SCC reference specimen for practical SCC sizing, and for NDT personnel qualification.

2. Materials and Methods

2.1. Metal Powders and Lubrication Binders

Small powder size and spherical shape are important parameters to be selected. Therefore, the gas atomized stainless steel (SUS 316L) powders which the average particles sizes were 1 μm and 10 μm as delivered by supplier were employed in this research. The chemical compositions are shown in Table 1. Additionally, the 99.9% of Tin(Sn) powder with the particles size of 1 μm was used to mix with SUS-316L powder 2 to 15% wt for acceleration of the sintering process with lower temperature by liquid phase forming effect.

Table 1. Chemical composition of SUS 316L powders

SUS 316L		Chemical composition				
Sample size	Fe	Cr	Ni	Mo	Mn	Others
1 μm	68.57	16.5	10.5	2.1	1.45	0.88
10 μm	68.41	16.2	11.0	2.4	1.26	0.73

Binder system for this works was used to hold and maintain the powder into desired shape. The binders should also have good flow characteristics during the compaction of the powder. In addition, they should be removed by heating with temperature lower than sintering temperature. As descriptions, the binder system for less pressure filling method was adopted by using the acrylic liquid-based binders. Duramax binder system including D-3019 as dispersant, mixture between B-1000 and B-1022 as emulsion of

acrylic polymer were used in this work with composition of 1%, 2% and 5% of total volume respectively.

2.2. Fabrication of SU-8 Pattern for Imitative Stress Corrosion Crack

The fabrication of pattern for imitative SCC in this research was based on the deep X-ray lithography (LIGA) technique which demonstrates in Figure. 1(a). The SU-8 2000 (MicroChem, USA), was used as the photoresist material in the process. The imitative SCC as shown in Figure. 1(b) was used to create a mask aligner of pattern. After prepared the SU-8 spread uniformly on the substrate, it was exposed under the mask pattern by LIGA (facilitated by Synchrotron Light Institute of Thailand) with X-ray accumulated dose at $36,000 \text{ mJ/cm}^2$ under synchrotron beam current of 111.4 mA. Then, the SU-8 pattern was developed by solvent washing and baking. Finally, the imitative SCC pattern was finished by taking pattern out from the substrate.

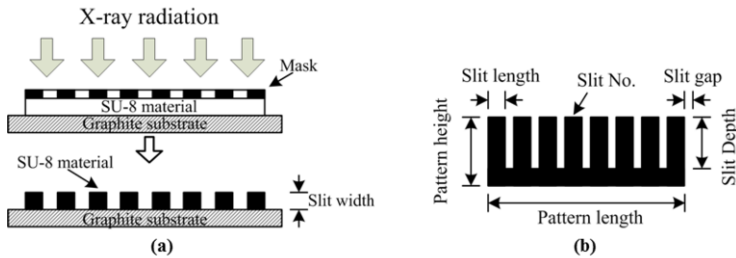


Figure 1. (a) Deep X-ray lithography process, (b) imitative stress corrosion crack pattern

2.3. Powder Metallurgy Processing

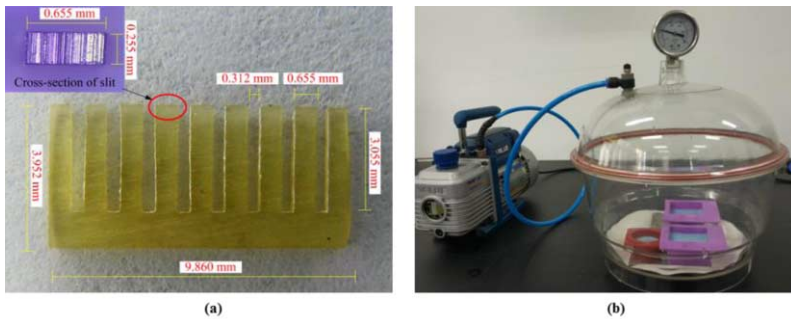


Figure 2. (a) Dimensions of the Imitative SCC pattern, (b) degassing process using vacuum chamber

Fabrication process by less pressure filling method for production of imitative SCC is described as follow: First, the imitative SCC pattern was cut to the length of 10 mm, and shaped to the thickness of about 0.25 mm as shown in Figure 2(a). Next, it was inserted into the soft mold diameter of 40 mm. Both $1 \mu\text{m}$ and $10 \mu\text{m}$ of SUS 316L powders and Sn powder were mixed together by mechanical stirrer for 10 min with weight percentage of $45\% \pm 5\% \text{ wt}$, $45\% \pm 5\% \text{ wt}$ and (2% to 15% wt) respectively. The dispersant with 1% of total volume was mixed with distilled water by using

mechanical method for 5 min. Then, well-mixed powder was added to the dispersant mixture about 50% solid loading. Mechanical stirrer was utilized to blend the solution for 20 min. acrylic binders of B-1000 and B-1022 with 2% and 5% of total volume were added to the whole mixture and continue mixing for 10 min. All of well-mixture was carefully poured to the soft mold then, put it in the degassing vacuum chamber to remove the bubble formation at -0.8 kg/cm^3 until the specimen was dried as shown in Figure 2(b). After that, the green compact specimen in the soft mold was carefully removed by hand. Finally, the specimen was de-binding at 700°C for 90 min and sintering at 1100°C for 90 min in a vacuum furnace with heating rate of 5°C/min .

3. Experimental Results

3.1. Characteristics of the imitative SCCs

Four specimens fabricated by using the X-ray LIGA and less pressure filling method with different composition were grinded and polished at thickness of 5mm to reveal the size of the imitative SCC as shown in Figure 3 by optical observation in macro scales.

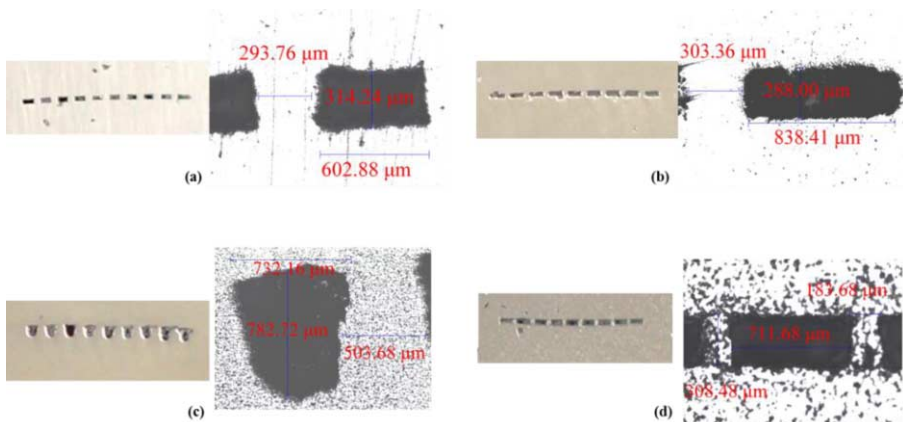


Figure 3. Optical observation in macro scales of (a) TP1 as pure SUS 316L powder, (b) TP2 as 2% of Sn addition, (c) TP3 as 5% of Sn addition, (d) TP4 as 15% of Sn addition

Figure 3(a) as “TP1” reveals shape and size of the imitative SCC resulting from the inserted SU-8 pattern dimensions 0.6 mm of slit length, 0.25 mm of slit width and 0.3 mm of gap between slits. By using only SUS 316L powders, some pores were observed on the surface of the specimen. The measured bulk density was 90% compared to the specific wrought SUS 316L density (7.98g/cm^3). In Figure 3(b), the optical observation result of added 2% of Sn is shown as “TP2”. The shape and size of the imitative SCC of TP2 were resulting from the inserted SU-8 pattern dimensions 0.8 mm of slit length, 0.25 mm of slit width and 0.3 mm of gap between slits. The observed surface of base material appeared better than using only pure SUS 316L powder. The density of the TP2 slightly increased to 91%. However, the results of increasing Sn powder affected to the material property as shown in Figure 3(c) of “TP3”. The shape and size of the imitative SCC of TP3 were obtained from SU-8 insertion of slit length 0.8 mm, slit width 0.8 mm and gap between slits 0.5 mm. Macro observation revealed great

numbers of small pores all over the surface of the specimen. This result of Sn addition at 5% slightly reduced the density of the specimen to 90%. Figure 3(d) as “TP4” additionally shows the effect of very high Sn addition at 15%. The larger pore sizes on the surface were occurrences, and decrease the bulk density to 88%. The size and shape of the imitative SCC were still closed to the SU-8 inserted pattern dimension of slit length 0.7 mm, slit width 0.25 mm and gap between slits 0.2 mm. The average dimensions of all specimens are summarized in Table 2. These dimensions are used to be the parameters for reconstruction of the equivalent conductivity of the imitative SCCs.

Table 2. Average dimensions of the imitative SCC specimen fabricated by LIGA and less pressure filling method

TP No.	Slit No.	Slit Depth (mm)	Slit Length (mm)	Slit Width (mm)	Slit gap (mm)
TP1	10	3	0.6	0.3	0.3
TP2	9	3	0.8	0.3	0.3
TP3	9	3	0.8	0.9	0.5
TP4	9	3	0.7	0.3	0.2

3.2. Microstructures

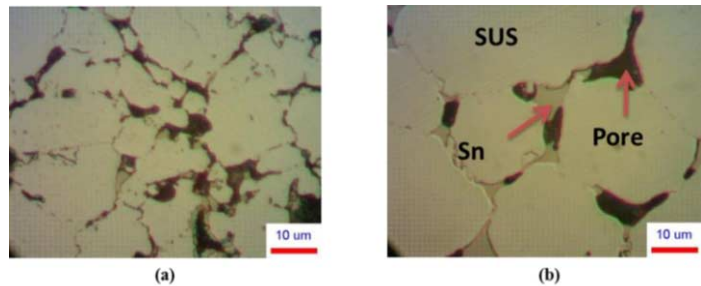


Figure 4. Microstructures of the different composition (a) pure SUS 316L, (b) SUS 316L with Sn powder

Comparison between microstructures of pure SUS 316L and SUS 316L with Sn addition is illustrated in Figure 4. The result of adding Sn powder with stainless steel affected to larger grain in Figure. 4(a) compared to Figure 4(b). Normally, grain growth effect can be observed in the sintered materials for prolonged times. However, larger grain size resulting in SUS316L with Sn addition occurred in normal condition of sintering. This may indicate that the liquid phase of the Sn accelerates the atomic diffusion during sintering process as high sintered temperature or prolonged sintering. On the other hand, the larger pore sizes are also indicated when the percentage of added Sn was increased.

3.3. Eddy Current Testing

There were two ECT experiments, which consisted of the conductivity measurement of the base materials and the imitative SCCs detection. The first experiment conducted to the compare the conductivity of base materials by using the liftoff ECT signals between the reference block and the imitative SCC specimens. The second experiment led to detect the ECT signals from the imitative SCCs area compared to the reference notch at

depth of 3 mm and width of 0.25 mm. The ECT experiments established to measure the ECT signals at test frequency of 50 kHz by using an absolute pancake coil with dimensions of 3.2 mm, 1.2mm and 0.8mm for outer diameter, inner diameter, and height respectively. The employed ECT instrument was the Ectane model of Eddyfi®, Canada. Both testing were run after calibration procedure of the ECT probe and instrument. Results of conductivity measurement are shown in Figure 5(a) by placing the ECT probe on the surface of the base specimens. It is indicated that the measured ECT signals of base material from less pressure filling method for both of TP1 and TP2 are closed to the reference signal of stainless steel block. Especially, the different angle between the TP2 signal and the reference signal is less than 5 degree. Note that, the shorter ECT signals of the TP1 and TP2 caused from the lower density of the specimens. This result imply that the fabricated specimen with SUS 316L plus 2% of Sn powder can represent to be a calibration block as the material similar to the part under examination referred to BS EN ISO 17643:2015 standard [7]. On the other hand, the ECT result of TP4 was much difference due to high admixing of Sn powder at 15%. The possibility may from Nickel composition in the SUS 316L dissolved with Sn, and then caused the conductivity and permeability changes.

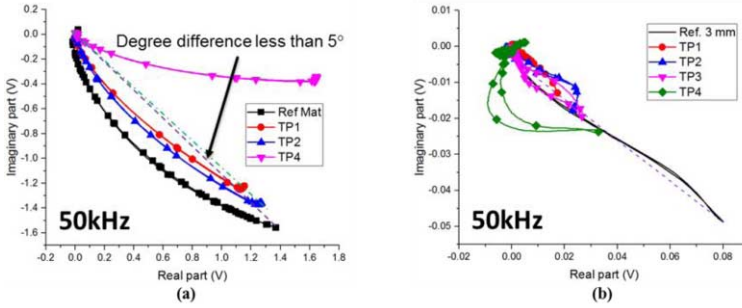


Figure 5. Eddy current testing results (a) conductivity measurement, (b) imitative SCCs detection

Figure 5(b) shows the ECT signals of the imitative SCC detection by scanning the ECT probe across the imitative SCCs compared to the reference notch depth of 3 mm. The results reveal different decreasing amplitudes of the measured ECT signals. Dimensions of multiple tiny slit of the imitative SCCs led to alter the local conductivity in the imitative SCCs. Increasing length of the slit resulted in decreasing equivalent conductivity. Oppositely, decreasing width of the slit resulted as increasing of the equivalent conductivity.

4. Reconstruction of Equivalent conductivity

4.1. Scheme of Equivalent Conductivity Reconstruction

The FEM-BEM fast forward solver was utilized to establish the database of ECT. The formulation as in an Eq. (1) describes the fast forward solver based on database method. Details of expressions can be found in [8].

$$[I - [H_{11}][\tilde{K}_{11}]]\{q_1^f\} = [H_{11}][\tilde{K}_{11}]\{q_1^u\} \quad (1)$$

where the coefficient metric of unflawed field $[\tilde{K}_{11}]$ and inverse metric $[H_{11}]$ are calculated stored as databases. The $\{q_1^f\}$ and $\{q_1^u\}$ are the potential vectors of the flawed and the unflawed fields respectively.

The 3D model used in the simulation had dimensions of $40 \times 40 \times 5$ mm in width, length, and depth respectively. The conductivity of the base material was set to 1.35 MS/m. The imitative SCC area was assumed at the center of the plate with a total length of 10 mm. Scanning path of the ECT probe with similar to experiment parameters was perpendicular to the crack direction. Average dimensions of the imitative SCC specimens measured by optical method as represented in Table 2 were given as parameters to calculate the corresponding ECT signals. An inversion method of the particle swarm optimization (PSO) algorithm was utilized to reconstruct the equivalent conductivity from the measured ECT signals [9]. The discrete objective function by the mean-square residual can be simplified as the Eq. (2).

$$\varepsilon(\sigma_i^k) = \sum_{m=1}^M |z_m(\sigma_i^k) - z_m^{osb}|^2, i = 1, \dots, n \quad (2)$$

where z_m^{osb} is the measured ECT signals, $z_m(\sigma_i^k)$ is the calculated ECT signal when conductivity is σ , M is the number of sampling points, n is the number of particles in a swarm, and σ_i^k is the position vector of the particle (conductivity) i in the k -th iteration. To finish the reconstruction process, the equivalent conductivity was modified to estimate the ECT signals until obtained the convergence result.

4.2. Results of Conductivity Reconstruction

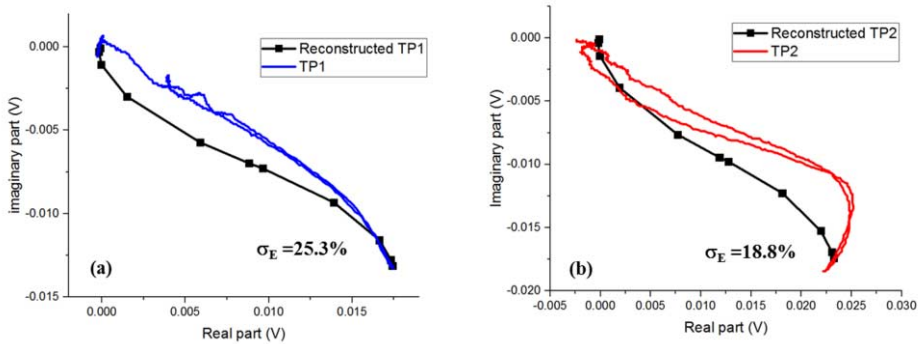


Figure 6. Reconstruction results of equivalent conductivity (a) TP1, (b) TP2, (c) TP3 and (d) TP4

As the results in section 3.3, TP3 and TP4 specimens had different conductivity of the base material compared to the reference SUS 316L. Therefore, only TP1 and TP2 were employed to reconstruct the equivalent conductivity of the imitative SCCs. Figure 6 presents the calculated ECT signals of the imitative SCC specimens with their equivalent conductivities. According to design of the imitative SCCs as the crack with uniform conductivity, the reconstructed results were efficiently done by using the peak

of the ECT signals. The results of equivalent conductivities were 25.3% and 18.8% of the base material for specimens TP1 and TP2 respectively. These results showed that the imitative SCC specimens fabricated by X-ray LIGA and less pressure filling method were reasonable for desired equivalent conductivities of actual SCC detection by the ECT.

5. Conclusion

Synchrotron deep X-ray lithography and powder metallurgy process by using less pressure filling method can be utilized to fabricate complex shapes of the imitative SCCs which provided the high bulk density up to 91%. The fabrication results revealed that shapes and dimensions of imitative SCC had small changes after sintering with few tens micrometer. By testing the imitative SCC specimens in this research, the obtained ECT signals were quiet smooth without noises of welded area. Moreover, the equivalent conductivities of the imitative SCCs were reconstructed by using Particle Swarm Optimization which showed appropriated values to be a desired reference specimen for ECT.

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Evaluation of Electrical Conductivity of Metals via Monotonicity of Time Constants

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Abstract. This manuscript addresses the “classical” problem of estimating electrical conductivity of metals. The novel contribution of this work is the use of a new set of features, namely time constants of the source-free response in a pulsed eddy current testing (PECT) experiment. Time constants characterize the source-free response and increase monotonically with the electrical conductivity of the specimen. Time constants form a particularly important set of features because they do not depend upon the probing system and, hence, they are not sensitive to probe lift-off and tilting which are responsible for significant experimental errors.

Keywords. Electrical conductivity evaluation, pulsed eddy current testing, Nondestructive testing, monotonicity of time constants

1. Introduction

Nondestructive testing (NDT) measurements of the electrical conductivity of conductors is extremely common and useful in many applications. Two distinct families of methods for evaluating the electrical conductivity are available: eddy current and contact (two/three/four points) based methods.

The conductivity of a material can be obtained by measuring the bulk resistance and the physical dimensions of the material, usually of canonical shape [1]. The two-point methods require to apply two terminal/wires to both ends of the material. A voltage source applies a voltage across the material and an ammeter, which is connected in series with the material and the source, measures the current passing through the material. The bulk resistance is then the ratio of the applied voltage and the measured current. Three and four-point techniques are an improvement of this simpler two-points method. For instance, for the four-point probe, a current source forces a current through the ends of the material. A voltage meter simultaneously measures the voltage across the inner part of the material, thus providing the electrical resistance and, hence, the electrical resistivity. This approach avoids the contact resistances affecting the two-point technique and, hence, is widely used for measuring the resistivity of soil, semiconductor and metals. The four-point method is able to measure the electrical

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conductivity of metals with very low uncertainty [2], however, there are several restraining factors for its practical application [3], for instance, it requires good electrical contact with specimen. In addition, a large excitation current is required to produce a sufficiently high potential drop signal for highly conductive materials.

The conductivity of a material can also be determined by measuring the eddy current induced when the material is exposed to an external alternating magnetic field. In practice, a primary external field is usually produced by a coil carrying an AC current. The magnetic flux density due to the induced eddy current can be measured via the voltage induced in a pick-up coil or by using a field sensor (Fluxgate, Hall effect sensor, magnetoresistance sensors, etc.). Compared to the four-point technique, the eddy current based method requires no electrical contacts and, hence, is suitable for high speed operation. The conductivity of the specimen is determined via calibrated standards or by solving an inversion problem. Also, eddy current is a rather localized phenomenon and may reveal the local property.

Conventional eddy current methods use single-frequency magnetic field as excitation. For non-ferromagnetic materials, a well-design eddy current system can measure the electrical conductivity with very small uncertainty [2, 4, 5]. However, conventional eddy current measurement is a function of many operation parameters, e.g. lift-off, frequency, sensor tilting, and requires extra effort to reduce these effects [6]. Pulsed eddy current method has recently emerged as a promising modality since it provides rich information in harmonic context. Current studies mostly investigate the time-domain signatures [3, 7] and the measured signal suffers from the effect of operation parameters. In [8], Ye etc. proposed using the decay time, which is insensitive to sensor lift-off, as a proper feature to evaluate minor changes in electrical conductivity of materials. This paper introduces a new set of features, the time constants, for estimating the electrical conductivity of a material. Time constants have been successfully introduced in Pulsed Eddy Current Imaging of defects in a conducting material ([9] - [12]). The present paper is the first contribution where time constants are exploited for estimating the electrical conductivity of a material. Time constants are attractive because they are monotonic in some sense w.r.t. the spatial distribution of the electrical conductivity [9]. In the specific problem addressed in this paper, in addition to being monotonic, time constants are also linear w.r.t. the electrical conductivity. This makes them very attractive for electrical conductivity measurement. Moreover, they depend only on the specimen geometry and the related conductivity. This set of features is completely independent of any operation parameters such as, probe lift-off, tilting, shape and nature, excitation waveforms, etc.

2. Monotonicity of time constants

In this paper, we consider an eddy current system comprising a planar coil and a controlled current source. The measurement is the voltage across the coil when a prescribed and known excitation current is injected into the same coil (Figure 1). Hereafter we assume linear and non-magnetic conductive materials.

The system is operated in the time domain. The source is periodically switched on and off. The free response of the system, i.e. the response of the system when the source is switched off, is characterized by a proper set of time constants:

$$\mathbf{J}(\mathbf{r}, t) = \sum_{i=1}^{+\infty} c_i \mathbf{J}_i(\mathbf{r}) e^{-t/\tau_i} \quad (1)$$

$$\mathbf{B}(\mathbf{r}, t) = \sum_{i=1}^{+\infty} c_i \mathbf{B}_i(\mathbf{r}) e^{-t/\tau_i} \quad (2)$$

$$v(t) = \sum_{i=1}^{+\infty} c_i v_i e^{-t/\tau_i} \quad (3)$$

where $\mathbf{J}$, $\mathbf{B}$ and v represent the eddy current density, the magnetic flux density and the pick-up coil voltage, respectively, and the c_i 's are linear coefficients. It is also worth noting that, thanks to the Biot-Savart law and the Faraday's law, all measured quantities in eddy current testing share same set of time constants. It can be proved that the τ_i 's are real, positive, bounded and approach to zero when ordered in decreasing order, as usually done. Moreover, the time constants depend monotonically upon the electrical conductivity [9, 10, 11].

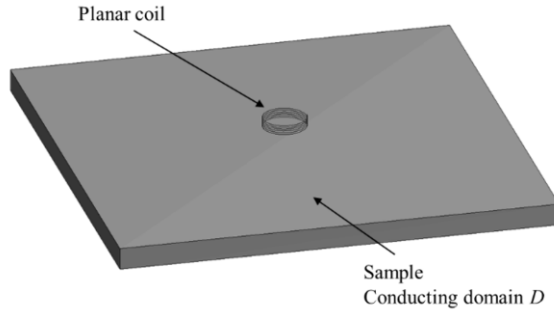


Figure 1. Schematic of an eddy current testing system.

2.1. Natural modes and time constants

Assuming that the conductive specimen is linear, spatially and temporally non-dispersive, isotropic and neglecting the displacement current, the eddy current density $\mathbf{J}(\mathbf{r}, t)$ satisfies the following time dependent equation:

$$\int_D \mathbf{J}'(\mathbf{r}) \cdot \sigma^{-1}(\mathbf{r}) \mathbf{J}(\mathbf{r}, t) dV = -\partial_t \int_D \mathbf{J}'(\mathbf{r}) \cdot \mathbf{A}_i[\mathbf{J}(\mathbf{r}, t)] dV - \partial_t \int_D \mathbf{J}'(\mathbf{r}) \cdot \mathbf{A}_c(\mathbf{r}, t) dV, \forall \mathbf{J}' \in \hat{H} \quad (4)$$

where D is the conducting domain, $\hat{H} = \{\mathbf{v} \in H(\text{div}; D) \mid \nabla \cdot \mathbf{v} = 0, \mathbf{v} \cdot \hat{\mathbf{n}} = 0 \text{ on } \partial D\}$, $\mathbf{J} \in \hat{H}$ for any given t , σ is the electrical conductivity of the conductor and $\mathbf{A}_c(\mathbf{r}, t)$ is the magnetic vector potential due to the driving currents and the operator $\mathbf{A}_i$, acting on the spatial coordinates only, is defined as:

$$\mathbf{A}_i[\mathbf{J}] = \frac{\mu_0}{4\pi} \int_D \frac{\mathbf{J}(\mathbf{r}', t)}{\|\mathbf{r} - \mathbf{r}'\|} dV'. \quad (5)$$

The discrete equivalent of (1) is given by [13, 14]:

$$\left(\mathbf{R} + \frac{d}{dt} \mathbf{L} \right) \mathbf{T}(t) = \mathbf{s}(t), \quad (6)$$

where $\mathbf{T}(t)$ is the electric vector potential and

$$R_{ij} = \int_D \nabla \times \mathbf{N}_i \cdot \sigma^{-1} \nabla \times \mathbf{N}_j dV, \quad (7)$$

$$L_{ij} = \frac{\mu_0}{4\pi} \int_D \int_D \frac{\nabla \times \mathbf{N}_i(\mathbf{r}) \cdot \nabla \times \mathbf{N}_j(\mathbf{r}')}{\|\mathbf{r} - \mathbf{r}'\|} dV dV', \quad (8)$$

$$s_i(t) = -\frac{d}{dt} \int_D \nabla \times \mathbf{N}_i(\mathbf{r}) \cdot \mathbf{A}_c(\mathbf{r}, t) dV, \quad (9)$$

and $\mathbf{N}_i$ is the i -th edge-element shape function.

The natural modes in pulsed eddy current testing are source-free solutions of (3) of the exponential type, i.e. $\mathbf{T}(t) = \mathbf{u}e^{-t/\tau}$. These solutions can be found by solving the following generalized eigenvalue problem [9]:

$$\mathbf{L}\mathbf{u} = \tau\mathbf{R}\mathbf{u}. \quad (10)$$

2.2. Monotonicity of time constants with respect to the electrical conductivity

It is worth noting that if we replace $\sigma(\mathbf{r})$ with a scaled version $\alpha\sigma(\mathbf{r})$, where α is a positive constant, then (10) admits the same generalized eigenvectors, whereas the corresponding generalized eigenvalues (the time constants) are scaled by α , i.e.

$$\begin{aligned} \sigma(\mathbf{r}) &\rightarrow \alpha\sigma(\mathbf{r}) \\ \mathbf{u} &\rightarrow \mathbf{u} \\ \tau &\rightarrow \alpha\tau. \end{aligned} \quad (11)$$

This makes the estimation through time constants very attractive because, from the practical point of view, this requires calibrating the measurements system onto a single reference specimen only. This is not the case when a nonlinear relationship between the measured feature and the electrical conductivity is observed, as in “traditional” eddy current methods.

2.3. Estimation of time constants

Estimation of time constants τ_i 's from the time domain waveform $v(t) = \sum_{i=1}^{+\infty} \beta_i e^{-t/\tau_i} + \xi(t)$, where $\xi(t)$ is the random noise, falls into the category of exponential analysis. In [15], Istratov gave a comprehensive review on available algorithms for extracting the exponential depending on how many exponential terms are subject to find (single exponential, multi exponential and exponential spectrum).

The problem of estimating time constants is naturally classified as multiexponential analysis. In this study, an algorithm modified from the Laplace-Padé approximant approach is exploited to estimate the time constants. The original Laplace-Padé approximant approach was presented in [16], which renders the benefit that it does not require any *a priori* knowledge of the number of exponential terms. A modified approach tailored for estimation of time constants is presented in [12]. This approach takes into account multiple measurements sharing same set of time constants and introduces new stopping criteria accounting for the noise level and the presence of non-physical (negative or complex) time constants.

3. Validation

A numerical model was built to calculate the time constants of a square plate (figure 1). A circular coil above the center of the plate was used to produce the primary field and to measure the response produced by the induced eddy current density.

The width, length and height of the plate is 76.2 mm, 76.2 mm and 3 mm, respectively. The numerical model was discretized into a $30 \times 30 \times 5$ mesh. The cross section of the circular coil is $1\text{mm} \times 1\text{mm}$ and its average radius is 5.5 mm. The lift-off between the coil and the plate is 0.5 mm.

The applied waveform was a train of square pulses of duration 1ms and duty cycle of 50%. We assumed the conductive plate to be homogeneous, and the baseline conductivity was $\sigma_0 = 5 \times 10^6$ S/m. The actual conductivity of the specimen varied from -10% to 10% of this baseline value. The source-free responses (pick-up coil voltages) corresponding to this range are given in figure 2(a). Time constants are numerically calculated according to (10) and are showed in figure 2(b). As expected from the theory, the time constants are linear with the electrical conductivity of the specimen.

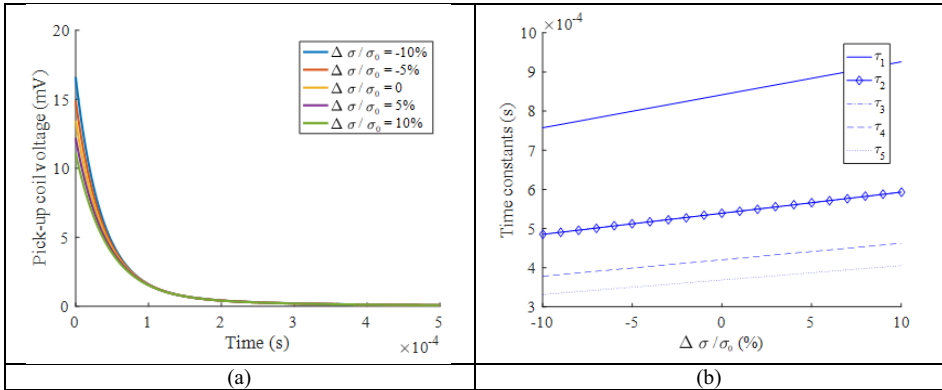


Figure 2. Voltages at the pick-up coil (a) and corresponding time constants (b) as a function of the electrical conductivity. The 2nd and 3rd time constants are identical due to the geometrical symmetry of the plate.

The transient signals in figure 2(a) were sampled at discrete time $t=k\Delta t$ where $\Delta t=2.5 \times 10^{-7}$ s. A total of 2000 samples per waveform were collected ($1\text{ms} \times 50\%$ duty cycle/ Δt). Assuming there is no noise on the measured waveforms, the first three time constants extracted from the signals by the modified Laplace-Padé approximant approach [12] are presented in figure 3.

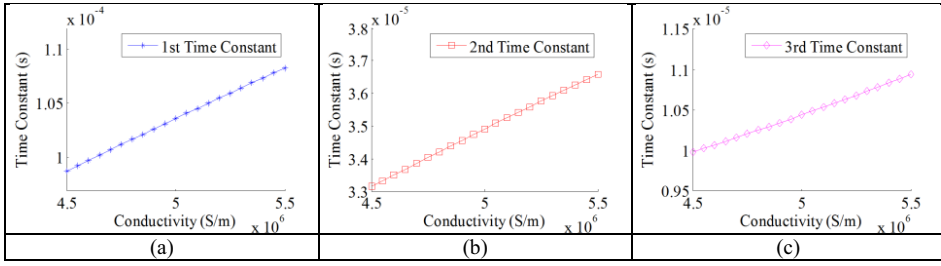


Figure 3. Changes in extracted time constants (noise free) with electrical conductivity.

Assuming an additive noise of about 1% of the peak of the signal amplitude, we performed the extraction of time constants on the signal obtained after averaging a total of 10,000 waveforms. Given 1ms for measuring an individual waveform, the total data collection time is 10s for 10,000 waveforms, thus making the average process feasible. The estimated time constants are presented in figure 4.

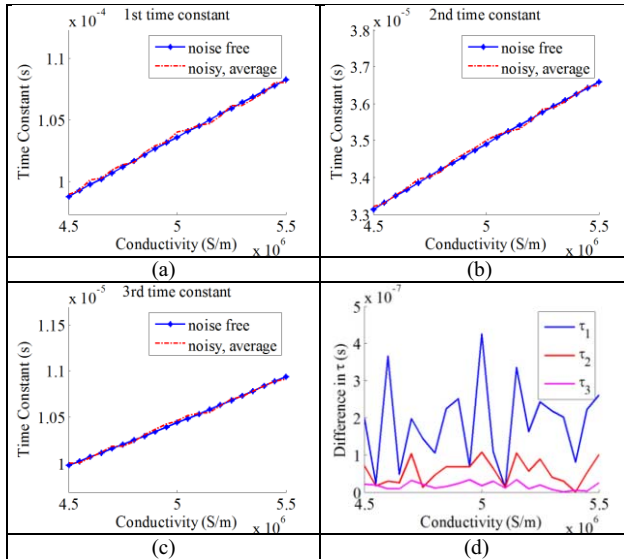


Figure 4. The extracted time constants (noise level = 1%) as a function of the electrical conductivity. Time constants from a noisy signal and from a noise-free signal (a)-(c). Errors (less than 1%) affecting the estimated time constants (d).

3.1. Sensitivity to lift-off

In principle, time constants are only a function of the electrical conductivity and geometry of the specimen. Time constants are independent upon probe lift-off and tilting, shape and nature of the probe and excitation waveforms. Despite of this, we found a weak dependency of the time constants upon the lift-off. There are two possible reasons for this: (i) when changing the lift-off, a different subsets of time constants is observable and/or (ii) the non-ideal performances of the algorithm (Laplace-Padé approximant) used to estimate the time constants.

The motivation underlying the first point is that with a single, or a limited number of measurement coils, only a subset of time constants can be observed. However, this

point can be easily investigated by considering the energy of the individual terms appearing in (3) as a function of the lift-off. The energy carried by each natural mode is given by:

$$E_k = \int_0^\infty (c_k e^{-t/\tau_k})^2 dt = \frac{1}{2} \tau_k c_k^2. \quad (12)$$

Figure 5 illustrates the energy of the first 20 natural modes with different coil lift-offs. As aforementioned, when using a single pick-up coil, only a subset of time constants is observable, as appears from Figure 5. However, the subset of observable modes does not depend on the lift-offs, for the typical variations (from 0.1mm to 1mm) considered in ECT, as the peaks in the energy occur at the same positions. Therefore, the sensitivity of time constants w.r.t. lift-off is due to the extraction algorithm.

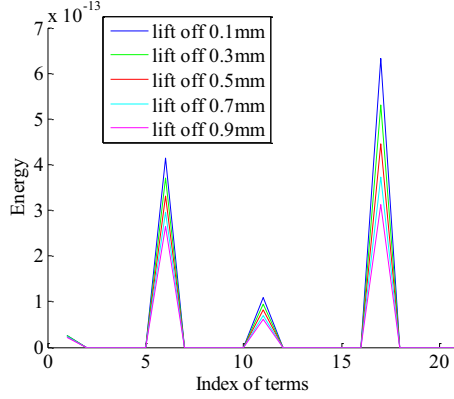


Figure 5. Energy of first 20 natural modes.

Eventually, we compared the performances in terms of sensitivity to lift-off of the proposed approach with a “traditional” eddy current methods based on time harmonic operations. Specifically, Figure 6 compares the changes in measured time constants and the changes in coil impedance with different lift-offs. Results indicate that extracted time constant is much less sensitive to lift-off if compared to the traditional approach.

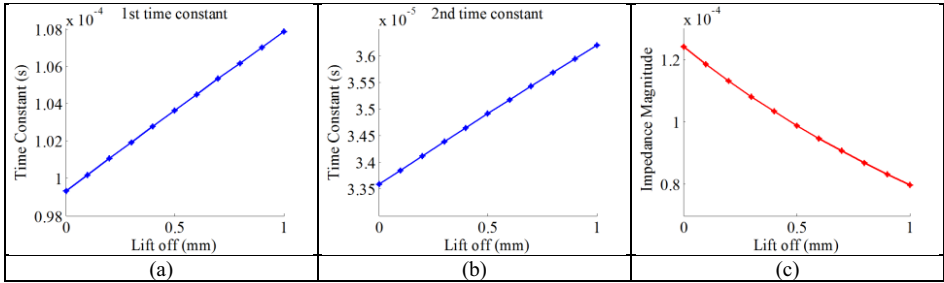


Figure 6. Changes in measured quantities with lift-off. (a) first time constant, relative difference = 8.2%. (b) second time constant, relative difference = 7.5%. (c) coil impedance, relative difference = 43%.

4. Conclusions and future work

In this paper, the monotonicity of time constants is exploited to measure the electrical conductivity of conductive materials. Moreover, we exploited a specific feature that time constants are homogeneous function of order one with the electrical conductivity of the specimen. This makes the estimation of the electrical conductivity through time constants very attractive because it requires a single reference conductive specimen to calibrate the measurement system.

Time constants depend on the electrical conductivity and geometry of the material only. They are completely independent of the type, shape, lift-off and tilting of the probe, as well as independent of the excitation waveforms. Despite of this, the intrinsic limit of the method to extract the time constants (Laplace-Padé approximant approach) makes the estimated time constants weakly depending on the lift-off of the probe. However, the method resulted to be much less sensitive to lift-off variation if compared to conventional frequency domain eddy current approaches.

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B-Spline Sparse Grids for Eddy-Current Testing Inverse Problems

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Abstract. A sparse grid surrogate model using hierarchical B-spline basis functions is used to approximate the objective function in an optimization-based inversion algorithm. The B-spline basis provides a smooth interpolant of the objective function and the gradient of the interpolant is readily available in closed-form. The latter is used in a gradient-based minimum search algorithm that results in the approximate solution of the inverse problem. The method is computationally more efficient than using gradient-free direct search methods, as illustrated by an example drawn from eddy-current nondestructive testing.

Keywords. inverse problem, sparse grid, B-spline, gradient-based optimization

1. Introduction

Surrogate models can facilitate the solution of inverse problems related to electromagnetic nondestructive evaluation. The benefit from a surrogate model is the cheap approximation of the usually heavy electromagnetic simulation. In this way, the model-based inversion—that relies on several subsequent evaluations of the forward model—can be significantly sped up.

Most surrogate modeling approaches consist in the interpolation of the input-output function based on its pre-calculated samples (training data or database). Various sampling and interpolation techniques have been applied in the context of electromagnetic nondestructive testing (NdT). The adaptive mesh-databases combined with linear [1], [2] or radial basis function interpolation [3] have been shown to apply well as surrogate models. Later, to cope with the curse-of-dimensionality (i.e., the exponential increase of storage and computational needs of the surrogate model with increasing number of input parameters), sparse grid surrogate models have been introduced. Traditionally, piecewise linear basis functions were used for the interpolation on sparse grids [4]. The efficiency of the method in the context of computational electromagnetics has been demonstrated, e.g., in [5]. In eddy-current nondestructive testing (EC-NdT), the inverse problem has also been targeted with sparse grids. A Monte Carlo sampling was built on the surrogate in [6], to jointly perform model selection and optimisation-based inversion. The idea of sparse grids in EC-NdT inversion is flashed also in other contributions, e.g., [7].

A bottleneck of piecewise linear interpolation (which is common with sparse grids) is the lack of continuous derivatives of the interpolant with respect to the input param-

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eters. This hinders the use of gradient-based algorithms of inversion through the optimization of a misfit function. This is the main motivation for using hierarchical B-splines as basis functions, e.g., in [8]. This technique yields a smooth interpolant of which the gradient is continuous and can be analytically expressed.

In the present work, the sparse grid surrogate model with hierarchical B-spline basis functions is outlined (Sec. 2) and its use for optimization-based inversion is presented (Sec. 3). Finally, a numerical example drawn from EC-NdT is discussed (Sec. 4).

2. Sparse grid interpolation with hierarchical B-spline basis functions

Let us consider a flaw model (e.g., a crack or void) with N parameters, each having lower and upper bounds. The region of interest in terms of parameters $\mathbf{x} = [x_1, \dots, x_N]$ is thus an N -dimensional space. Let us assume that the allowed domain of the parameters—say, the *input space*—is the N -dimensional unit-hypercube $[0, 1]^N$ (this is possible in most practical cases by some appropriate transformation of the parameters). The output signal (e.g., a surface scan of impedance variation of a probe coil) corresponding to the parameter vector $\mathbf{x}$ is denoted by $\mathbf{y} = \mathbf{f}(\mathbf{x})$. The vector-vector function $\mathbf{f}$ represents the NdT forward problem as an input-output function, and $\mathbf{f}$ is usually evaluated by means of numerical simulation, involving a brute-force method such as the finite element or the moment method. To reduce the computational burden associated with the evaluation of $\mathbf{f}$, one seeks for an approximation $\hat{\mathbf{f}} \approx \mathbf{f}$, i.e., the surrogate model of the original problem.

To this end, sparse grid interpolation is proposed, which is briefly summarized as:

1. definition of a hierarchical set of basis functions in 1-dimension (1-D);
2. generation of a set of hierarchical N-D basis functions as tensor product of the 1-D bases;
3. truncation of the set of N-D basis functions such that the resulting set is “sparse” in a certain sense.

Let us denote the set of 1-D basis functions at level ℓ by

$$\Psi_\ell(x) = \left\{ \psi_i^{(\ell)}(x) \mid i = 1, 2, \dots, m_\ell^{(1)} \right\} \quad (1)$$

where $\ell = 0, 1, 2, \dots, d$, with d being the depth of the hierarchical interpolation. The sparse tensor product [4] of N number of such 1-D bases at its level ℓ is given by

$$\Phi_\ell(\mathbf{x}) = \left\{ \Psi_{\ell_1}(x_1) \otimes \Psi_{\ell_2}(x_2) \otimes \dots \otimes \Psi_{\ell_N}(x_N) \mid \sum_{i=1}^N \ell_i = \ell \right\}. \quad (2)$$

A linear truncation is used herein, i.e., all N -D basis functions involved in a depth= d interpolation satisfy the linear constraint for the level indices $\sum_{i=1}^N \ell_i \leq d$. With this sparse hierarchical basis, the interpolant $\hat{\mathbf{f}}$ has the form of

$$\hat{\mathbf{f}}(\mathbf{x}) = \sum_{\ell=0}^d \sum_{i=1}^{m_\ell^{(N)}} \mathbf{c}_i^{(\ell)} \phi_i^{(\ell)}(\mathbf{x}), \quad (3)$$

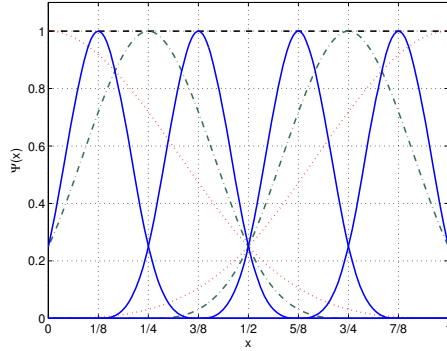


Figure 1. Hierarchical B-splines ($\psi_i^{(\ell)}(x)$) of order 3. Level 0 (---), level 1 (···), level 2 (- · -) and level 3 (—) basis functions.

where $\mathbf{c}_i^{(\ell)}$ and $\phi_i^{(\ell)}(\mathbf{x})$ are the i -th vectorial coefficient and the related basis function at level ℓ , the latter is an element of $\Phi_\ell(\mathbf{x})$ in (2), and $m_\ell^{(N)}$ is the number of basis functions at level ℓ in an N -D sparse grid, respectively.

Linear (“hat”) basis functions $\psi_i^{(\ell)}(x)$ are commonly used with sparse grids. In the present work, *hierarchical* B-spline basis functions are applied, and similarly to [8], 3rd order B-splines have been chosen. This choice ensures the continuity of the 2nd order derivatives of interpolant (3). The 3rd order *cardinal* B-spline is defined in the interval $0 \leq x \leq 4$ and it is expressed as

$$b^{(3)}(x) = \begin{cases} 1/4x^3, & 0 \leq x < 1; \\ -3/4x^3 + 3x^2 - 3x + 1, & 1 \leq x < 2; \\ 3/4x^3 - 6x^2 + 15x - 11, & 2 \leq x < 3; \\ -1/4x^3 + 3x^2 - 12x + 16, & 3 \leq x \leq 4. \end{cases} \quad (4)$$

In our implementation, the level 0 basis function is chosen as constant ($\psi_1^{(1)}(x) = 1$) and all other hierarchical basis functions ($\ell \geq 1$) are derived from a cardinal B-spline (4) via an affine transformation of the input variable, as illustrated in Fig. 1.

To determine the coefficients $\mathbf{c}_i^{(\ell)}$ in (3), the equality $\hat{\mathbf{f}}(\mathbf{x}_k) = \mathbf{f}(\mathbf{x}_k)$ is enforced at n number of control points ($\mathbf{x}_k, k = 1, 2, \dots, n$). These points are chosen as the center points of the corresponding basis functions (i.e., where $\phi_i^{(\ell)}(\mathbf{x}) = 1$ hold, for all $\ell \geq 1$, and by definition, it is $1/2[1, 1, \dots, 1]$ for $\phi_1^{(0)}(\mathbf{x})$). The number of control points is thus equal to the cardinality of the sparse basis (2), and the location of the set of these control points (nodes) forms the so-called *sparse grid*. In 2-D, the distribution of the sparse grid nodes is illustrated in Fig. 2, along with the contour lines of the sparse grid interpolant of the well-known Branin test function [9].

The sparse grid interpolation can cope with the curse-of-dimensionality in the following sense. The number of grid nodes n in a sparse grid with depth d in dimension N is in the order $\mathcal{O}\{K(\log K)^{N-1}\}$, where $K = 2^d + 1$ being the number of hierarchical basis functions per dimension. However, the increase of n is much faster when a classical full grid is used, herein $\mathcal{O}\{K^N\}$ applies [4]. The numerical example presented in Sec. 4 involves $N = 4$ dimensions, for which the node numbers of sparse and full grids

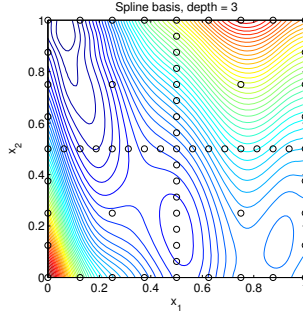


Figure 2. Contour lines of the interpolant of the Branin test function by means of a depth=3 sparse grid with hierarchical B-spline basis functions, along with the grid nodes.

Table 1. Node numbers of sparse and full grids in $N = 4$ dimension, in function of the depth d .

d	0	1	2	3	4	5	6
n (sparse)	1	9	41	137	401	1105	2929
n (full)	1	81	625	6561	83521	1185921	17850625

are compared in Table 1 as an illustration of the gain benefited from sparse grids. In spite of the large reduction of node numbers, the interpolation accuracy of sparse grids is only slightly deteriorated compared to full grids, as detailed in [4].

3. Inversion using the sparse grid surrogate model

Once the sparse grid surrogate model (3) is available, and hierarchical B-splines are used as basis functions therein, one can solve the inverse problem by means of optimizing a misfit function, as it follows. Let us assume that the output $\mathbf{y}$ of the numerical simulation $\mathbf{f}(\mathbf{x})$ (as already mentioned in Sec. 2) is a real row vector of M elements: $\mathbf{y} = [y_1, y_2, \dots, y_M]$. In the case of complex output (e.g., complex impedance variation in EC-NdT), one can always introduce the real output vector as $\mathbf{y} := [\text{Re}\{\mathbf{y}\}, \text{Im}\{\mathbf{y}\}]$. The measured data vector is denoted by $\tilde{\mathbf{y}}$. Let us define the quadratic misfit function as the squared norm of the discrepancy between simulated and measured data:

$$u(\mathbf{x}) = \|\tilde{\mathbf{y}} - \mathbf{y}\|^2 \equiv [\tilde{\mathbf{y}} - \mathbf{y}][\tilde{\mathbf{y}} - \mathbf{y}]^T \quad (5)$$

The regularized inverse problem then consists in solving the constrained optimization problem

$$\mathbf{x}' = \arg \min_{\mathbf{x} \in [0,1]^N} u(\mathbf{x}) \quad (6)$$

that yields the solution $\mathbf{x}'$. In order to reduce the computational burden, one approximates $\mathbf{y} = \mathbf{f}(\mathbf{x})$ by $\hat{\mathbf{y}} = \hat{\mathbf{f}}(\mathbf{x})$, and the approximate misfit function $\hat{u}(\mathbf{x}) = \|\tilde{\mathbf{y}} - \hat{\mathbf{y}}\|^2$ is to be minimized according to (6). Since $\hat{\mathbf{f}}(\mathbf{x})$ is based on a sparse grid interpolation with hierarchical B-spline basis functions, its gradient

$$\nabla \hat{\mathbf{f}}(\mathbf{x}) = \begin{bmatrix} \partial \hat{y}_1 / \partial x_1 & \partial \hat{y}_2 / \partial x_1 & \cdots & \partial \hat{y}_M / \partial x_1 \\ \vdots & & \ddots & \vdots \\ \partial \hat{y}_1 / \partial x_N & \partial \hat{y}_2 / \partial x_N & \cdots & \partial \hat{y}_M / \partial x_N \end{bmatrix} \quad (7)$$

can be easily expressed in closed form based on (3). Therefore, one can also write the gradient of the quadratic approximate misfit function in closed form (see, e.g., [10]) as

$$\nabla \hat{u}(\mathbf{x}) = 2[\tilde{\mathbf{y}} - \hat{\mathbf{f}}(\mathbf{x})][\nabla \hat{\mathbf{f}}(\mathbf{x})]^\top. \quad (8)$$

This gradient (8) can be used in the gradient-based solution of the optimization problem (6). In order that one can choose among the large variety of classical unconstrained optimization algorithms, the constrained problem (6) has to be re-formulated by using some nonlinear transformation of the optimization variable $\mathbf{x}$. Herein a new variable $\xi = [\xi_1, \xi_2, \dots, \xi_N]$ is introduced such that

$$x_i = (\arctan \xi_i) / \pi + 1/2, \quad x_i \in [0, 1], \quad \xi_i \in (-\infty, \infty), \quad i = 1, 2, \dots, N, \quad (9)$$

and the optimization is performed with respect to ξ in an unconstrained domain:

$$\xi' = \arg \min_{\xi \in \mathbb{R}^N} u(\mathbf{x}(\xi)). \quad (10)$$

The gradient-based local optimization strategies unfortunately suffer from the possibility of stalling in a local minimum and thus miss the global minimum. This risk is reduced when one runs the algorithm with different initial guesses. However, this was not the case in the numerical example which is presented in the next section.

4. Numerical example

4.1. Configuration

Let us consider the eddy-current testing arrangement shown in Fig. 3. A homogeneous, non-ferromagnetic plate is considered as specimen. The conductivity is $\sigma = 10^6$ S/m and the thickness is $t = 1.55$ mm. The other dimensions of the plate are assumed to be infinite. The coil is an air-cored probe (inner and outer diameters: 2 mm and 3.25 mm; height: 2 mm; no. of turns: 328; lift-off h : 0.303 mm), driven with a time-harmonic current with frequency of 300 kHz. The variation of the coil impedance is observed at regularly spaced coil positions:

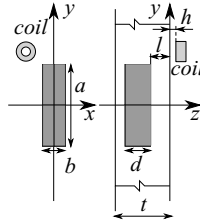
$$x_c = \{-15 : 1 : 15\} \text{ mm} \quad \text{and} \quad y_c = \{-5 : 1 : 5\} \text{ mm}, \quad (11)$$

that is, $31 \times 11 = 341$ complex impedance variations are available in the surface scan.

A rectangular-shaped defect (with zero conductivity) is present within the plate, with sides parallel to the x , y and z axes; the defect is centered on the origin of the xy plane. The bounds given in Table 2 apply for the 4 geometrical parameters of the defect. The input domain is non-rectangular but narrowed by a linear inequality constrain for d and l . This

Table 2. Parameter ranges in the single narrow crack example.

parameter		range
length	a	$[4, 22]\text{mm}$
width	b	$[0.001, 0.3]\text{mm}$
depth	d	$[10, 100]\%$ (of plate thickness)
ligament	l	$[0, t - d]$

**Figure 3.** Sketch of the configuration.

is taken into account when transforming the input domain to the $[0, 1]^N$ hypercube, on which the sparse grid and inversion algorithms are performed, as already seen in Sec. 2.

The electromagnetic simulation is based on an integral equation formulation, implemented in the CIVA software [11].

4.2. Test of the forward interpolation

The interpolation error is defined as $\varepsilon(\mathbf{x}) = \|\hat{f}(\mathbf{x}) - \mathbf{f}(\mathbf{x})\|$ at each point $\mathbf{x}$. The overall interpolation performance is characterized by the maximal and root mean-squared (rms) interpolation error, both are approximated based on a finite number of test samples $\mathbf{x}_j$ ($j = 1, 2, \dots, J$):

$$\varepsilon_{\max} = \max_j \varepsilon(\mathbf{x}_j), \quad \varepsilon_{\text{rms}} = \sqrt{\frac{1}{J} \sum_j \varepsilon(\mathbf{x}_j)^2} \quad (12)$$

In this numerical test, 4000 random test samples are used. In order to facilitate the interpretation of the error, the normalized (dimensionless) interpolation error is presented, with a normalizing factor that is chosen as $\max_j \|\mathbf{f}(\mathbf{x}_j)\|$. The change of the normalized error with respect to the depth of the sparse grid interpolation is shown in Fig. 4. For reference, linear basis functions [5] have also been used. In terms of interpolation accuracy, the linear basis slightly outperforms the B-splines. Yet, the B-splines are better in optimization-based inversion schemes, as shown in the next section.

4.3. Test of the inversion algorithm

As summarized in Table 3, the output signal associated with two randomly chosen defects (denoted as “true”) has been calculated. The inversion procedure detailed in Sec. 3 resulted in the reconstructed parameters denoted as “grad”. For comparison, a similar optimization-based inversion has been performed using the sparse grid with linear basis functions (the same as presented in the previous section). Due to the lack of continuous

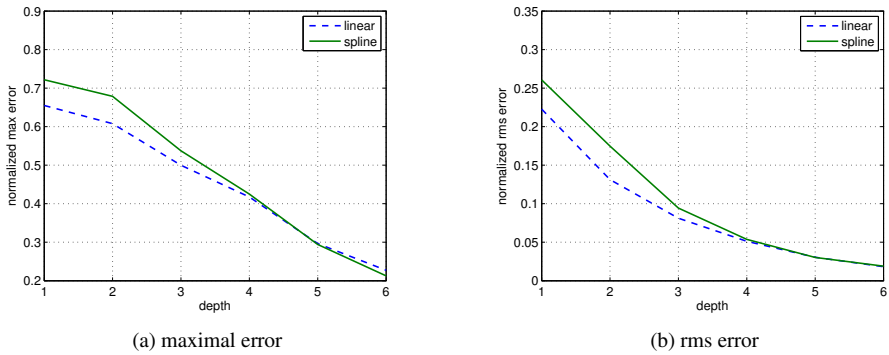


Figure 4. Normalized interpolation error in the numerical example.

Table 3. Results of the inversion in the 1st (left) and 2nd (right) test cases. The “true” parameters correspond to the defect to be reconstructed, and the gradient-based (“grad”) and simplex optimisation-based (“direct”) inversion methods yielded the parameters as given in the tables. Depth=6 sparse grid is used in all cases.

	<i>a</i>	<i>b</i>	<i>d</i>	<i>l</i>
true	16.04	0.290	0.258	0.413
grad	16.30	0.012	0.474	0.324
direct	20.42	0.118	0.348	0.389

	<i>a</i>	<i>b</i>	<i>d</i>	<i>l</i>
true	15.10	0.157	0.525	0.942
grad	14.96	0.188	0.507	0.957
direct	16.20	0.137	0.478	0.850

gradient of the interpolant, a simplex optimization (referred to as “direct”) method has been applied with the linear basis. According to Table 3, both approaches found reliable solutions, except for parameter *b*, which is the width of the defect and it is well-known to have only a weak influence on the impedance change. This measurement setup thus has limited capacity in reconstructing *b*.

The main advantage of the B-spline based scheme becomes clear when looking at the evolution of the iterative optimization routines in Fig. 5. In both examples, the gradient method converged much faster to a better local minimum than the simplex method.

Implementation has been made in Matlab, with the functions `fminunc` (a quasi-Newton method) and `fminsearch` (a simplex method for direct search).

5. Conclusion

The use of B-spline basis functions for interpolation on sparse grids has been found to be an efficient tool when performing optimization-based inversion using the sparse grid surrogate model. The gradient of the approximate misfit function is continuous, giving rise to gradient methods. Spare grids *per se* have previously proven a good performance in NdT inversion; and the present contribution is an extension of this solid framework.

Future work will include the expression the Hessian of the misfit function to use second order optimization methods. Furthermore, one will study the case when the gradient of the forward model **f** is available at hand (e.g., via the adjoint problem), and it can be used to fit the surrogate model not only to **f** but to **∇f** as well. Considerations on the adaptive generation of the sparse grid in combination with B-spline basis functions will also be taken.

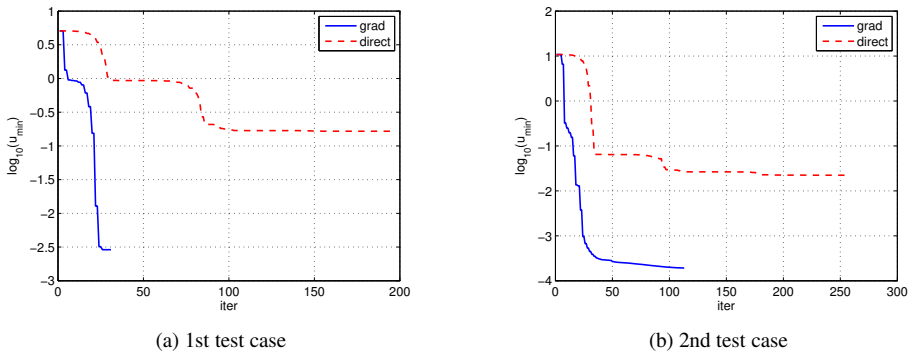


Figure 5. Performance of the inversion with gradient-based (“grad”) and simplex optimisation-based (“direct”) inversion methods: minimum misfit function found with respect to iteration number.

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Imaging of External Corrosion in Nonmagnetic Pipes via GPEC

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Abstract. Due to corrosive and hostile environment, in-service nonmagnetic pipes are prone to such anomalies as the external corrosion which poses a severe threat to structural integrity. Gradient-field Pulsed Eddy Current technique (GPEC) has been found superior to other non-destructive evaluation (NDE) techniques particularly pulsed eddy current testing in evaluation and imaging of subsurface corrosion in planar conductors. In this paper, GPEC for evaluation and imaging of external corrosion in nonmagnetic pipes is investigated. Two orthogonal gradient-field signals are simultaneously acquired and used for visualization of external corrosion. Through experimental investigation, it has been found that: (1) the depth of the external corrosion can be implied by the acquired corrosion images; and (2) the corrosion boundaries can be detected and identified via GPEC.

Keywords. Electromagnetic non-destructive evaluation, gradient-field pulsed eddy current, external corrosion, corrosion imaging, nonmagnetic pipe

1. Introduction

Pipes of such nonmagnetic materials as aluminum, copper, titanium, etc. are widely used in industrial fields involving aerospace, energy, chemical engineering, etc. The hostile and corrosive environment leaves in-service nonmagnetic pipes vulnerable to external corrosion which severely undermines the pipe integrity and is one of the critical causes leading to the structural failure of the nonmagnetic pipeline. It is highly demanded that the external corrosion can be non-intrusively detected and quantitatively evaluated during periodical inspection of the nonmagnetic pipeline. Although Electromagnetic Non-destructive Evaluation (ENDE) techniques including Eddy Current Testing [1, 2], Pulsed Eddy Current Testing (PECT) [3-5] and Ultrasonic Testing [6, 7] have been found to be advantageous in inspection of tubular conductors, their technical drawbacks have been identified especially in regard to the imaging of subsurface corrosion [8].

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By integrating PECT with gradient magnetic field measurement, Gradient-field Pulsed Eddy Current technique (GPEC) has been proposed for detection, evaluation and imaging of subsurface defects in conductors [9, 10]. Previous research has revealed that GPEC is superior to PECT in terms of higher sensitivity and accuracy in evaluation and imaging of subsurface corrosion in planar conductors [8]. This indicates that complementary to aforementioned techniques, GPEC could be a promising technique in corrosion imaging for tubular conductors. As a result, in this paper GPEC imaging of external corrosion in nonmagnetic pipes is intensively investigated through experiments. A GPEC probe consisting of a bobbin coil and gradient-field sensors measuring two orthogonal gradient fields is proposed. Through a series of experiments, the advantage of GPEC in evaluation and imaging of external corrosion in nonmagnetic pipes is further identified.

2. Proposed GPEC probe and analysis of the gradient-field characteristics

A GPEC probe comprising a bobbin coil for generation of the incident magnetic field and magnetic sensors sensing the gradient field in different directions is proposed for evaluation and imaging of external corrosion in nonmagnetic pipes. The probe structure is portrayed in Figure 1(a). With this configuration, the circumferential distribution of the induced eddy current in a flawless pipe section facing the coil lateral winding is relatively uniform. Any anomalies occurring in the pipe body perturb the eddy-current distribution, and thus break the field uniformity over the pipe circumference. This leads to the perturbation of the net magnetic field, and subsequently gives rise to the gradient field.

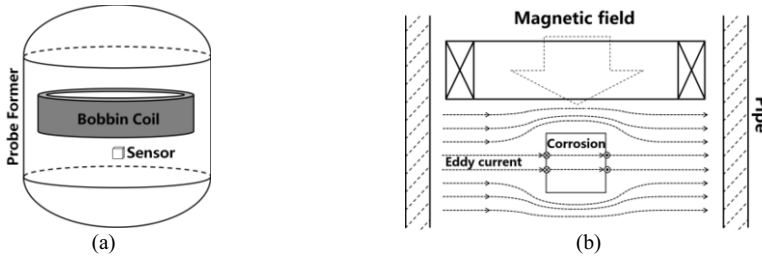


Figure 1. Schematic illustrations regarding: (a) structure of capsule-type probe; and (b) interaction of induced eddy currents in the pipe with external corrosion

The schematic illustration exhibiting the interaction of eddy currents induced in a pipe with external corrosion is shown in Figure 1(b). It can be indicated from Figure 1(b) that when the probe is placed over the pipe section free of corrosion, the circumference-direction gradient field of the circumferential component of the net magnetic field $g_\phi(B_\phi)$ is null due to the uniform eddy-current distribution over the pipe circumference. In the presence of the external corrosion, the axial component of the eddy current J_z arising from the material discontinuity introduced by the corrosion boundary particularly perpendicular to the flowing direction of the eddy current becomes non-zero. This results in the non-zero $g_\phi(B_\phi)$, the signal of which can thus be adopted to detect the corrosion boundary in the axial direction.

Similarly, significant perturbation in the circumferential component of the eddy current J_ϕ arises from the material discontinuity along the circumferential direction. Therefore, dramatic variation regarding the axial-direction gradient field of the axial component of the net magnetic field $g_z(B_z)$ which is dependent of J_ϕ can be found. The signal of $g_z(B_z)$ is thus applicable for identification of the corrosion boundary in the

circumferential direction. By simultaneously analyzing the signals of $g_{\varphi}(B_{\varphi})$ and $g_z(B_z)$, profiling and sizing of the external corrosion could be realized via corrosion visualization.

3. Experimental investigation

3.1. Experimental system

Based on the analysis of the gradient-field characteristics, a GPEC probe, which consists of a bobbin coil and two magnetic sensors individually measuring $g_{\varphi}(B_{\varphi})$ and $g_z(B_z)$, is fabricated for experimental investigation. The schematic illustration of the probe is shown in Figure 2 whilst the probe parameters are listed in Table 1. The sensors for gradient-field measurement are Tunnel Magneto-resistance (TMR) sensors (MultiDimension TMR-4002), and deployed at the same side of the coil lateral winding. The GPEC probe is connected with the testing system whose setup is exhibited in Figure 3. A rectangular-waveform excitation current drives the bobbin coil with the pulse repetition frequency, duty cycle and maximum amplitude of 100Hz, 50%, and 100mA, respectively.

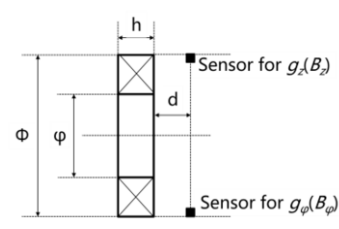


Figure 2. Schematic illustration of the GPEC probe

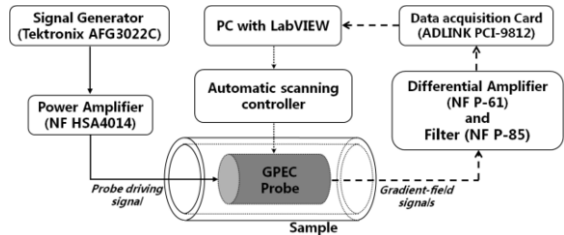


Figure 3. Schematic illustration of the GPEC system

Table 1. Parameters of the GPEC probe

Φ (mm)	φ (mm)	h (mm)	d (mm)	Number of turns (N)
48	36	10	4	1157

Four aluminum-alloy pipes which are used for fuel delivery in an aircraft are taken as the specimens adopted in the experiments. The nominal thickness, length, conductivity and relative permeability of each specimen are 6.2mm, 200mm, 34.2MS/m and 1, respectively. External corrosion in different profiles and sizes are fabricated on the external surfaces of the specimens. The cross-sectional view of each specimen with external corrosion is shown in Figure 4. It is noted that Corrosion R1 and Corrosion R2 are rectangular-shaped corrosion with the side length of 24mm whilst Corrosion C1 and Corrosion C2 are circular-shaped with the diameter fixed at 24mm.

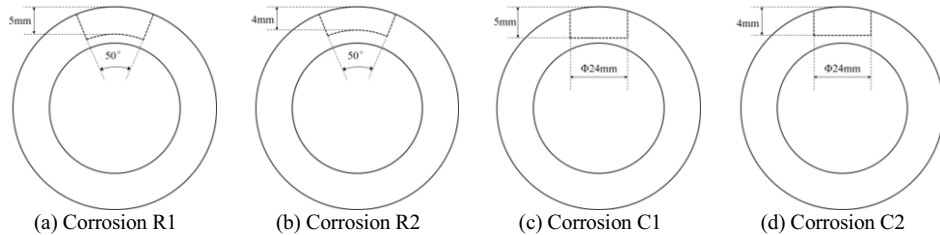


Figure 4. Schematic illustration of specimens with external corrosion (cross-sectional view)

3.2. Imaging of external corrosion

During the investigation of corrosion imaging via GPEC, the probe scans axially and circumferentially over the inner surface of each specimen with intervals of 0.5mm and 0.5° , respectively. At each scanning position, the signals of $g_\phi(B_\phi)$ and $g_z(B_z)$ are acquired from the sensors, and subsequently subtracted into the reference signals when the GPEC probe is placed in the corrosion-free area. This brings about the so-called differential signals for feature extraction and corrosion imaging. The corrosion signals of $g_z(B_z)$ and $g_\phi(B_\phi)$ when the probe is deployed right over Corrosion R1, reference signals and acquired differential signals are shown in Figure 5. It can be seen from Figure 5 that in the presence of the external corrosion the magnitude of the gradient-field signal decreases. The Peak Value (PV) which is extracted from the differential signal and null for the corrosion-free scenario could be used for detection, evaluation and imaging of external corrosion.

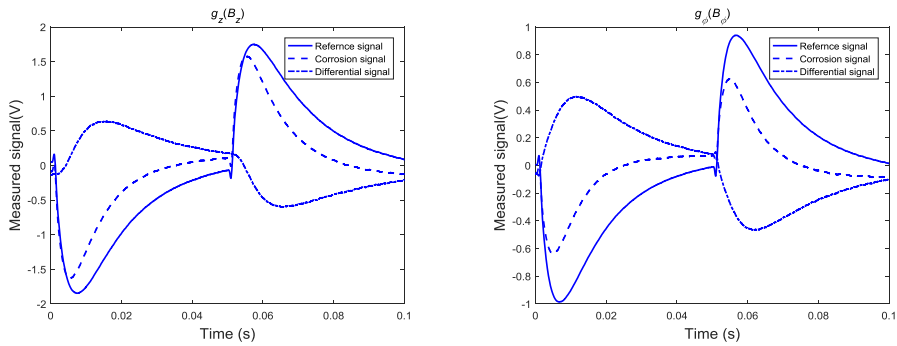


Figure 5. Measured gradient-field signals

In order to investigate the characteristics of the scanning curves for external corrosion, the axial and circumferential scanning curves depicting PVs of the differential signals against the sensor positions along the corrosion central lines are exhibited in Figure 6.

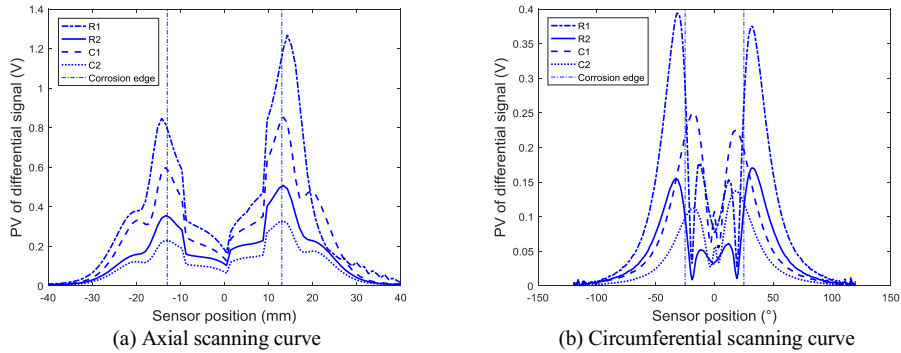


Figure 6. Acquired scanning curves

It can be observed from Figure 6 that for the corrosion in the same profile, the sensor positions corresponding to the maximum values of the scanning curve imply the edges of the external corrosion. The corrosion with deeper depth offers the increase in the scanning curve amplitude due to larger perturbation of the induced eddy currents in the specimen. The cross-comparison between Figure 6(a) and Figure 6(b) reveals that for the external corrosion with same depths, the scanning curve amplitude of the

rectangular-shaped corrosion is higher than that of the circular-shaped corrosion. This is because the averaged depth of rectangular-shaped corrosion is larger than that of the circular-shaped defect. It can also be noticeable that compared with the circumferential scanning curve which is almost symmetric, the axial scanning curve is non-symmetric. The reason lies in the fact that there is offset distance (d in Figure 2) between the excitation coil and sensor. This causes the difference in amplitudes of the signals from the sensor particularly when the sensor enters and leaves the corrosion region. In addition, compared with the corrosion edge in the axial direction, the corrosion edge in the circumferential direction can be identified with higher accuracy by using the axial scanning curve. To enhance the accuracy in identification of the corrosion edge in the axial direction by using the circumferential scanning curve, the coil height (h in Figure 2) could be decreased, which could adversely lead to decline in the scanning-curve amplitude. This opens up the follow-up investigation focused on the optimization of the probe parameters.

Based on multiple scanning curves (in the axial and circumferential directions) for each corrosion, corrosion images of $g_z(B_z)$ and $g_\phi(B_\phi)$ are obtained and exhibited in Figures 7 and 8 together with the real corrosion profiles indicated by the solid line. It is noted that the corrosion centre is set as $Z=0\text{mm}$, $\Phi=0^\circ$ whilst the scanning ranges are: $-40\text{mm} \leq Z \leq 40\text{mm}$ (for the axial scanning); and $-120^\circ \leq \Phi \leq 120^\circ$ (for the circumferential scanning). The image values are normalized to the maximum image value of Corrosion R1. It is noticeable from Figures 7 and 8 that: (1) the external corrosion in each specimen is detected by the proposed GPEC probe; and (2) edges of the detected corrosion can subsequently be indicated by using the sensor positions corresponding to the extrema in the acquired images.

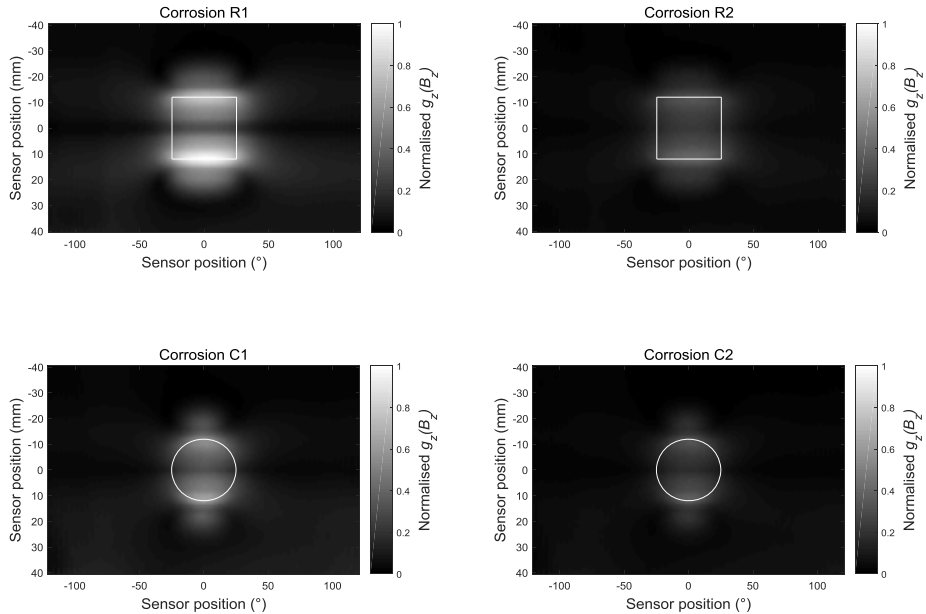


Figure 7. Corrosion images of $g_z(B_z)$

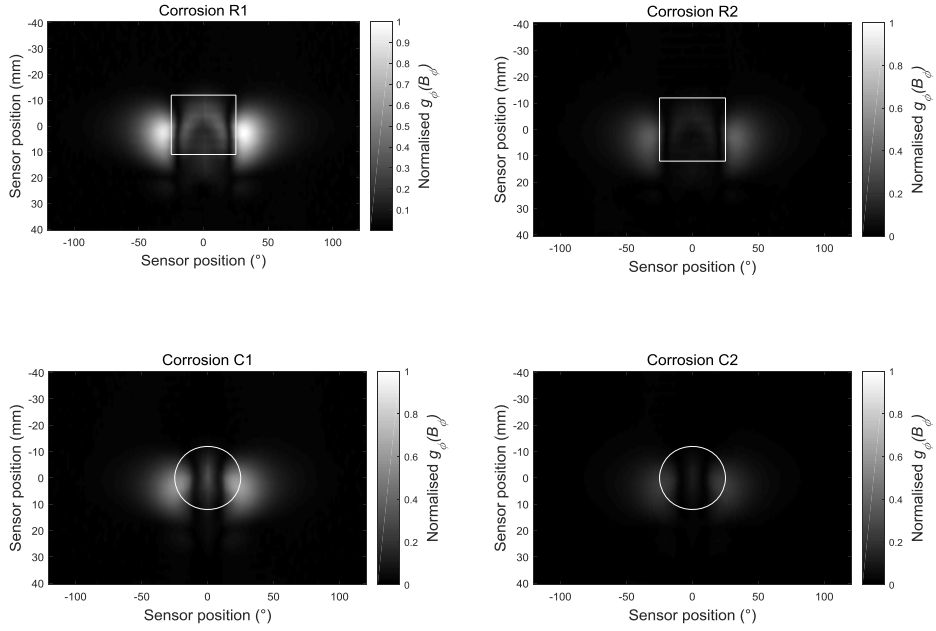


Figure 8. Corrosion images of $g_\phi(B_\phi)$

In order to further identify the corrosion profile, the contour map is derived from the corrosion image. Following this, the contour lines corresponding to the value equal to zero are extracted and used to produce the plot for corrosion profile identification. For example, regarding Corrosion R1 the obtained plots for $g_z(B_z)$ and $g_\phi(B_\phi)$ are portrayed in Figure 9. It can be found from Figure 9 that the contour lines in each plot indicate the edges of Corrosion R1 particularly the edges along the circumferential direction (as can be seen in Figure 9(a)). Compared with Figure 9(a), the corrosion edges along the axial direction, which is indicated by Figure 9(b) has discrepancy against the true corrosion profile. This could result from the dissipation and non-uniformity of electromagnetic field especially subject to axial material discontinuity in the presence of external corrosion.

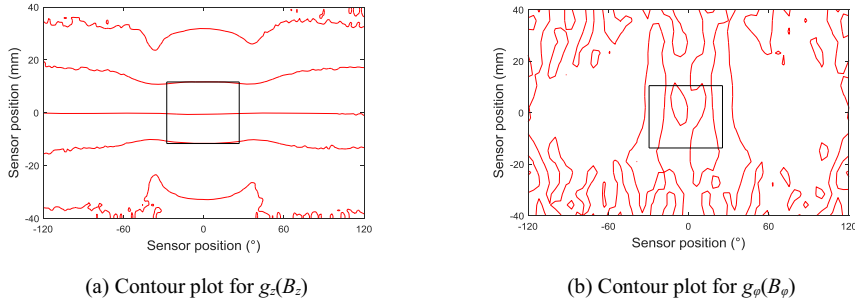


Figure 9. Contour plots with the line value equal to zero for corrosion R1

For each external corrosion scenario, the acquired corrosion images are subsequently processed by superimposing the images for $g_\phi(B_\phi)$ and $g_z(B_z)$ in an effort to generate a single image for each corrosion. The processed results are presented in Figure

10 along with the real profiles and estimated profiles acquired from contour plots. It is noted that the estimated corrosion profile is derived from superposition of the contour plots for $g_\varphi(B_\varphi)$ and $g_z(B_z)$.

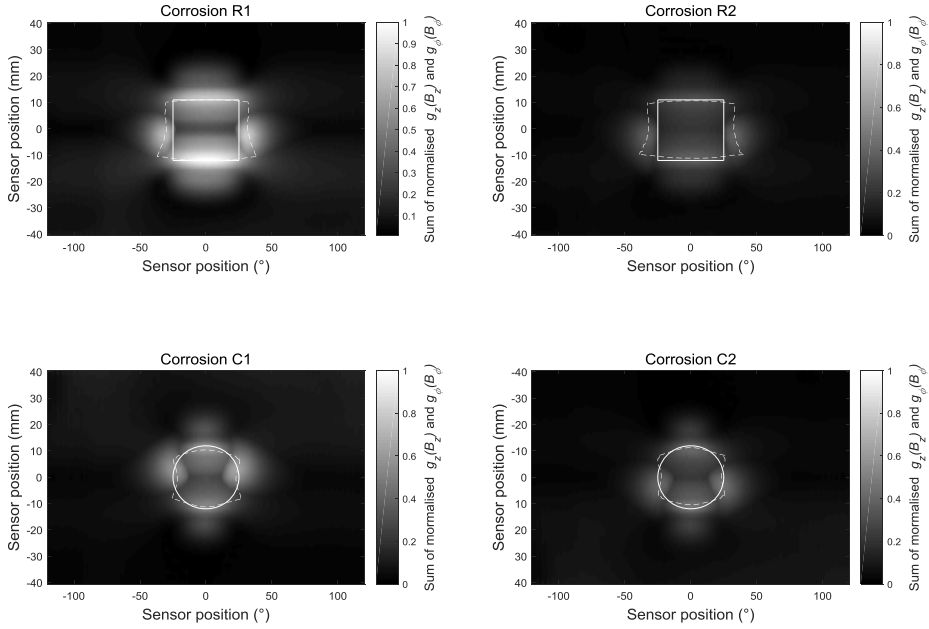


Figure 10. Acquired corrosion images along with the true and estimated corrosion profiles

It can be observed from Figure 10 that: (1) all external corrosion can be detected by using the proposed GPEC probe; and (2) the image contrast for rectangular-shaped corrosion is higher than that for circular-shaped corrosion. This agrees with the finding from analysis of images in Figures 7 and 8. For the corrosion with the same profile, the image contrast is directly proportional to the corrosion depth, which implies that the depth of external corrosion could be quantitatively evaluated by analyzing the image contrast of the acquired corrosion image. From comparison between the estimated and true profiles for each corrosion scenario, it can be seen that the imaging by using the proposed GPEC probe gives higher accuracy in identification of the material discontinuity particularly in the axial direction. This opens up the further work focused on the probe optimization and imaging processing.

4. Concluding remarks

In this paper, GPEC imaging of external corrosion in nonmagnetic pipes is intensively investigated. Based on the analysis regarding characteristics of the gradient field, a GPEC probe with sensors measuring the gradient field in two orthogonal directions is proposed. A series of experiments have been conducted to further investigate imaging of external corrosion by using the proposed GPEC probe which consists of a bobbin coil for field excitation and two sensors acquiring signals of $g_\varphi(B_\varphi)$ and $g_z(B_z)$. Multiple scanning curves obtained in circumferential and axial directions give two 2D matrices corresponding to $g_\varphi(B_\varphi)$ and $g_z(B_z)$ for corrosion imaging. Following this, the

axial/circumferential contours of the 2D matrices are applied to identify the corrosion profiles. It is noticeable from experimental results that: (1) the depth of the external corrosion could be implied by analysis of the acquired corrosion images; and (2) the corrosion boundaries can be detected and identified via GPEC. The material discontinuity particularly in the axial direction can be localized with higher accuracy. The presented work opens up the further research involving optimization of the proposed GPEC probe and characterization of external corrosion in nonmagnetic pipes via GPEC imaging.

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Void Detection of Steel Products for Process Monitoring System Using Electromagnetic Acoustic Transducer

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Abstract. This study is concerned with void detection of steel products using electromagnetic acoustic transducer (EMAT). The dynamical behaviors of the pulser-receiver EMAT measurement system are described by a transient eddy current analysis. The ultrasonic wave propagation inside steel products is then given by an initial-boundary elastic problem defined on the steel domain with unknown void parameters. Thus the parameter-to-output mapping is developed by time dependent finite element model with unknown parameters. An inversion methodology is proposed using computational intelligent approach based on the reduced order expansion. Results of computational experiments to demonstrate the efficacy of the proposed method are reported.

Keywords. Inversion, electromagnetic acoustic transducer, ultrasonic testing, signal processing, computational intelligence

1. Introduction

Quality control of continuous casting process during metal solidification is a critical issue on steel industries. Recent advances in simulation technologies are able to provide valuable information on the main causes of segregation, cracking, and fluid flow phenomena during the solidification processes. As a result, it is much important to predict their dynamic behaviors using in-line measurements. Ultrasonic testing using the electromagnetic acoustic transducer (EMAT) is a feasible method for in-line measurements in steel product monitoring system because of non-contacting device and robustness to high temperature [1][2]. Figure 1 illustrates the overall configuration of the system. At the process of the solidification process, it is extremely important to avoid the existence of internal cracking. In-line measurements by ultrasonic testing make it possible to analyze elastic properties of metal products. Our attempt in this study is to develop an inversion method for identifying profiles of metal solidification process under in-line measurements.

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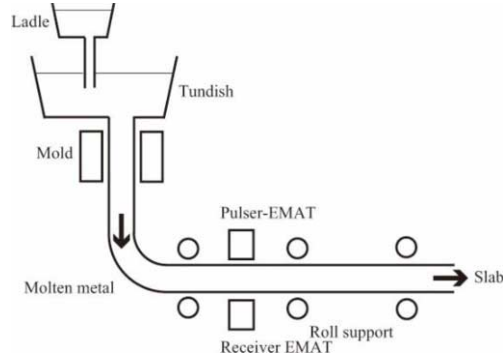


Figure 1. Overall configuration of continuous casting process

2. System configuration

EMAT consists of a magnet and a coil of wire and relies on electro-magnetic acoustic interaction for elastic wave generation. The device and the surface of molten metal interact and generate an acoustic wave. Under high temperature more than eight hundred centigrade, Lorentz force plays an essential roles in generating ultrasonic wave. Figure 2(a) illustrates the mechanism of signal generation process. The static bias magnetic field by the permanent magnet can be governed by a static magnetic potential problem. Let $A_m = A_m(x_1, x_2)$ be the x_3 component of magnetic vector potential and let M_{x_i} ($i = 1, 2$) be the magnetic moment. For economy of notations, we use $x = (x_1, x_2)$ in the sequel. Then the bias magnetic field caused by permanent magnets can be governed by

$$-\nu_m \left(\frac{\partial^2}{\partial x_1^2} + \frac{\partial^2}{\partial x_2^2} \right) A_m(x) = -\frac{\partial M_{x_2}}{\partial x_1} + \frac{\partial M_{x_1}}{\partial x_2} \quad (1)$$

where ν_m denotes the magnetic reluctivity given by

$$\nu_m = \begin{cases} 1/\mu_r\mu_0 & \text{in the magnets} \\ 1/\mu_0 & \text{Otherwise} \end{cases}$$

and where μ_0 and μ_r denote the magnetic permeability of the air and the relative permeability of the permanent magnets, respectively.

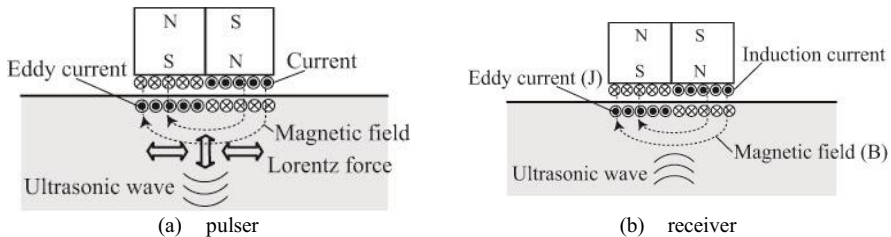


Figure 2. Illustration of pulser-receiver EMAT

The dynamic magnetic field caused by coil parts can be also governed by the same potential problem. Let $A_e^P = A_e^P(t, x)$ and $\phi^P(t, x)$ be the electromagnetic vector potential and the electrical scalar potential at time t at location x . We denote the domain of the metal product by Ω . Then the eddy currents in two spatial dimensions are governed by a time dependent potential problem:

$$\sigma\chi_\Omega \left\{ \frac{\partial A_e^P(t, x)}{\partial t} + \frac{\partial \phi^P(t, x)}{\partial x_3} \right\} - \frac{1}{\mu_0} \left(\frac{\partial^2}{\partial x_1^2} + \frac{\partial^2}{\partial x_2^2} \right) A_e^P(t, x) = J_s(t, x) \quad (2)$$

where σ , λ , and $J_s(t)$ imply the electrical conductivity of metal product, the characteristic function, and the applied current to the coil elements of EMAT device.

The excitation of ultrasonic testing signal is assumed to be governed by Lorentz force. The exciting force in two dimensions are written as ([3])

$$f^{(L)}(t, x) = \left\{ -J_e(t, x) \frac{\partial A_m}{\partial x_2}(x), -J_e(t, x) \frac{\partial A_m}{\partial x_1}(x) \right\} \quad (3)$$

where $J_e(t, x)$ denotes the eddy current given by

$$J_e(t, x) = -\sigma \left\{ \frac{\partial A_e^P}{\partial t}(t, x) + \iint_\Omega \sigma \frac{\partial A_e^P}{\partial t}(t, x) dx / \iint_\Omega \sigma dx \right\}. \quad (4)$$

The domain of the metal product is assumed to be given by the bounded domain $\Omega \in R^2$ with the boundary $\Gamma = \Gamma_P \cup \Gamma_S$. The surface confronting the EMAT device is denoted by Γ_P and the remained boundary is given by Γ_S . Let $\tilde{\sigma} = \{\tilde{\sigma}_{ij}\}$, $\varepsilon^D = \{\varepsilon_{ij}^D\}$, and $u = \{u_{ij}\}$ be the stress, the deviatoric strain, and displacement at time $t \in [0, t_f]$ ($t_f > t_w$) and at location $x \in \Omega$ for $i, j = 1, 2$. Then the wave propagation is governed by

$$\begin{aligned} \frac{\partial^2 u(t, x)}{\partial t^2} + \nabla \cdot \tilde{\sigma}(t, x) &= 0 \\ \tilde{\sigma}_{ij}(t, x) &= \lambda \{ \nabla \cdot u(t, x) \} \delta_{ij} + 2G \varepsilon_{ij}^D(u) \\ \varepsilon_{ij}^D(t, x) &= \frac{1}{2} \left\{ \frac{\partial u_i(t, x)}{\partial x_j} + \frac{\partial u_j(t, x)}{\partial x_i} \right\} - \frac{1}{2} \{ \nabla \cdot u(t, x) \} \delta_{ij} \\ &\text{in } (0, t_f) \times \Omega, (i, j = 1, 2) \end{aligned} \quad (5a)$$

with the initial and boundary conditions

$$u(0, x) = \frac{\partial u(0, x)}{\partial t} = 0 \quad \text{on } \Omega \quad (5b)$$

$$\tilde{\sigma}^t \cdot n = f^{(L)}(A_e^P) \quad \text{on } (0, t_f) \times \Gamma_P, \quad \tilde{\sigma}^t \cdot n = 0 \quad \text{on } (0, t_f) \times \Gamma_S \quad (5c)$$

where λ and G denote the Lamé's first parameter and the modulus of transverse elasticity, respectively. Furthermore, the additional absorbing boundary conditions are assumed on Γ_S .

The dynamics of the receiver EMAT is governed by the same potential model as in the pulser-EMAT. Let $A_e^R = A_e^R(t, x)$ and $\phi^R = \phi^R(t, x)$ be the electromagnetic vector

potential and the electrical scalar potential at time t at location x . Then the magnetic potential A_e^R of the reflecting signal is described by

$$\sigma\chi_\Omega \left\{ \frac{\partial A_e^R(t, x)}{\partial t} + \frac{\partial \phi^R(t, x)}{\partial x_3} \right\} - \frac{1}{\mu_0} \left(\frac{\partial^2}{\partial x_1^2} + \frac{\partial^2}{\partial x_2^2} \right) A_e^R(t, x) = J_0(t, u) \quad (6)$$

where

$$J_0(t, u) = -\sigma\chi_\Omega \left\{ \frac{\partial u_{x_1}}{\partial t} \frac{\partial A_m}{\partial x_1}(x) + \frac{\partial u_{x_2}}{\partial t} \frac{\partial A_m}{\partial x_2}(x) \right\}. \quad (7)$$

The driven forces of the receiver EMAT (6) imply the interaction between the particle velocities of ultrasonic waves and the bias magnetic fields, as simply depicted in Fig. 2(b). Consequently, the detecting voltage is given by

$$V_d(t) \propto \sum_n N \frac{d}{dt} \oint A_e dl_n \quad (8)$$

where N , dl_n , and A_e denote the number of coil turns, line element of the n -th coil, and the magnetic potential at the coil, respectively.

3. Forward problem

The forward problem considered here is to formulate the parameter-to-output mapping. As illustrated in Fig. 3(a), the void domain is defined by an ellipsoidal shape,

$$\Omega(q) = \left\{ x = (x_1, x_2) \mid \frac{(x_1 - x_1^c)^2}{a^2} + \frac{(x_2 - x_2^c)^2}{b^2} < 1 \right\} \quad (9)$$

where the parameter vector $q = \{a, b, x_1^c, x_2^c\}$ is characterized by the void shape to be identified. To discretize the system model by a finite element method, the metal product Ω is divided into a finite number of elements $\Omega_h = \bigcup_h \Delta_h$. The associated finite element model to Eqs. (2), (5), and (6) can be represented by

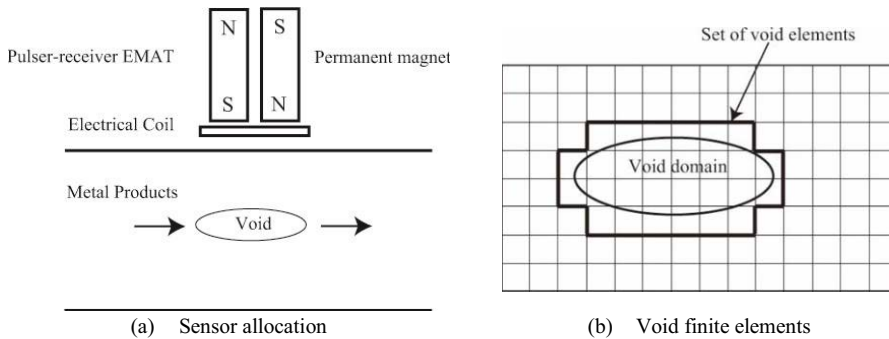


Figure 3. Void shape and unknown parameters

$$[M_A^{P,h}] \left\{ \frac{dA_e^{P,h}}{dt} \right\} + [K_A^{P,h}] \{A_e^{P,h}\} = \{J_s^h\} \quad (10a)$$

$$[M_E^h(\Delta_h)] \left\{ \frac{d^2 u^h}{dt^2} \right\} + [C_E^h(\Delta_h)] \left\{ \frac{du^h}{dt} \right\} + [K_E^h(\Delta_h)] \{u^h\} = \{f^{(L)}(A_e^{P,h})\} \quad (10b)$$

$$[M_A^{R,h}] \left\{ \frac{dA_e^{R,h}}{dt} \right\} + [K_A^{R,h}] \{A_e^{R,h}\} = \{J_o^h(u^h)\}. \quad (10c)$$

Define the void finite elements $\Delta_h^c \in \Omega(q)$. Let $\Omega_h(q) = \bigcup_h \Delta_h^c$ be the minimum coverage of the set $\Omega(q)$ as shown in Fig. 3. Then the forward analysis $\Omega_h(q) \xrightarrow{\text{yields}} Y(\Omega_h(q)) := \{y(n; \Omega_h(q))\}_{n=1}^{N_d}$ can be performed by solving the finite element models corresponding to Eq. (10) with the set of the finite elements $\Omega_h - \Omega_h(q)$. Figure 4 demonstrates simulation results corresponding to three different voids.

4. Inverse problem

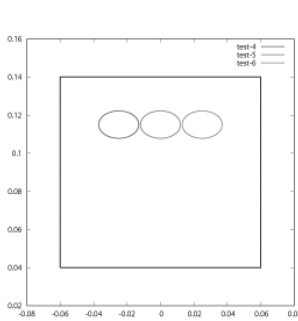
Our inversion is to estimate the position and the dimension of void $\Omega(q)$ inside metal product. To this end, a steady-state genetic algorithm (SSGA) was adopted to the inversion [4]. The SSGA simulates the continuous model of the generation, and eliminates and generates a few individuals at each generation. Let $\{q_i\}_{i=1}^M$ be a set of individuals given by

$$q_i = \{q_i^j\}_{j=1}^4 = \{a_i, b_i, x_1^c(i), x_2^c(i)\} \quad \text{for } i = 1, 2, \dots, M. \quad (11)$$

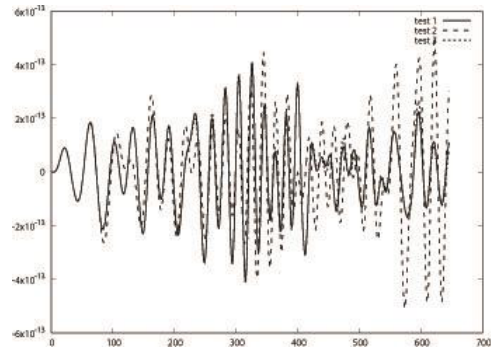
Since the shape parameters of void are given by real numbers, the following adaptive mutation was applied to the inverse problem [5]:

$$q_i^j(k+1) = q_i^j(k) + \gamma(k) \left| \frac{fit_i(k) - \min_i fit_i(k)}{\max_i fit_i(k) - \min_i fit_i(k)} \right| \theta(k) \quad (12)$$

for $i = 1, 2, \dots, M, j = 1, 2, 3, 4$ in probability $\eta(> 0)$



(a) Allocation of voids



(b) Detected voltage

Figure 4. Example of the forward problem

where $\max_i fit_i(k)$ and $\min_i fit_i(k)$ are the maximum and minimum fitness values, respectively and where $\theta(k)$ denotes the sequence of mutually independent random variables $\theta(k) \sim N(0, 1)$. The fitness value for the i -th individual is evaluated by

$$fit_i\{\Omega(q_i)\} = \sum_{n=1}^{N_d} \frac{1}{\{\eta_d(n)\}^2} |y(n, \Delta_{q_i}) - y_d(n)|^2 \quad (13)$$

where $y(n, \Delta_{q_i})$ and $\eta_d(n)$ denote the observed data and their standard deviation at time instant n , respectively. In the numerical experiments, the syntheses data were constructed as follows. First, an arbitrary true void shape was preassigned by $\Omega(q_{true})$. Secondly, the corresponding finite element models were solved numerically for the prescribed domain $\Omega_h - \Omega_h(q_{true})$. Thus the observed data $Y_d = \{y_d(n)\}_{n=1}^{N_d}$ can be prepared by

$$y_d(n) = \{1 + \xi(n)\}y\{n; \Omega_h(q_{true})\} \quad (14)$$

where $\xi(n)$ denote the independent sequence of Gaussian random variables with $N(0, \zeta)$. Consequently, the standard deviation in Eq. (13) becomes

$$\eta(n) = \xi(n)|y\{n; \Omega_h(q_{true})\}| \quad for \ n = 1, 2, \dots, N_d. \quad (15)$$

Table 1 reports the true and estimated parameters for the syntheses data with 0.32% noise. Figure 5 represents the true and recovered void shape for the test examples listed in Table 1.

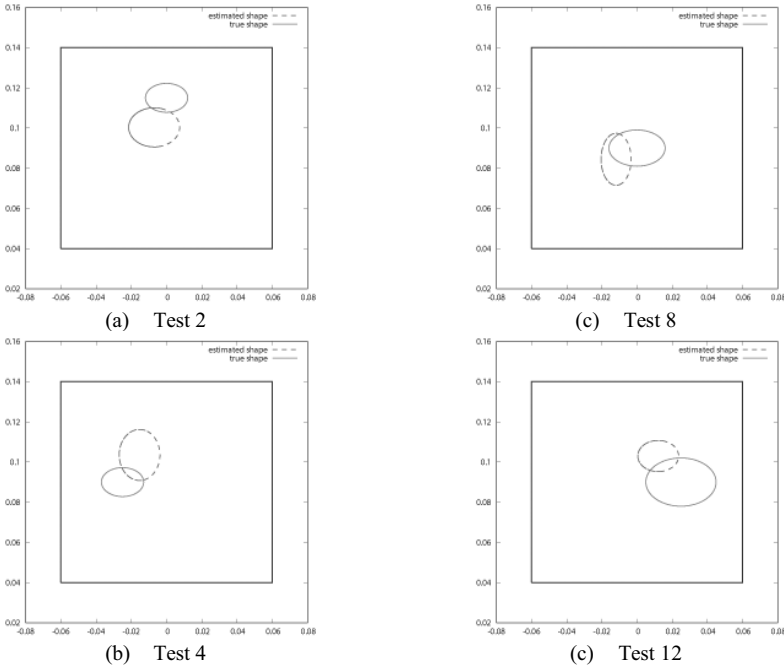


Figure 5. Estimated void shape

Table 1. Test summaries

No.		a	b	x_1^c	x_2^c
1	true	1.20e-2	7.20e-3	-2.50e-2	1.15e-1
	estimate	8.37e-3	9.61e-3	-2.11e-2	9.71e-2
2	true	1.20e-2	7.20e-3	0.0	1.15e-1
	estimate	1.45e-2	9.73e-3	-7.03e-3	1.00e-1
3	true	1.20e-2	7.20e-3	2.50e-2	1.15e-1
	estimate	9.17e-3	1.12e-2	-1.36e-3	9.60e-2
4	true	1.20e-2	7.20e-3	-2.50	9.009e-2
	estimate	1.16e-2	1.27e-2	-1.53e-2	1.04e-1
5	true	1.60e-2	9.00e-3	-2.50e-2	9.00e-2
	estimate	1.45e-2	8.42e-3	1.34e-2	7.86e-2
6	true	2.00e-2	1.20e-2	-2.50e-2	9.00e-2
	estimate	2.31e-2	1.72e-2	-6.19e-4	8.46e-2
7	true	1.20e-2	7.20e-3	0.0	9.00e-2
	estimate	9.19e-3	1.21e-2	2.01e-2	9.85e-2
8	true	1.60e-2	9.00e-3	0.0	9.00e-2
	estimate	8.58e-3	1.29e-1	-1.19e-2	8.44e-2
9	true	1.60e-2	9.00e-3	0.0	9.00e-2
	estimate	9.42e-3	1.14e-2	1.12e-2	9.92e-2
10	true	1.20e-2	7.20e-3	2.50e-2	9.00e-2
	estimate	1.79e-2	1.62e-2	-5.32e-3	9.05e-2
11	true	1.60e-2	9.00e-3	2.50e-2	9.00e-2
	estimate	1.25e-2	8.16e-3	1.02e-2	7.69e-2
12	true	2.00e-2	1.20e-2	2.50e-2	9.00e-2
	estimate	1.16e-2	7.75e-3	1.20e-2	1.03e-1
13	true	1.20e-2	7.20e-3	-2.50e-2	6.50e-2
	estimate	9.53e-3	1.03e-2	1.62e-2	8.62e-2
14	true	1.20e-2	7.20e-3	0.0	6.50e-2
	estimate	7.73e-3	8.37e-2	-1.51e-2	7.35e-2
15	true	1.20e-2	7.20e-3	2.50e-2	6.50e-2
	estimate	9.18e-3	1.73e-2	3.31e-3	8.44e-2

5. Concluding remarks

An inversion methodology is proposed using computational intelligent approach based on evolutionary strategies. To this end, the series of forward problems were formulated by electromagnetic analyses and an ultrasonic wave propagation model and the corresponding numerical schemes were developed by finite element methods. It was argued that the algorithm was worked in the test experiments. Further refinement might be necessary in order to apply the laboratory experiments.

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Product Uniformity Control – A Research Collaboration of European Steel Industries to Non-Destructive Evaluation of Microstructure and Mechanical Properties

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Abstract. In steel manufacturing, the conventional method to determine the mechanical properties and microstructure is by offline, destructive (lab-)characterisation of sample material that is typically taken from the head or the tail of the coil. Since coils can be up to 7 km long, the samples are not always representative for the main coil body. Also, the time delay (typically a few days) between the actual production and the availability of the characterisation results implies that these results cannot be exploited for real-time adaptation of the process settings.

Information about the microstructure and material properties can also be obtained from electromagnetic (EM) and ultrasonic (US) parameters, which can be measured in real-time, non-destructively, and over the full length of the steel strip

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product. With the aim to improve the consistency in product quality by use of inline EM and US measurements, a European project called "Product Uniformity Control" (PUC) has been set up as a broad collaboration between 4 major European Steel Manufacturers and 10 Universities / Research institutes.

Using both numerical simulations and experimental characterisations, we study the inline measured EM and US parameters in regard of the microstructural and mechanical properties. In this way, we aim to establish an improved understanding of their mutual relationships, and to apply this knowledge in existing and new non-destructive evaluation techniques.

In this paper, the concerted approach of modelling and experimental validation will be addressed, and results of this work will be shown in combination with inline measured data.

Keywords. steel, material characterisation, mechanical properties, non-destructive evaluation, inline monitoring

1. Introduction

The majority of steel used for automotive applications is produced as 'infinitely' long flat (strip) steel, with typical width of 1 – 2 m, and thickness ranging from 0.6 – 1.6 mm. For the ease of transport to the customer, the steel industry ships the strip steel as coils to the customers, see figure 1. Each coil typically measures several km of strip length and weights 10 – 20 tons. Next to the timely delivery of the product according to the required dimensional tolerances, the steel manufacturer strives for high quality standards, in particular regarding the surface condition, as well as the forming properties and the mechanical strength. This paper focuses on the latter quality aspects, i.e. the mechanical properties.



Figure 1. Steel coils ready for dispatch to customers.

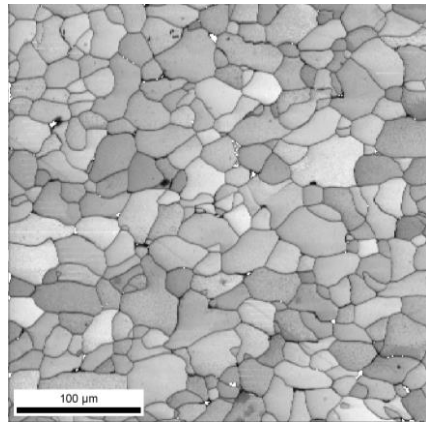


Figure 2. Example of granular microstructure of steel, here of a traditional low-carbon steel.

The mechanical properties are governed by the microstructure of the material. Traditionally, the microstructure and the mechanical properties of steel strip are determined from samples taken from the head or tail parts of a coil. From these samples, specimens are prepared for (automated) tensile testing and for microstructure analysis. Typical engineering parameters deduced from the stress-strain curve obtained by tensile testing are the yield strength (R_p) and tensile strength (R_m). Microstructural parameters that have a relation to the mechanical properties are grain size (distribution), crystal

orientation (texture) (distribution) and phase (or phase constituent) volume fractions. These can be imaged by an electron backscatter diffraction (EBSD) measurement, as illustrated in figure 2, which shows the grain pattern of a low carbon steel.

In recent years, the strong drive to reduce the weight of cars has stimulated the development and manufacturing of new high-strength steel grades. These grades owe their higher strength from the finer dispersion of their microstructure. In particular, the microstructure of these ‘advanced high-strength steels’ (AHSS) is featured by reduction of the grain size and the introduction of (hard) secondary phases, small precipitates, and highly irregular grain boundaries, and combinations thereof, as illustrated in figure 3.

In comparison to the regular grades, the microstructure and hence the mechanical properties of these AHSS grades are more sensitive to the exact thermo-mechanical processing, i.e. the heating and cooling trajectories.

The partly batch-wise processing of the steel, in the form of slabs / coils, intrinsically results in the impossibility to have a fully identical thermal treatment for each part of the strip. For instance, the heating of slabs in furnaces is non-uniform, speed variations occur in the process (acceleration and deceleration of the strip), and the cooling down of the hot-rolled coil is non-uniform. As shown in figure 4, these phenomena may lead to a significant variation of the microstructure and hence in the mechanical properties. This typically occurs in the extremities (head and/or tail) of a strip. Figure 4 shows the variation of a magnetic measurement (here: IMPOC) over the length of an AHSS strip (Dual Phase grade (DP)). Clearly, the first 50 m of the strip deviates from the bulk part.

It is notably this type of inline measurement technology that can help to get insight into the occurrences of microstructure variations, i.e. to detect their occurrence, evaluate their severity and study the frequency of occurrence in relation to (given combinations of) processing parameters. Further along this road, microstructure monitoring can provide checks to keep up with the required tighter process window for AHSS and to ensure that the product meets the mechanical property specification over its entire length.

To investigate the topic of the uniformity of steel strip in more detail, and to develop instrumentation for the fast, inline characterisation thereof, a large European consortium, comprising 4 major European steel manufacturers, has been established in the project ‘Product Uniformity Control’, started in 2013.

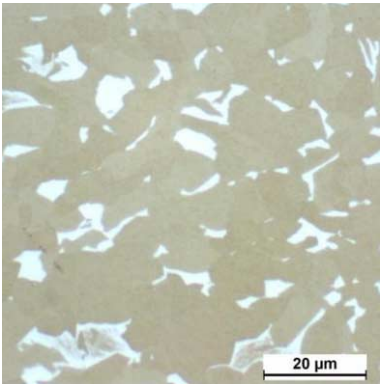


Figure 3. Example of AHSS microstructure, here with about 15% martensite (white area in picture), 12% bainite, and 73% ferrite.

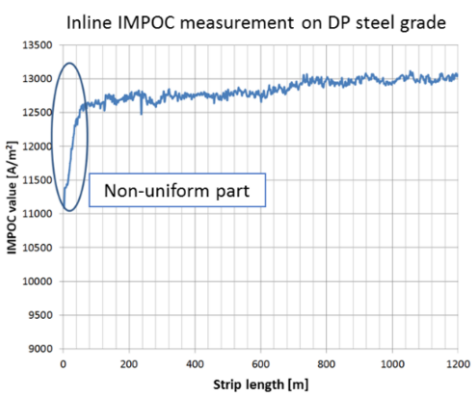


Figure 4. Example of inline magnetic measurement (IMPOC) on cold-rolled, galvanized AHSS steel (Dual Phase steel).

2. Background

2.1. *Microstructure of steel*

The microstructure of steel is formed during the manufacturing process of subsequently casting, hot-rolling, cold-rolling, (continuous or batch) annealing and finishing (temper rolling, coating and skin-pass rolling). Low-alloyed, low carbon steel has a ferritic grain like structure with average grain size in the 10 – 50 micron range. Higher alloying contents, often in combination with faster cooling trajectories, result in the creation of (additional, i.e. ‘secondary’) phases or phase constituents like pearlite, bainite and martensite, which are generally harder than ferrite.

In the study of the microstructure, different microstructure parameters are distinguished², i.e.:

- Grain size (distribution)
- Grain morphology, grain elongation (aspect ratio)
- Texture (crystal orientation)
- Secondary phase fractions (and connectivity of phases)
- Precipitate density and precipitate size
- Dislocations

Precipitates are small particles (1 nm – 1 μ m) of carbides and nitrides of (typically) Nb, Ti or V and dislocations are (tiny) lattice distortions due to straining of the material. All the microstructure parameters listed above play a role in the determination of the mechanical properties of steel, and hence they are of importance for the aim of materials characterisation. However, during the manufacturing of steel, these microstructural parameters will not vary one by one, but their variation of almost always coupled. This makes it difficult to attribute a change in a (macroscopic) property to its individual microstructural contributions in a quantitative way.

2.2. *Non-destructive evaluation techniques used in steel manufacturing*

Past research work [1-17] has investigated and developed electromagnetic (EM) and ultrasonic (US) techniques that are sensitive to individual features of the steel microstructure. Advantages of the EM and US techniques are their non-destructive nature and the potential to apply them in non-contact mode, and in harsh environments. These capabilities give them preference over X-ray diffraction (XRD) based methods: although providing more direct information on the microstructure in a lab environment, XRD measurements are (still) regarded to be too vulnerable and too expensive for use in inline monitoring applications in the steel processing industry, where water, dust and the fast movements of the product would spoil the signals. In the present scope of electromagnetic non-destructive evaluation (ENDE), this contribution will address only the EM related work that is carried out in the PUC project.

In the application of EM techniques for the NDE of the microstructure, an imminent problem is the fact that, generally spoken, magnetic properties are sensitive to all the microstructure parameters of steel listed in section 2.1, and hence are not selective. Accordingly, this poses a huge challenge in the interpretation of a variation in a measured

² The chemical composition of the steel is supposed to be known from the chemical analysis in the steelmaking plant.

EM signal in terms of a variation in microstructure. The next section highlights the approach which is followed by the consortium to investigate this problem in detail.

3. Approach

From the previous section, it is clear that the inverse problem in EM has too many solutions to give direct answers without further knowledge of the possible microstructure configurations that could occur within the scope of the process. Therefore, an approach has been defined consisting of both an experimental characterisation chain and a modelling chain, which has been depicted schematically in figure 5. In the experimental characterisation chain, samples from different stages in the steel production chain have been taken to investigate their (uniformity in) microstructure, EM properties and NDE instrument output parameters. In addition, a selection of samples are used for dedicated laboratory experiments to investigate influences from operating conditions, like the influence of lift-off, (transversal) speed and zinc coating thickness (since most steel suppliers galvanise the steel strip for automotive applications). Moreover, small reference samples have been prepared specifically to explore systematically the effect of variations in the microstructure parameter space on the fundamental (ultrasonic and) magnetic properties. The experimental data from these experimental studies partly serves as input, and partly as validation data for the simulation chain. In the simulation chain, models are built to investigate the direct problem: i.e. given a certain well-defined microstructure, what are its EM properties? And given these EM properties, what are the responses of the EM measurement equipment (in given operating conditions)?

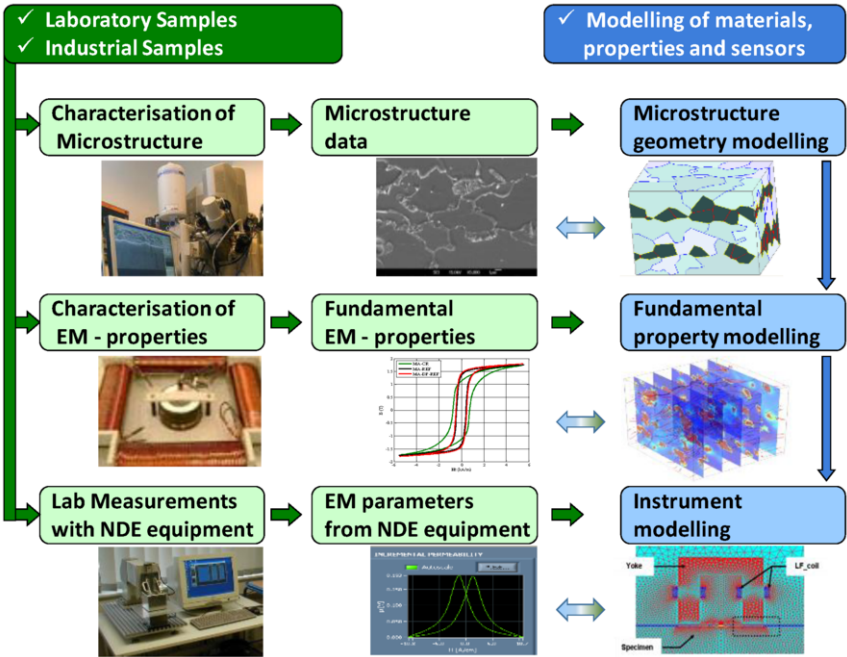


Figure 5. Experimental and modelling chains in the PUC project.

With the use of computer generated microstructures, it is straightforward to vary just one microstructure parameter at a time to study its individual effect on the magnetic properties (and on the NDE measurement), and rank the effects according to importance in a given position in the microstructure parameter space. Another possibility is to generate a multitude of different microstructures, even beyond physically realistic structures, to explore the full microstructure parameter space.

4. Description of models

Modelling of fundamental magnetic properties of polycrystalline microstructures like steel has to be considered as a 3D problem. Model reduction to 2D model descriptions is deemed to fail because even in the simpler case where isotropic behaviour is assumed, the reduction of the number of degrees of freedom limits the possible magnetic flux paths, resulting in solutions that falsely predict a harder magnetic behaviour. As a consequence, 2D microstructure descriptions that are readily obtained from micrographs or EBSD measurements cannot be used as direct input to magnetic property modelling, which underlines the need to rely on 3D microstructure geometry models for this purpose.

4.1. Microstructure geometry modelling

The 3D microstructure geometries are generated using multi-level Voronoi modelling [18-20], see figure 6 for illustrations. The microstructure generation tool allows to control grain size, grain morphology, texture, phase ratios and their distributions. The resulting geometries are described in text files with a format similar to EBSD output, providing a simple and efficient way to interface to fundamental property models.

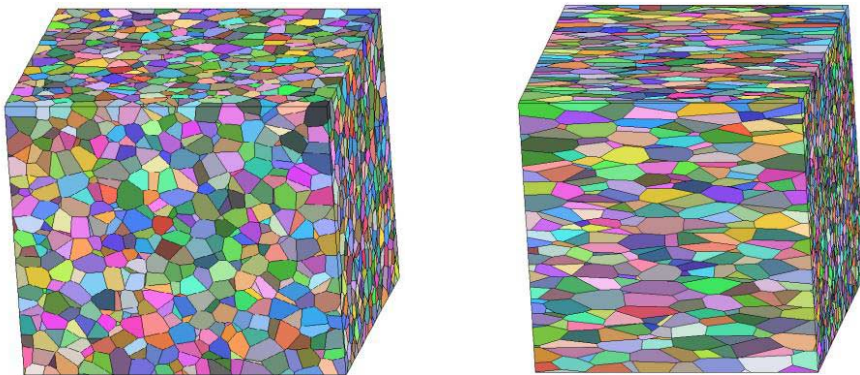


Figure 6. Example of computer generated 3D microstructures. Left: with equal grain aspect ratio in all 3 dimensions; Right: with 1:1:3 aspect ratio in grain size, allowing to study the effect of grain morphology.

4.2. Fundamental EM property modelling

For the modelling of the fundamental magnetic behaviour, two courses are taken: a micromagnetic approach to solve the magnetic problem on the (sub-)micron scale, and a finite element approach to deduce effective magnetic permeability on a mesoscale.

4.2.1. Micromagnetic modelling

Micromagnetism, introduced by William Fuller Brown Jr in 1963 [21], treats the physics of magnetism at sub-micrometer scale. The small scale is required to resolve domain walls and their interaction with matter. Due to the fine spatial resolution that is required, the numerical computation of micromagnetic problems demands high computational power and the solutions are often limited to small volumes.

The solving technique applied here uses a finite difference like discretization scheme of the Landau and Lifshitz equations to find the energy minimum of the sum of these contributions at each step of a magnetisation cycle, resulting in the magnetic hysteresis loop [22,23]. It appears that a stable solution can be achieved when the configuration is set to saturation magnetization in the initial state, and relaxed to zero applied field.

4.2.2. Modelling of the effective relative magnetic permeability

For the particular purpose of phase ratio evaluation in dual phase and multiphase steels, a fast and effective solution to model the effective magnetic permeability is found by 3D finite element modelling in COMSOL of the granular structure, in which the grains have been assigned to different phases with different magnetic permeabilities. The same approach has been used to account for the grain size effect on the magnetic permeability by assigning a different (higher) permeability to a narrow zone around the grain boundary compared to the bulk.

4.3. EM-based NDE instrument modelling

EM NDE instruments that are used for inline measurement in steel manufacturing and that are studied within the PUC project, are IMPOC, HACOM, EMSPEC (MFIA) and 3MA. These instruments have been described in [24,25].

The modelling of these instruments is performed by different institutes using different software packages. For IMPOC [26], a combination of Finite Integration Techniques and a mesh-less modal approach is used to calculate the field gradient after a pulse excitation. HACOM [27] and EMSPEC [28] have been modelled in ANSYS-MAXWELL [29] and the 3MA [30] in ALTAIR-FLUX [31].

5. Results

Low carbon steel samples of identical chemistry have been heat treated to produce a range in grain size. These samples have been characterised on microstructure (grain size) and on magnetic properties, see figure 7. Their microstructures have been simulated in 3D geometry models, serving as input for the fundamental modelling of the magnetic behaviour. Figure 7 compares the initial magnetic permeability ($\mu_{r,i}$) and the coercive field H_c as function of the grain size. Details on samples, measurements and models are given in [32]. A good agreement is found between model prediction and experimental results, in particular for the $\mu_{r,i}$ -model while the micromagnetic model may suffer from a too coarse mesh for a proper description of H_c in the lower grain size region.

A demonstration of the capabilities of the instrument modelling is given in figure 8, showing the modelled and measured incremental permeability (IP) as sensed by the 3MA instrument for different lift off settings. As can be observed, the IP curve shape is well

reproduced. The effect of lift off on the signal drop is qualitatively well reproduced, albeit from a quantitative perspective further improvements could be made.

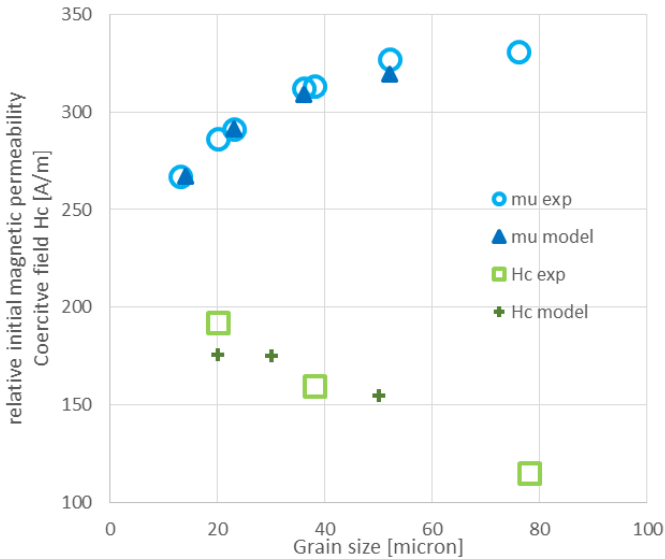


Figure 7. Model and experimental results for the initial magnetic permeability (μ) and coercive field H_c as a function of the grain size, for a fully recrystallised low carbon steel.

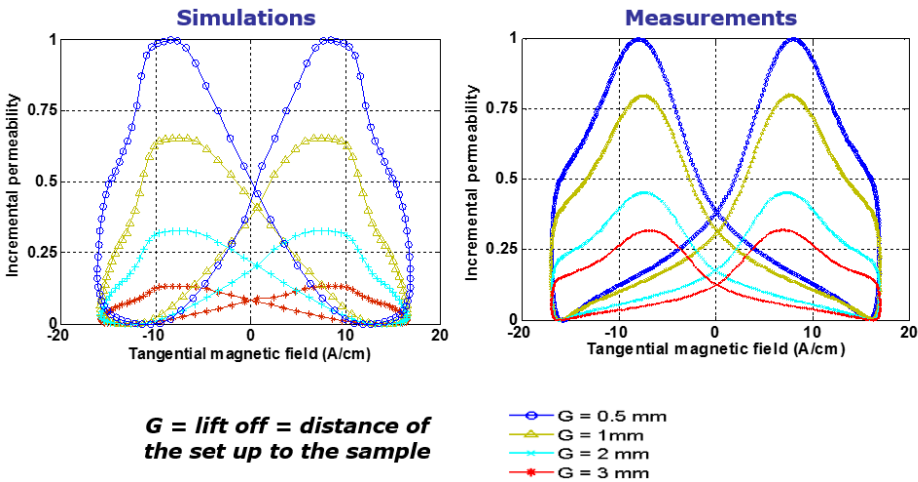


Figure 8. Simulated and measured incremental permeability of the 3MA, for different lift off settings.

An illustration how the inline magnetic data can be converted to microstructure information is given in Figure 9. Here, the data from the inline IMPOC system, situated in the pickling line of one of the steel manufacturers in the consortium, has been analysed on its deviation from the baseline of a given coil of an interstitial-free steel grade. The

deviations stem from the positioning of the steel slabs on support bars in the reheating pusher-furnace prior to hot rolling. At the support bar positions, the slab temperature remains lower than in the unsupported length sections, resulting in slight differences in grain size for this grade.

From measurements on samples, phenomenological relations have been deduced between IMPOC values and coercive field H_c , and between H_c and grain size (see Figure 7), which form the basis for the calculation of the grain size evolution over strip length as depicted in Figure 9. In the calculation, a baseline grain size of 28 microns has been assumed based on microstructure characterisations on multiple samples from this grade.

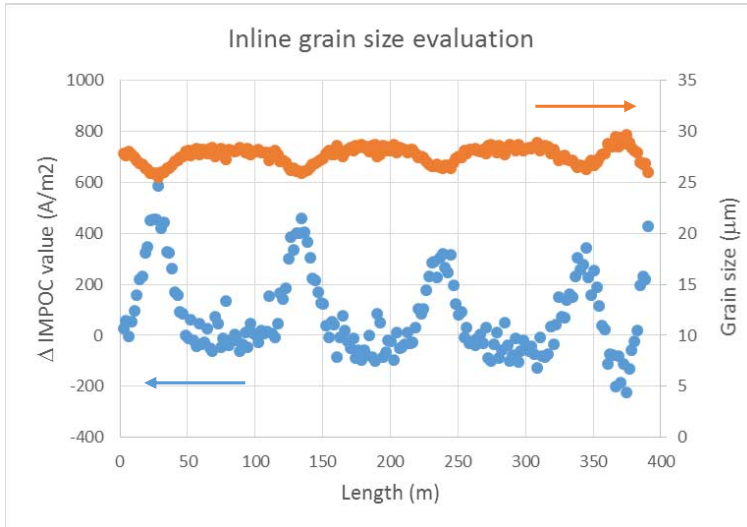


Figure 9. Microstructure evaluation over length using the electromagnetic IMPOC instrument. The blue markers (bottom, left axis) represent the deviation of the IMPOC data from the baseline. The orange markers (top, right axis) shows the grain size deduced from the measurement.

6. Conclusions & Outlook

Within the European collaboration project PUC, we have been able to develop direct models to predict magnetic behaviour and sensor response. Microstructure variations at grain size scale could be measured and modelled and appear to have significant effect on the magnetic properties, with good correspondence between models and experimental observations. We have shown that the 3D microstructure geometry modelling is a powerful tool to generate a multitude of well-defined microstructure variations. A next step is to broaden the microstructure model space to study the inversion problem, e.g. by the use of reduced models or surrogate modelling, in order to interpret the sensor measurements back to microstructure variances. From the relations between instrument data, fundamental electromagnetic properties and microstructure, we managed for certain steel grades to deduce the grain size over length from inline magnetic measurements.

Albeit significant progress have been made, there are still phenomena that need to be studied in a quantitative way, like the effects on the magnetic behaviour caused by microstructure details at the submicron scale, i.e. precipitates and dislocations.

Furthermore, texture, stress and depth gradients are potentially present in the material due to the rolling processes to which the steel strip is subjected. The incorporation of these effects is therefore also important for a good understanding of the measurement data. The large number of microstructure parameters that have influence on the magnetic properties underline that a deep knowledge of the production process and its resulting metallurgical variations is a prerequisite in the interpretation which microstructure variation(s) may cause the measured (magnetic) variations.

7. Acknowledgements

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Characterizing of Corrosion Backside of Steel Plates Using Extremely Low-Frequency Eddy Current Testing with Multiple-Frequency Magnetic-Field Exposure

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Abstract. In recent years, aging and deterioration of social infrastructure has become a major problem. A non-destructive testing technique for early detection of degradation (such as internal corrosion) is required to extend the lifetime of these structures. Currently, there are various methods used for inspection. Among these, non-destructive testing using magnetic fields is widely used due to advantages such as ease of use and high speed. However, magnetic inspection is generally employed only for inspection of the surface and subsurface, since it is difficult to inspect internal corrosion using this technique. Therefore, we have developed a device for detecting thinning of steel plates due to corrosion using multiple extremely-low frequencies and a high-sensitivity magnetic sensor. However, it is necessary to sweep multiple frequencies to obtain a frequency spectrum, and this is very time-consuming, especially in extremely low-frequency ranges. In this study, we have developed an analysis method based on a multiple-frequency applied magnetic field and FFT (fast Fourier transform) analysis of the detection signal. Thus, we have successfully reduced the measurement time while maintaining high detection accuracy.

1. Background

Recently, corrosion defects due to age-related deterioration of social infrastructure have become a serious problem. Therefore, non-destructive testing for early detection of these defects is required. Among the available methods, non-destructive testing using magnetic fields has advantages such as ease of use and high speed, and is therefore widely used [1]. However, due to the skin effect, the signal is greatly reduced as the distance from the surface increases [2]. Therefore, application of magnetic methods is limited to the surface and subsurface, and investigating internal corrosion remains difficult. Therefore, application to internal inspection has been studied [3], [4].

We have developed extremely low-frequency eddy current testing (ELECT) which enables internal detection by using extremely low-frequency magnetic fields and a high-sensitivity magnetic sensor. In this study, we use our technique to detect thinning of steel plates by corrosion. In addition, as a method of estimating the plate thickness, a spectrum analysis of the magnetic field (SAM) using the intensity and phase of the magnetic signal detected by sweeping multiple extremely-low frequencies is reported [5]-[7]. However, when performing ELECT-SAM, since the frequency is very low, sweeping multiple

frequencies is very time-consuming. In the previous report, we have swept a single applied magnetic field frequency and detected using a lock-in amplifier [8]. In this study, we have improved the measurement system to shorten the measurement time. Since the system combines multiple frequency applied magnetic fields and FFT analysis of the detection signal, it does not use a lock-in amplifier. Because of this improvement, a shortening of measurement time was possible while maintaining high detection accuracy equivalent to the previous report.

2. Methods

I. Measurement system

Figure 1 shows a schematic diagram of the measuring system, which consists of an oscillator, a current source, a magnetic measurement probe with an excitation coil, a cancelation coil, an anisotropic magnetoresistive (AMR) sensor, a sensor circuit, a DAQ (data acquisition module), and a PC. The multiple-frequency waveform is generated by the oscillator, the current is controlled by the current source, and samples are exposed to a multiple-frequency AC magnetic field induced by the excitation coil. The amplitude of each frequency in the multiple-frequency AC current was set to the same value, 65 mA. The secondary magnetic field, generated by the eddy current in the sample, is detected using the high-sensitivity AMR sensor. After the A/D (analog-to-digital) conversion of the detected signal with the DAQ, the intensity and phase of the detected signals at each frequency undergo FFT analysis. The built-in cancelation coil in the magnetic measurement probe is used to reduce exposure of the AMR sensor to the directly-coupled excitation magnetic field. The excitation coil and the cancelation coil have 50 turns and 10 turns, respectively. Samples of 12 mm-thick steel plate were used, with areas of 200 mm $\times$ 200 mm and different corrosion depths. The corroded, or reduced-thickness, area was 50 mm $\times$ 50 mm in the center of each plate. The thickness reduction is from 10% to 80%.

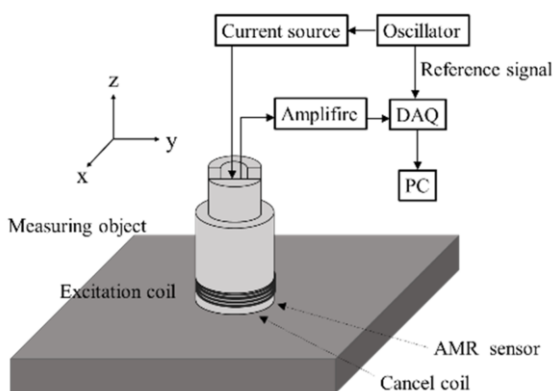


Figure 1. Measurement system

II. Spectrum analysis of magnetic field

ELECT uses an extremely low-frequency magnetic field to detect the steel plate thickness. Therefore, the magnetic field was measured with an AMR sensor with high sensitivity. The reported SAM technique was used to estimate the thickness. Figure 2 shows the concept of the SAM and estimation of plate thickness from analysis of the magnetic field. Because of sweeping frequencies we call the change of magnetic vectors as the magnetic-field spectrum. The magnetic-field spectrum is represented by the detected magnetic vectors of intensity and phase at each frequency. The start point of each spectrum is shifted to the origin to correct for the magnetization of the steel plate. Since the magnetic spectrum is thickness dependent, the thickness can be evaluated by comparing the magnetic field spectra at each measurement point.

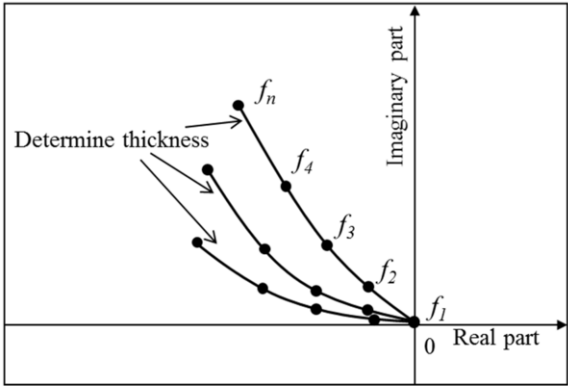


Figure 2. Magnetic-field spectrum

III. Multiple-frequency waveform and FFT analysis

Figure 3 shows the time waveform and frequency characteristics of the synthesized multiple-frequency waveform used in this study. The synthesized waveform, obtained by synthesizing five frequencies (1, 5, 10, 20, and 50 Hz), was used for the applied magnetic field. The current provided by the synthesized waveform was set to 65 mA at each frequency. To obtain a sufficiently high S/N ratio, the sampling frequency was set to 65536 Hz. Additionally, after receiving two cycles of the synthesized waveform from the DAQ, one cycle was FFT analyzed from the point where the half cycle phase was delayed. By performing this process, we ensure that FFT analysis is performed on a range which avoids both ends of the acquired signal and thus further reduce noise.

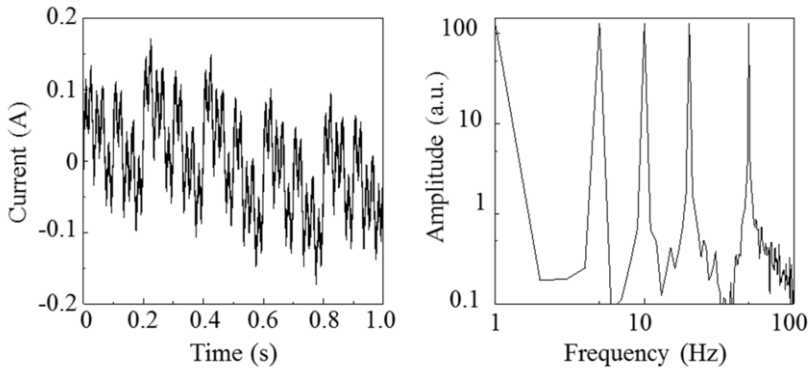


Figure 3. (a) time waveform and (b) frequency of waveform

3. Results

I. Evaluation of plate thickness

Six steel plates, SS400 and 12 mm thick and with different corrosion depths from 10% to 80% of total thickness were measured. Figure 4 shows the magnetic-field spectrum of the steel plates with different corrosion depths. As the corrosion depth increases, the spectrum curve shifts counterclockwise. From this measurement, we have obtained the spectrum change corresponding to corrosion depth. Therefore, we can estimate the thickness change caused by the corrosion.

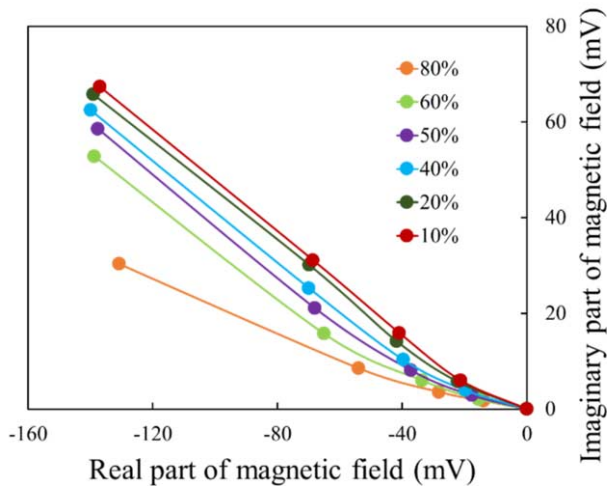


Figure 4. Magnetic-field spectrum of steel plates with different corrosion depths.

II. Imaging

For two-dimensional imaging of corrosion, an XYZ stage for moving the measurement probe was added to the measuring system shown in Fig. 1. Magnetic field were measured at several points to create an imaging of corrosion from the surface opposite the corrosion, i.e. a backside corrosion measurement was performed. The distance between the sensor and the measurement sample was 1 mm, and the 100 mm $\times$ 100 mm region in the center was measured at 10 mm intervals. To obtain the mapping that shows the corrosion area, it is necessary to evaluate the characteristics of the magnetic-field spectrum using one parameter. Therefore, after measuring with five frequencies from 1 Hz to 50 Hz, two frequencies suitable for thickness estimation were selected, and a differential vector was made. The magnetic vector at 1 Hz was subtracted from that at 20 Hz to obtain the differential vector. We use the phase parameter of the differential vector to reduce the effects of the steel's magnetization fluctuation. Figure 5 shows the results from the backside corrosion samples (steel plates with thickness reduction from 10% to 80%) using the differential vector of 20 Hz to 1 Hz. The dotted lines in Fig. 5 correspond to the actual position of the backside corrosion. As demonstrated in the figure, we have successfully imaged thickness changes due to backside corrosion of more than 20% using the phase change of the differential vector.

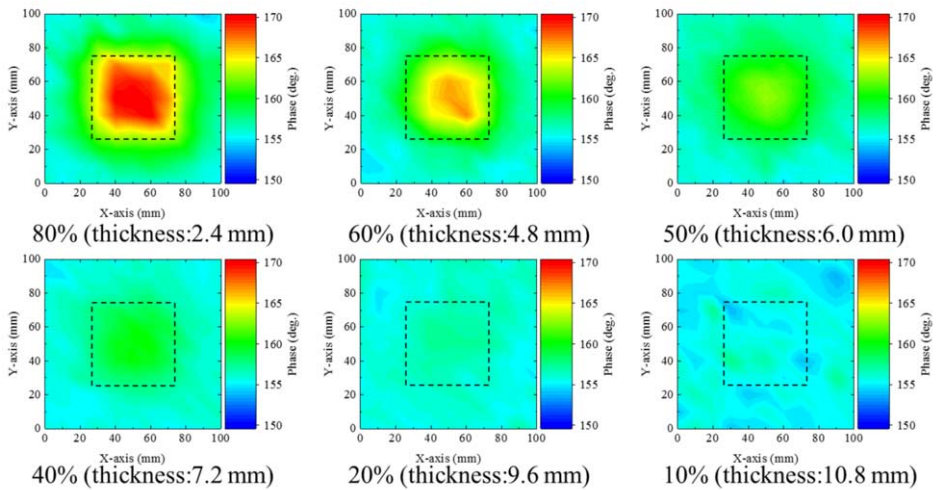


Figure 5. Backside corrosion imaging by subtraction of 20 and 1 Hz phase signals

4. Conclusion

For the ELECT-SAM method, we have developed an analysis method using multiple-frequency magnetic field exposure and FFT analysis. Using our method, we successfully imaged the backside corrosion of steel plate samples by plotting the local phase change of the magnetic-field spectrum, while reducing overall measurement time.

Acknowledgment

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Design of an UHF Antenna Insensitive to the Concrete Dielectric Characteristics

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Abstract. In this communication, we present a design methodology for an insensitive radiating structure for concrete health monitoring applications. It takes advantage of an optimized smaller vacuum box surrounding a patch antenna. The Ansys HFSS software is used to simulate the studies of different configurations. The patch antenna consists of a radiating metallic element with a ground plane printed on a low-cost FR4 substrate. It is optimized to be linearly polarized and to be able to operate in free space medium at the first two resonant frequencies of 1.55 GHz and 1.95 GHz. This study demonstrates that a 5 mm high vacuum box above main radiating antenna, is sufficient to reduce the influence of three surrounding dielectric materials studied in this work, such as a PVC, concrete and limestone. Theoretical predictions, for all types of surrounded materials, show reasonably good agreement with experimental results.

Keywords. Insensitive antenna, concrete medium, nondestructive testing, UHF band.

1. Introduction

The main cause of deterioration of reinforced concrete in civil engineering structures is damage resulting from steel corrosion. Corrosion occurs when some aggressive agents, particularly chlorides in marine environments, penetrate through the protective cover concrete layer by means of various transfer mechanisms.

Nondestructive testing (NDT) techniques have become a useful tool for infrastructure experts in order to establish accurate diagnosis [1]. Electromagnetic NDT techniques are sensitive to the presence of water and chlorides in concrete, unlike mechanical NDT that are rather sensitive to porosity and to the mechanical properties of concrete structures [2].

From a durability point of view, periodical survey using electromagnetic NDT techniques is necessary. That makes structural health monitoring (SHM) with embedded devices allowing wireless transmission an extremely interesting tool.

The global objective of the present study is to develop an insensitive antenna that is able to work in presence of surrounding concrete medium. In fact, the electromagnetic properties of the concrete are difficult to quantify exactly and may

vary with frequency and moisture content. Consequently, we fixed two criteria to select the optimized antenna technology: the topology and the operating frequency.

A patch antenna was chosen due to its weak electric field in the near-coupling area [3-4]. Once the antenna selected, we performed various parametric studies in order to minimize the coupling effect between the antenna and the surrounding media. This leads to introducing and optimizing a smaller vacuum box around the proposed antenna.

2. Concrete characteristics and electromagnetic' properties

Before presenting parametric studies leading to the choice of the antenna structure, the global characteristics of concrete have to be recalled. Hydraulic concrete is an heterogeneous mixture of various materials: sand, aggregates, cement, water, ... It is well known [5] that the dielectric permittivity of the mixture depends on the nature of the cement, as well on the mineralogical nature of the aggregates and the water-to-cement ratio (W/C). Moreover, the permittivity of concrete mixtures evolves over time (due to hydration and aging phenomena) and is highly sensitive to the water, chloride and carbonation ingresses [6-9].

Fig .1 shows an experimental result [10] of the frequency dependence for the complex relative permittivity of concrete at different humidity conditions (0, 50, 75 and 100%). The characteristics of the studied mixture are calcareous aggregates, CEMI cement + AEA (air-entrained agent), cement (350 kg.m^{-3}) and W/C = 0.35. Although this mixture has a low porosity (low W/C), we can see a notable dependence of the complex relative permittivity on the water content. Furthermore, it is important to note that it remains almost constant for frequencies higher than 500 MHz.

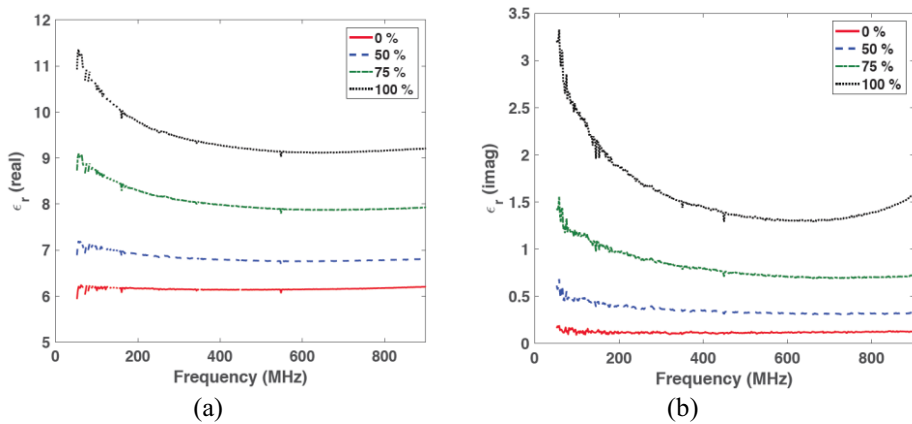


Figure 1. Real (a) and imaginary (b) parts of the permittivity of concrete versus frequency for four different humidity conditions.

As the globally dielectric properties of concrete depend on the mixture, on infrastructure aging, and on human/environmental impacts (uncontrolled phenomena), the main criterion for the antenna selection in our case should be its insensitivity to the varying dielectric properties of the surrounding medium.

3. Antenna structure

A patch antenna was chosen for its specific characteristics and, *inter alia*, a low electric in its near field zone [3]. It allows a very low coupling with the concrete surrounding medium and can meet the expectations of this study. Figure 2 shows the geometry of the used patch antenna. It consists of a metallic radiating element with ground plane printed on a low-cost FR4 substrate with relative permittivity $\epsilon_r = 4.5$. It was optimized by HFSS software [11] for efficient behavior in free space medium at the first two resonant frequencies around 1.55 GHz and 1.95 GHz. An example of the antenna positioning inside the reinforced concrete medium is shown in Fig. 3. This figure illustrates a smaller vacuum box surrounding the patch antenna. As the wireless transmission performance through a dielectric medium strongly depends on the coupling between the antenna and the material, we chose to minimize specifically the coupling conditions and their effects by optimizing the free space distance d between the antenna and the surrounding dielectric medium.

In the next sections, HFSS software simulations and experimental results will be compared, knowing that for both studies (numerical and experimental):

- the free space distance d varies from 0 to 50 mm in 1 mm steps.
- the relative permittivities of the transmission media were fixed to 2.7, 4.5, and 6.0, corresponding respectively to the permittivity values of PVC, limestone, and concrete.

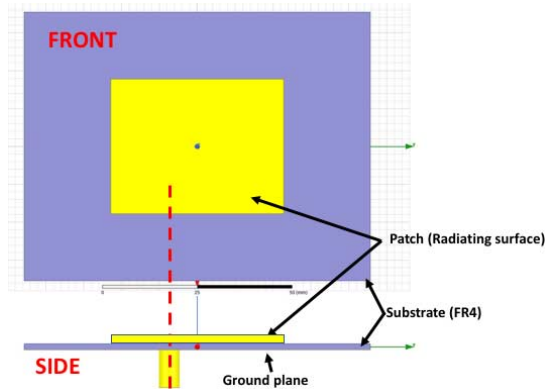


Figure 2. Optimized patch antenna layout: frontal view and side view.

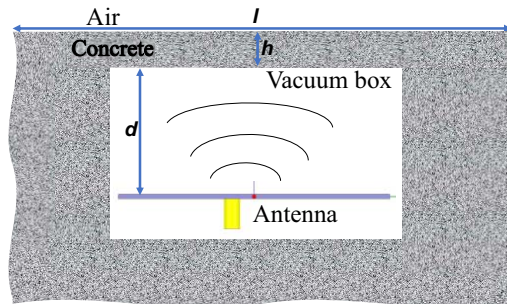


Figure 3. Geometry of study configuration

4. Results and discussion

4.1. HFSS simulation results

Fig. 4 shows HFSS numerical results corresponding to the evolution of the reflection coefficient (S_{11} parameter) of a patch antenna coupled to various modeled media (vacuum, PVC, limestone and concrete) at 3 positions ($d = 0$ mm, 5 mm and 50 mm).

When the antenna is close to the transmission medium (Fig. 4a), it can be clearly seen as permittivity increases, resonant frequencies decrease and matching is degraded. At $d = 5$ mm (Fig. 4b), a very small frequency shift can be observed, however there is still decrease the antenna matching between the vacuum case and the other media case and especially at the second resonant frequency (1.95 GHz). When the d distance between patch antenna and dielectric medium is further increased ($d = 50$ mm, Fig. 4c), almost no frequency shift is visible anymore. At least in the case of the first resonance mode, the mismatch observed for the embedded antenna is almost similar to that of the antenna in vacuum. Therefore, it can be stated that the antenna is insensitive to the varying dielectric properties of the transmission/coupling medium.

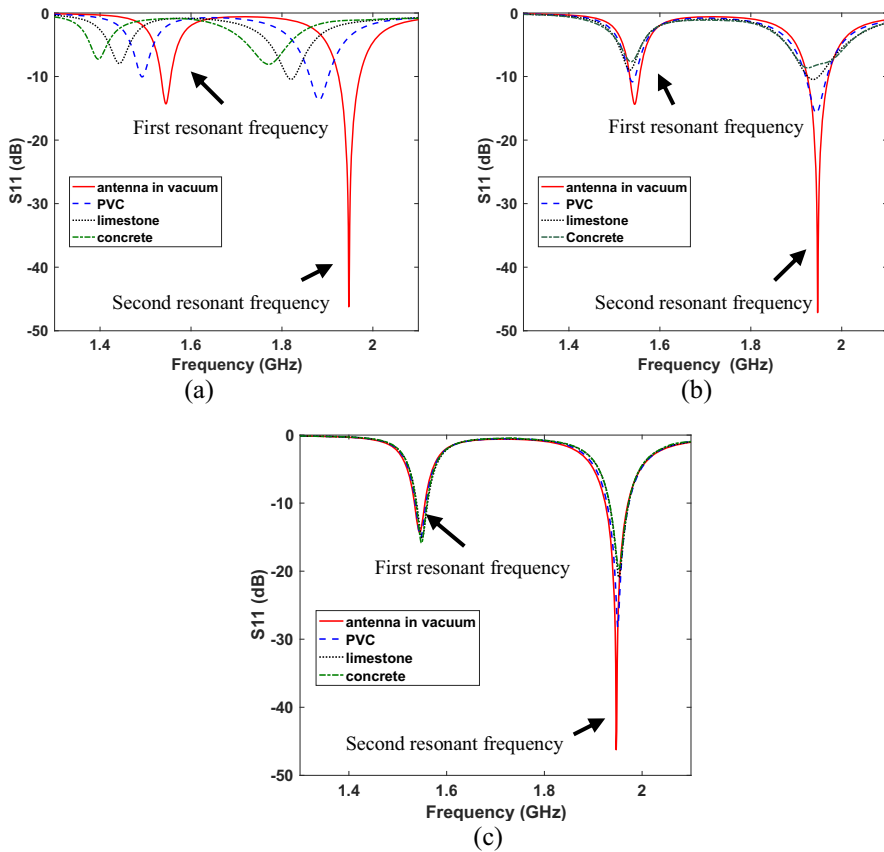


Figure 4. Simulated reflection coefficient S_{11} versus frequency for three distances d between the antenna and the three transmission medium studies. $d = 0$ mm (a), $d = 5$ mm (b), $d = 50$ mm (c).

Fig. 5 illustrates the evolution of the two first resonant frequencies as a function of the distance between the antenna and the considered transmission medium. The horizontal solid line (red line) corresponds to the vacuum resonance frequency, which can be considered as a reference.

For the first resonance modes, we can observe that both resonant frequencies are constant for all media at a distance d around 5 mm. At that specific distance, the resonant frequency is very close to that observed for the vacuum case. Except for large distances ($d > 50$ mm), where obtained vacuum resonant frequency is get closer and closer to one another. The distance $d = 5$ mm seems to be adapted for EM wave transmission with very low dependence on the varying dielectric properties of the transmission/coupling medium.

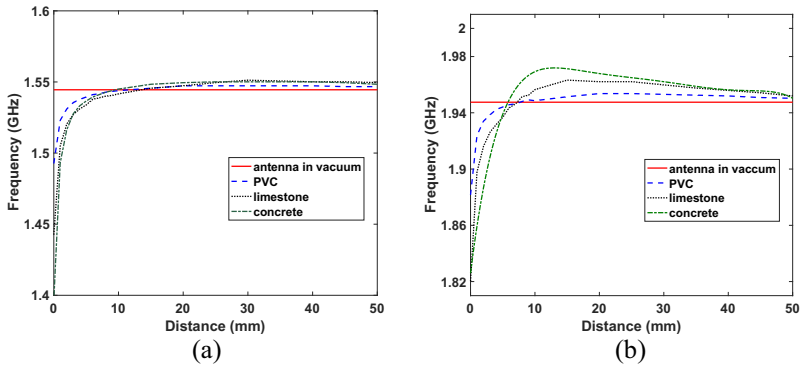


Figure 5. Antenna resonance frequencies versus distance between the antenna and the considered transmission media ($\epsilon_{r,PVC} = 2.7$, $\epsilon_{r,limestone} = 4.5$, $\epsilon_{r,Concrete} = 6.0$). First (a) and second (b) resonant frequency.

4.2. Experimental results

The patch antenna characteristics (in free space and coupled to PVC, limestone and concrete slabs) were experimentally determined. The dielectric characteristics of selected materials were measured using a coaxial transmission line. The PVC, concrete, and limestone slab dimensions were respectively $1\text{ m} \times 1\text{ m} \times 0.2\text{ m}$, $1\text{ m} \times 1\text{ m} \times 0.12\text{ m}$, and $0.5\text{ m} \times 0.5\text{ m} \times 0.2\text{ m}$.

The S_{11} measurements were done per millimeter up to $d = 50$ mm using an automatic experimental platform and a vector network analyzer (VNA).

The experimental evolution of the reflection coefficient (S_{11} parameter) of the patch antenna coupled to the studied media (vacuum, PVC, limestone and concrete) at three positions ($d = 0$ mm, 5 mm and 50 mm) is shown in Fig. 6.

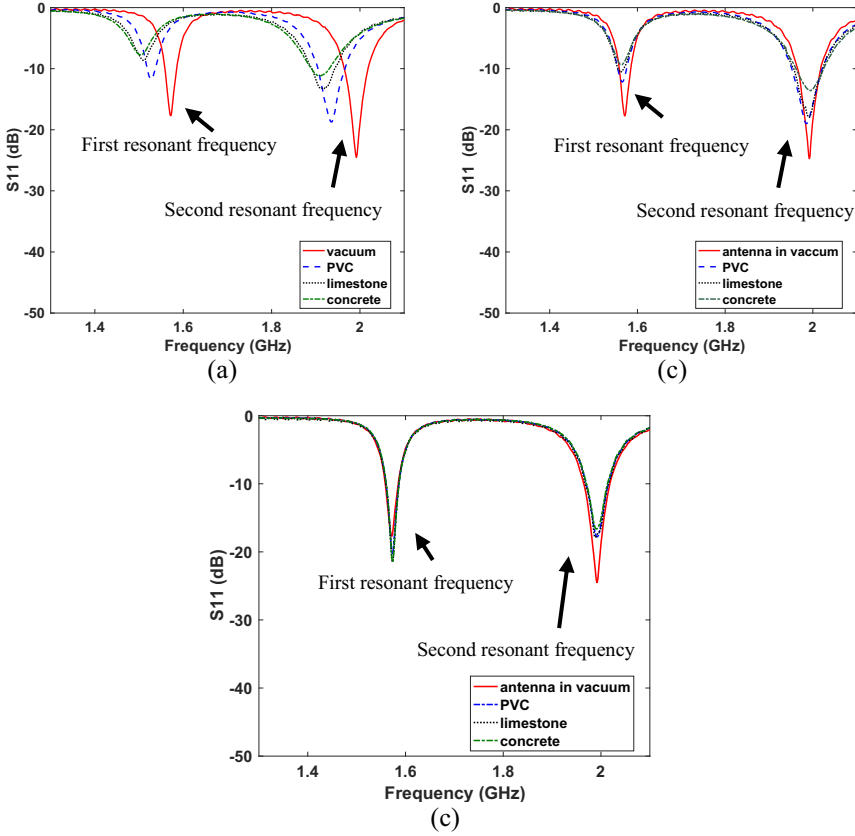


Figure 6. Measured reflection coefficient S_{11} versus frequency for three distances d between the antenna and the transmission medium as PVC, limestone, and concrete. $d = 0$ mm (a), $d = 5$ mm (b), $d = 50$ mm (c)

The experimental results are very similar to those of the numerical study even though the relative frequency shifts are smaller for limestone and concrete (Fig. 6a). Limestone is rather sensitive to the ambient humidity that cannot be controlled, thus probably causing modified dielectric properties compared to the initial characterization. The fabricated concrete slab is not homogenous unlike what was defined in the HFSS simulation.

At $d = 5$ mm (Fig. 6b), only a small frequency shift can be observed, however, there is still an increase in impedance mismatch for the three used media. When the antenna/medium distance is further increased ($d = 50$ mm, Fig. 6c), almost no frequency shift is visible anymore, which is in accordance with the numerical results. We can conclude that above $d = 5$ mm, the studied patch antenna is insensitive to the varying dielectric properties of the environmental medium.

The evolution of the resonant frequencies as a function of the distance between the antenna and the transmission medium for the three modeled media is illustrated in Fig. 7. Again, the horizontal solid line (red line) corresponds to the vacuum resonance frequency.

For the first resonance modes, we can observe a both resonant frequencies are constant for all media at a distance d around 5 mm. At this specific distance, the

resonance frequency is very close to that observed in the vacuum case. In agreement with the simulation results, the 5 mm high vacuum box seems to be adapted for the proposed antenna without any dependence on the varying dielectric properties of the surrounding medium.

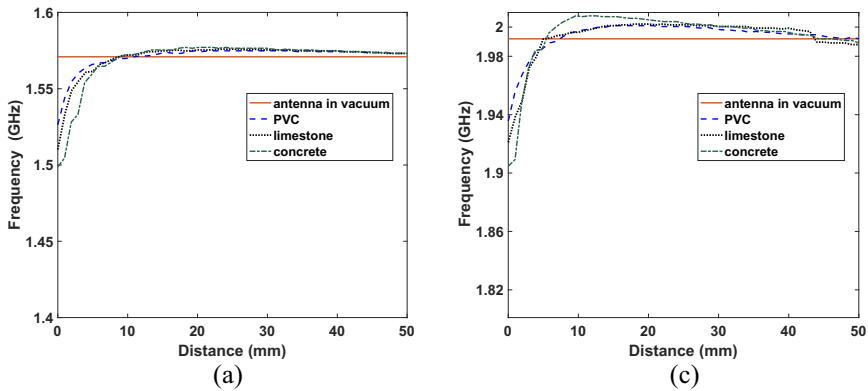


Figure 7. Measured antenna resonance frequencies *versus* distance between the antenna and the surrounding media. First (a) and second (b) resonance frequency.

Table 1 provides numerical and experimental results for the case study where $d = 5$ mm (Figs. 4b and 6b). This highlights the very good correspondence between simulations and experiments results. We can especially note that the chosen patch antenna is insensitive to the varying dielectric properties of the surrounding medium for a 5 mm high air gap of in above the antenna.

Table 1. S_{11} magnitude, resonant frequencies, and frequency shift Δf with respect to free space transmission for the different antenna surrounding media. Exp = experiment, Sim = simulation.

	FIRST RESONANCE FREQUENCY						SECOND RESONANCE FREQUENCY					
	S_{11} (dB)		f_1 (GHz)		$\frac{\Delta f(\%) = f_1 \text{ vacuum} - f_1 \text{ medium}}{f_1 \text{ vacuum}}$		S_{11} (dB)		f_2 (GHz)		$\frac{\Delta f(\%) = f_2 \text{ vacuum} - f_2 \text{ medium}}{f_2 \text{ vacuum}}$	
					Sim	Exp					Sim	Exp
	Sim	Exp	Sim	Exp	Sim	Exp	Sim	Exp	Sim	Exp	Sim	Exp
Vacuum	-14.37	-17.72	1.54	1.57	-	-	-47.14	-24.72	1.95	1.99	-	-
PVC	-10.83	-12.17	1.54	1.56	0.3%	0.4%	-15.76	-19.03	1.94	1.99	0.2%	0.4%
Limestone	-8.99	-10.58	1.53	1.56	0.7%	0.4%	-10.47	-17.97	1.94	1.99	0.5%	0,0%
Concrete	-7.66	-9.35	1.53	1.56	0.5%	0.4%	-7.66	-13.58	1.93	1.99	1.1%	0,0%

5. Conclusions

The main goal of the present study was to obtain an insensitive antenna configuration to allow precise transmission across a concrete structure. This includes only little dependency on the dielectric properties of its surrounding medium, which vary with

humidity, aging and several environmental conditions like pathogen agents. The performed simulation and experiments confirm that a 5 mm high vacuum box above main radiating antenna, allows the antenna to be insensitive to the surrounding dielectric characteristics. This is due to the intrinsic characteristic of a patch antenna, which radiate a weak electric in its near-field zone.

Under the investigated conditions, a frequency shift of less than 1% around the first two resonant frequencies (1.55 GHz and 1.99 GHz) first of the antenna has been obtained, proving the insensibility of the antenna to the surrounding medium.

This study was focused on the analyses of the reflection coefficient of the antenna surrounded by homogenous dielectric medium. In the future, the radiating characteristics of the embedded antenna, taking in to account medium losses, will be studied.

6. Acknowledgements

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Vibration Parameters Based Debonding Defect Detection for Sandwich Plate with Pyramidal Truss Cores

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Abstract. As a kind of super-light lattice material, the sandwich plate with pyramidal truss cores has wide application prospect. In this paper, a vibration parameter based nondestructive testing method using Uniform Load Surface (ULS) curvature was proposed to detect the delamination defects of the pyramid-type lattice sandwich plate. The feasibility of the approach is that the surface smoothing method and curvature mode change rate method are adopted to overcome the dependence on the unflawed model. The validity and efficiency of the proposed methods were demonstrated through numerical simulations.

Keywords. Sandwich plate with pyramidal truss cores, debonding defect, vibration characteristics, difference method, surface smoothing method, curvature mode change rate.

1. Introduction

As a kind of super-light material, the sandwich plate with pyramidal truss cores (Fig. 1-2) has been applied in many fields. The welding defects, especially the delaminations between the inner layer and surface layer occurred during processing and service process, are the major concern for the quality control of the lattice sandwich plate as it can reduce the strength significantly [1-3]. Therefore, a reliable and convenient NDT technique is required to guarantee the safety of a structure made by the pyramid-type lattice sandwich plate. Though many method exist for detecting damage in a structure, few of them can be used to identify the debonding defects ideally. Recently, based on Uniform Surface Load (ULS) [4], Wu and Law [5] have proposed a strategy to detect damage in a plate by investigating changes in the ULS curvature that is interpolated with central difference method and Chebyshev approximation. Tian and Chen [6-7] applied this difference method to detect the debonding defect and identified all of the damages of difference numbers and locations successfully. However, those methods need the parameters of undamaged structures as the reference signal. To overcome the

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dependence problem, based on ULS curvature, two methods independent of original intact structures were adopted to identify the debonding defects of lattice sandwich plate in this paper. One is surface smooth method and the other is curvature mode change rate method. The feasibility of the approaches were investigated through numerical simulation by using commercial software for different damage configurations.

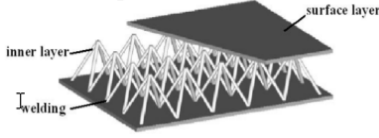


Figure 1. An example of lattice sandwich plate

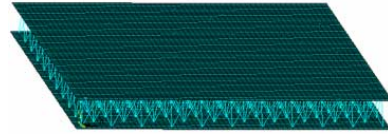


Figure 2. Finite element model of lattice sandwich plate

2. Numerical methods of damage detection for lattice sandwich plate

2.1. Damage detection algorithms based on difference method

The debonding defect of the pyramid-type lattice sandwich plate, which is usually caused by stress concentration around the welding zone during service, causes changes in structural physical properties. These changes in turn alter the dynamic response behavior of structure from its initial pre-damage condition. Therefore different static and dynamic characters will be presented between the damaged and normal structure.

To detect delaminations of the lattice sandwich plate, numerical simulation by using commercial software for different damage configurations were completed at first. As the second step, the ULS curvature values can be calculated from the measured modal parameters by using central difference method [6, 8]. The accuracy of the central difference method is well known depending on the density of the measurement grid. If the ULS values are estimated on a sparse grid, it will induce a very large error in calculating the curvature from differentiation [7]. The following Chebyshev polynomial in two variables was adopted to model the ULS distribution so as to avoid the error:

$$u(x, y) = \sum_{i=1}^N \sum_{j=1}^M C_{ij} T_i(x) T_j(y), \quad (1)$$

where, $T_i(x)$, $T_j(y)$ are the standard Chebyshev polynomials with the plane domain of $[-1, 1]$, N , M are their orders, and $u(x, y)$ is the uniform load surface at point $k(x, y)$.

$u(k) = \sum_{r=1}^m \frac{\phi_r(k) \sum_{l=1}^n \phi_r(l)}{\omega_r^2}$, ϕ_r is the mass normalized mode, ω_r is the r -th natural frequency.

For $P=N \times M$ measuring points, it can be written in a matrix form:

$$\{u(x_i, y_j)\}_{P \times 1} = [T_i(x) T_j(y)]_{P \times P} \{c_{ij}\}_{P \times 1}, \quad (P = N \times M), \quad (2)$$

The coefficient vector $\{c_{ij}\}_{P \times 1}$ can then be solved as

$$\{c_{ij}\}_{P \times 1} = [T(x_i) T(y_j)]_{P \times P}^{-1} \{u(x_i, y_j)\}_{P \times 1} \quad (3)$$

Then the ULS curvature can be approximated by the second derivatives of the Chebyshev polynomials in Eq. (1) as

$$\begin{aligned}
u_{xx}(x, y) &= \sum_{i=1}^N \sum_{j=1}^M c_{ij} \frac{\partial T_i^2(x)}{\partial x^2} T_j(y), \\
u_{yy}(x, y) &= \sum_{i=1}^N \sum_{j=1}^M c_{ij} T_i(x) \frac{\partial T_j^2(y)}{\partial y^2}, \\
u_{xy}(x, y) &= \sum_{i=1}^N \sum_{j=1}^M c_{ij} \frac{\partial T_i(x)}{\partial x} \frac{\partial T_j(y)}{\partial y}
\end{aligned} \tag{4}$$

By subtracting the curvature parameters of the intact state from those of the damage state, the damage index map can be formulated as follows,

$$d(x_i, y_j) = \left[\alpha_{xx} |u_{xx}^D - u_{xx}| + \alpha_{yy} |u_{yy}^D - u_{yy}| + \alpha_{xy} |u_{xy}^D - u_{xy}| \right]^2 \tag{5}$$

For undamaged structure, the difference between ULS curvatures of intact structure and damage structure is only due to measurement noise. Therefore, values of the damage index map, $d(x_i, y_j)$, slightly oscillate around zero without any distinct peak. In the contrast, peaks or slopes will clearly show up at damaged zone of the plate if the structure is damaged.

2.2. Damage detection algorithms based on surface smoothing method

Most damage detection based on a baseline or healthy data is usually unreliable for practical application, because a priori measurement in a healthy condition is affected a lot by the surroundings. To avoid this difficulty, Ratcliffe [9] and Wu [5] proposed the ‘‘Gapped-smoothing’’ technique to detection damages in beam and plate structure respectively. The basic idea of the method is that the ULS curvature of the sandwich plate, without any damage, has a smooth surface. The surface smoothing method can eliminate the peaks hiding in the ULS curvature caused by debonding defects. Therefore, the approximate ULS curvature without debonding defect can be established for the difference method.

Based on the surface smoothing method, the smoothed ULS curvature at each point (x_i, y_j) can be approximated by the following formula,

$$u_{fitting}(x, y) = \sum_{i=0}^2 \sum_{j=0}^2 C_{ij} x^i y^j \tag{6}$$

where the coefficient c_{ij} for an approximate ULS curvature without debonding defect can be evaluated by a curve-fitting process. Particularly, to obtain the smoothed ULS curvature at point (x_i, y_j) as shown in Fig. 3, curvature data at all the adjacent points, but not the point (x_i, y_j) itself, are used to evaluate the coefficient c_{ij} in Eq (6). This process is repeated for each measuring point to give a smooth ULS curvature to model the undamaged sandwich plate structure.

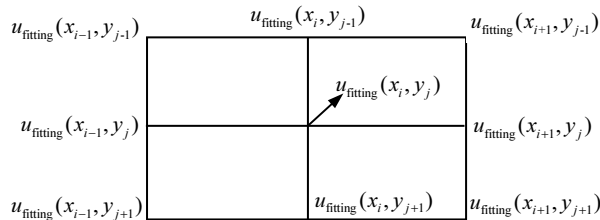


Figure 3. Gapped grid for curve fitting at measuring point (x_i, y_i)

So the presence of the peak in the ULS curvature due to debonding defect can then be detected by subtracting the smoothed curvature from the estimated curvature of the damaged structure. The damage index map is given as:

$$d(x_i, y_j) = \left(u_{measured}(x, y) - u_{fitting}(x, y) \right)^2 \quad (7)$$

2.3. Damage detection algorithms based on curvature mode change rate

The same to the surface smoothing method, the method based on curvature mode change rate need not the parameters of the initial intact structures. Zhang, et al [10] has applied this method to a beam structure and got ideal detection results. In this paper, the detection method based on curvature mode change rate was adopted to the debonding defect identification for four-side simply supported sandwich plate by using the ULS curvature values.

The basic idea of this method is that the local stiffness degradation caused by delamination will be expressed by the volatility of change rate of ULS curvature. As the corresponding relationship between curvature mode and displacement mode, the change rate of ULS curvature is relative to structure position. Therefore, the change rate of ULS curvature μ''' can be used to detect the local debonding defects. The calculation of μ''' is as follows,

$$\mu_k'' = \frac{\dot{\mu}_{k(i+1)} - \dot{\mu}_{ki}}{h} \quad (8)$$

where, $\mu_{k(i+1)}''$ is the ULS curvature mode of the modal vector coefficient at the $(i+1)$ measurement point of the k -th unit mass-normalized mode vector. h is the uniform grid spacing in the corresponding direction.

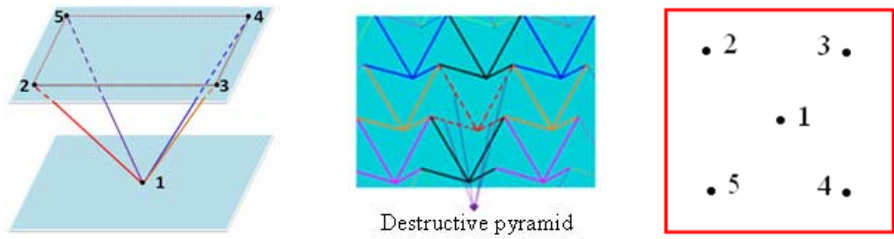
For the delamination identification of the pyramid-type lattice sandwich plate, the defect information will be extracted by the change rate curve of ULS curvature mode.

3. Results of Numerical simulation for damage detection

3.1. Finite element model

Concrete pyramid-type lattice sandwich plates with delaminations of different configurations were adopted as examples to demonstrate the efficiency of the above methods. The constraint condition with four-side simply supported was selected for the numerical simulations.

Lattice truss structures are usually made by perforating metal sheets to create a periodic array of diamond shaped holes [11]. Therefore, it can be consider in a periodic pyramidal structure of conjoint four beams cells as shown in Fig. 4(a).



(a) Periodic pyramidal structure (b) Diagram of the simplified delamination (c) Relationship of the five nodes

Figure 4 Periodic pyramidal structure and the diagram of the simplified delamination

As a preliminary study, four beams which are bounded to a welding point were deleted from the finite element model to simulate a debonding defect (Fig. 4(b)). The detection of the delamination is very difficult as the complexity of the topological structure. However, the dynamic characteristics of the corresponding position on the plate change with the deletion of the inner trusses. This characteristic can be used as the basis for the debonding defect identification. As shown in Fig. 4, once such a defect occurs, the dynamic response of the five nodes that are directly related to the debonding will change significantly. Therefore, the key point of the damage recognition of the lattice sandwich plate with pyramidal truss cores is to detect the location of these five nodes.

Figure 2 shows the 3D finite element model of the pyramid-type lattice sandwich plate founded by using commercial software ANSYS. The numerical models used for feasibility analysis of the proposed method are shown in Fig.5, where, (a) is the structure of the intact plate, (b) and (c) are models with 2 and 3 debonding defects.

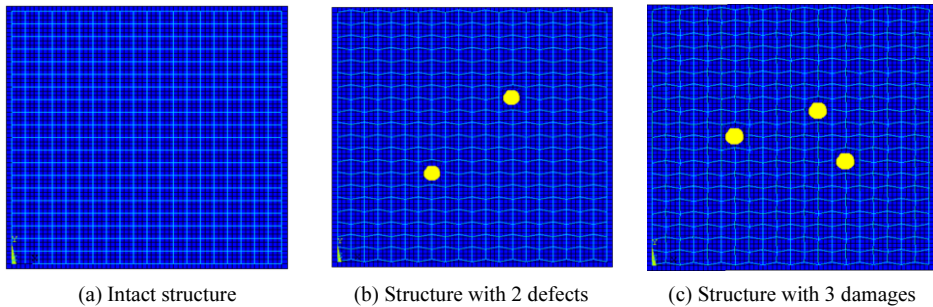


Figure 5 Numerical models for feasibility analysis

3.2. Damage identification results

Based on the mentioned methods, numbers of numerical simulations were completed. In this paper, only the displacements of translational DOF in direction perpendicular to the surface plate were used to calculate the ULS curvature. Eqs. (5), (7), (8) were used to calculate of the damage index based on changes in the ULS curvature. The results together with the results of central difference method are shown in Fig. 6, Fig.7, Fig.8 and Fig.9.

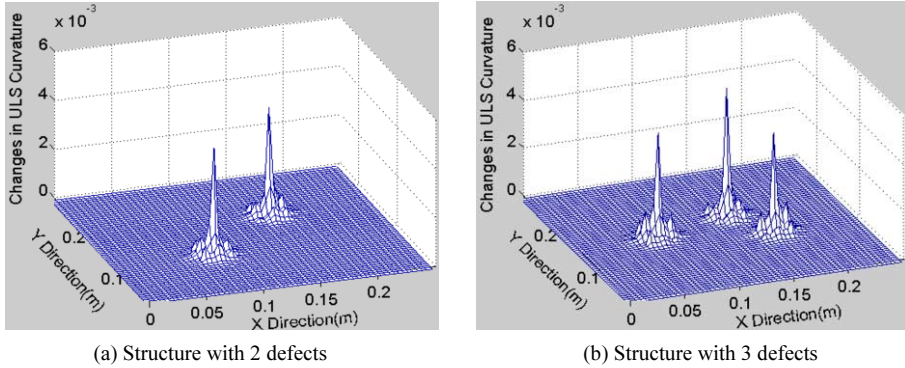


Figure 6. Results of central difference method for four-side simply supported lattice sandwich plate

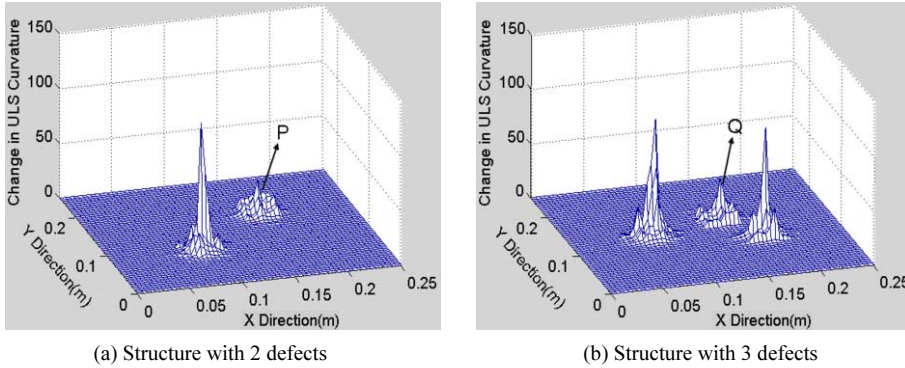


Figure 7. Results of Chebyshev polynomial method for four-side simply supported lattice sandwich plate

From Fig. 6 and Fig. 7, it can be clearly seen that there is a sharp peak located at each debonding defect. For a four-side simply supported plate, both maximum bending moment and flexural displacement occur at the geometrical center of the plate, they will decrease when the geometrical position moving to the boundary of the plate. The influence of damage to ULS curvature shows the same feature. In the case with same damage form and magnitude, the damage index shows different size.

Comparing the results of the two above, one can find that the damage index of the latter is much bigger than the former. So the latter method seems to be more sensitive than the former in delamination defect recognition of the lattice sandwich plate with pyramidal truss cores. However, in Fig. 7, some of the damage indexes are smaller than they should be, such as point P and Q in the figures. That's because the method of data block processing was utilized in the calculation to quickly deal with data. Then, error will unavoidably exist in the data of the junction. If the delamination just located at the position of the segmentation, error will occur in the defect identification index to make the indexes smaller than they should be as shown in Fig. 7.

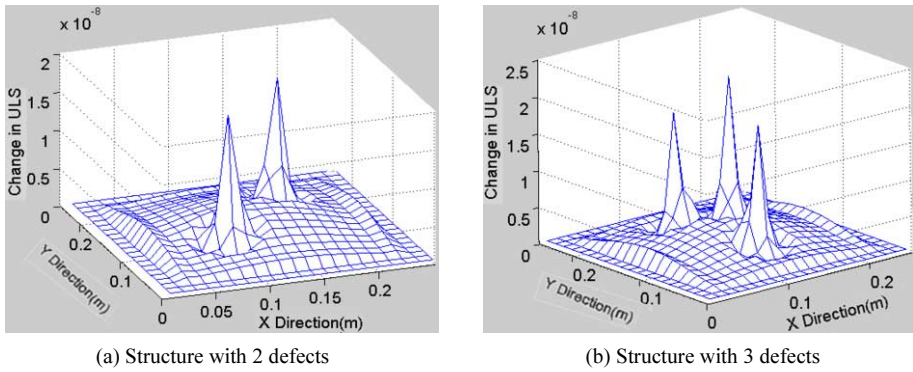


Figure 8. Results of surface smoothing method for four-side simply supported lattice sandwich plate

As shown in Fig. 8, independent of the unflawed model, the surface smoothing method can be used to identify the delaminations of the pyramid-type lattice sandwich plates effectively. Though there's some noises exist in the recognition results, the number and position of the debonding defects can be properly identified.

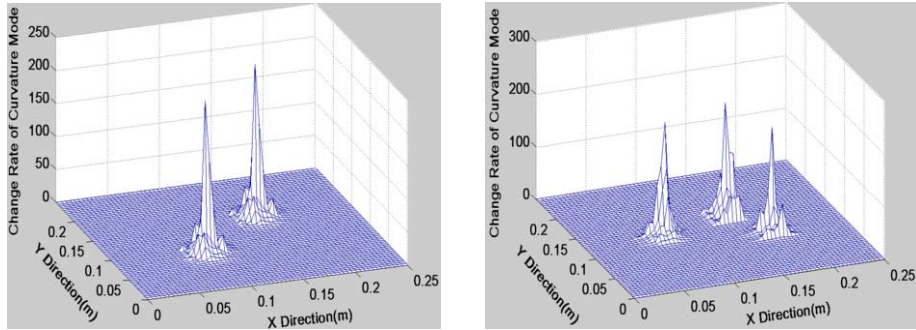


Figure 9. Results of curvature mode change rate for four-side simply supported lattice sandwich plate

It can be clearly seen in Fig. 9 that all of the delaminations are recognized accurately by using of the ULS curvature change rate. Comparing with the results of surface smoothing method, the defect identification index is much bigger. So the latter method seems to be more sensitive than the former in debonding defect recognition of the pyramid-type lattice sandwich plate.

4. Conclusion

In this paper, based on the ULS curvature calculated by using Chebyshev polynomial, difference method, surface smoothing method and curvature mode change rate method are applied in delamination detection of the lattice sandwich plate with pyramidal truss cores. Numbers of numerical simulations are performed to verify the efficiency of the proposed methods. The major conclusions obtained are as follows:

1) All of the processing methods and numerical codes proposed in this paper were demonstrated to be effective for the debonding defect identification of sandwich plate with pyramidal truss cores.

2) The difference method based on Chebyshev polynomial approximation is more sensitive than the method of central difference method.

3) Independent of the unflawed model, the identification method based on surface smoothing method and change rate of ULS curvature can be used to identify the delaminations of the lattice sandwich plates effectively. They will have more wide application in practice than the difference method.

4) The method based on change rate of ULS curvature is more sensitive than the surface smoothing method.

Acknowledgements

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Sensitivity Analysis Using a Sparse Grid Surrogate Model in Electromagnetic NDE

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Abstract. The global sensitivity analysis of electromagnetic nondestructive evaluation (NDE) by means of Sobol' indices are considered in this work. To reduce the computational burden, a sparse grid surrogate model is used. The latter can simply replace the true simulator to some extent, but it can also be used to numerically evaluate the integrals defining the Sobol' indices. In most of the NDE setups, the output is not a scalar quantity but functional data (e.g., a surface scan); a method is presented to take this into account. The sparse grid based sensitivity analysis is compared to classical techniques via examples drawn from electromagnetic NDE.

Keywords. sensitivity analysis, Sobol' index, sparse grid

1. Introduction

To characterize the uncertainty of the output of a simulation model due to its uncertain input parameters, Sobol' indices are commonly used. However, the calculation of the Sobol' indices can be computationally demanding, especially in the case when a heavy simulator is considered and/or the number of input parameters is high. The application of surrogate models provides a way to reduce this computational burden. Polynomial Chaos Expansion (PCE) supplies an efficient, low-cost technique to compute the Sobol' indices. Recently, the use of sparse grids (SGs) in electromagnetic nondestructive evaluation (ENDE) has been proposed [1]. SG can be used both as an approximation of the model (e.g., for Monte Carlo method) and as a numerical quadrature to evaluate integrals involved in the calculation of Sobol' indices. The surrogate model-based sensitivity analysis in the context of ENDE is considered, e.g., in [2]. The present work aims at (i) comparing different surrogate model based approaches of Sobol' index calculation for ENDE problems and (ii) extending the formulations to the case when the output consists of functional data, being typical in ENDE (e.g., surface scan of an eddy-current probe).

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2. Sensitivity Analysis

In this section a brief overview is given on the purpose of the sensitivity analysis, particularly with respect to the method of Sobol' indices as well as the possibilities of its numerical calculation.

Suppose that we have a mathematical model $y = f(\mathbf{x})$ as a multivariate, square integrable function over a domain defined by a set of independent variables, $\mathbf{x} = (x_1, \dots, x_n)$. In general the aim of sensitivity analysis is the quantification of how the uncertainty of the model output is affected by the uncertain input parameters. The *black-box* approach is commonly used, i.e., one relies only on certain number of input samples and the corresponding output ones. This approach provides the flexibility of applications in wide range of industrial and scientific fields.

2.1. Sobol' indices

The technique of Sobol' indices is a variance-based sensitivity analysis method. After its introduction in [3], by now it has become a widely used approach in several fields. A detailed presentation is given in, e.g., [4]. A brief summary is given below, such that the novelty of the present work can be clearly pointed out.

Without the loss of generality we can assume that all input variables are uniformly and independently distributed in the unit hypercube $\mathcal{D}_{\mathbf{x}} = [0, 1]^n$. The key idea is to decompose $f(\mathbf{x})$ into the sum of subfunctions of increasing dimension, such that

$$f(\mathbf{x}) = f_0 + \sum_{i=1}^n f_i(x_i) + \sum_{1 \leq i < j \leq n} f_{ij}(x_i, x_j) + \dots + f_{1,2,\dots,n}(x_1, x_2, \dots, x_n) \quad (1)$$

with f_0 being the expected value of $f(\mathbf{x})$ and the integral of the subfunctions with respect to any of their arguments is zero, i.e., the subfunctions are pairwise orthogonal. This Sobol-decomposition is shown to be uniquely exist in [3], and the subfunctions are defined in a recursive manner:

$$f_0 = \int_{\mathcal{D}_{\mathbf{x}}} f(\mathbf{x}) d\mathbf{x} \quad (2a)$$

$$f_i(x_i) = \int_0^1 \dots \int_0^1 f(\mathbf{x}) d\mathbf{x}_{\sim i} - f_0 \quad (2b)$$

$$f_{ij}(x_i, x_j) = \int_0^1 \dots \int_0^1 f(\mathbf{x}) d\mathbf{x}_{\sim i,j} - f_0 - f_i(x_i) - f_j(x_j) \quad (2c)$$

where $x_{\sim i}$ is a set without x_i . In addition the variance of $f(\mathbf{x})$ can be partitioned into the sum of sub-variances:

$$D = \sum_{\mathbf{v} \subseteq \mathbf{x} \setminus \{0\}} D_{\mathbf{v}}, \quad (3)$$

with $D_{\mathbf{v}}$ denoting the variance of the subfunction described by a group of variables $\mathbf{v} = (x_{i_1}, x_{i_2}, \dots, x_{i_s})$. Each sub-variance can be considered as the contribution of a group

of variables to the total variance. The measure of their importance, namely the Sobol' indices thus can be defined as

$$S_{\mathbf{v}} = D_{\mathbf{v}} / D. \quad (4)$$

The sum of all possible indices is 1, providing us a simple interpretation of their meaning. Groups consist of only one variable yield the 1st order indices. The 2nd order ones mean the effects caused by the interaction of two parameters without their 1st order effects; higher order indices can be analogously defined. Traditionally, the 1st order indices are estimated by Monte Carlo formulas:

$$\hat{f}_0 = \sum_{k=1}^M f(\mathbf{x}^{(k)}) \quad (5a)$$

$$\hat{D} = \sum_{k=1}^M f(\mathbf{x}^{(k)})^2 - \hat{f}_0^2 \quad (5b)$$

$$\hat{D}_i = \sum_{k=1}^M f(x_i^{(k)}, \mathbf{x}_{\sim i}^{(k)}) f(x_i^{(k)}, \mathbf{x}'_{\sim i}^{(k)}) - \hat{f}_0^2 \quad (5c)$$

where $\mathbf{x}^{(k)}$ is the k -th representation of the M samples and $\mathbf{x}'_{\sim i}^{(k)}$ denotes a sample independent from $\mathbf{x}_{\sim i}^{(k)}$

2.2. Extension of Sobol' indices

In the original framework of Sobol' indices, the output is a scalar function and the variables have to be independent. In certain cases this can be sufficient (e.g., scalar-output indices for POD studies), however, several NDE related applications cannot be directly treated due to these limitations. For example, vector output functions occur when a surface scan of impedance variation is considered or geometric constraints apply for the defect parameters, making them dependent on each other. Therefore, the original definition needs to be extended and generalized regarding the above cases, which is the main contribution of this work.

2.2.1. Multiple output functions

Supposing we have a multiple output function $\mathbf{F}(\mathbf{x}) = (F_1(\mathbf{x}), F_2(\mathbf{x}), \dots, F_P(\mathbf{x}))$, a natural extension of Sobol' indices is taking the average of the previously calculated indices of the component functions, i.e.,

$$S_i^{avg} = \frac{1}{P} \sum_{j=1}^P S_{F_j}^{(i)}. \quad (6)$$

This method has its drawback by not taking into account the vectorial nature of the output and the correlation between the component functions. A more suitable solution requires the definition of global indices to the entire functional output. A method with

this consideration has recently been introduced in [5]. Hereby we give another, simpler approach to extend the 1st order indices, relying on an equivalent stochastic definition:

$$S_i = \text{Var}_{x_i} [\mathbb{E}_{\mathbf{x}_{\sim i}} [f(\mathbf{x})|x_i]] / \text{Var} [f(\mathbf{x})]. \quad (7)$$

In 1 dimension (1-D), variance can be considered as the expected value of a squared error function, defined by the deviation of $f(\mathbf{x})$ from its own expected value. In N -D, variance can be defined analogously with the error being a vector and the 2-norm is used as a metric of distance instead of absolute value:

$$D^{\text{vec}} = \mathbb{E} [\|\mathbf{h}\|_2^2] = \mathbb{E} [\|\mathbf{F}(\mathbf{x}) - \mathbf{F}_0\|_2^2], \quad (8)$$

where $\mathbf{F}_0$ denotes the vectorial expected value of $\mathbf{F}(\mathbf{x})$. This leads us the definition of vectorial Sobol' indices:

$$S_i^{\text{vec}} = \text{Var}_{x_i} [\mathbb{E}_{\mathbf{x}_{\sim i}} [\mathbf{F}(\mathbf{x})|x_i]] / \text{Var} [\mathbf{F}(\mathbf{x})] = D_i^{\text{vec}} / D^{\text{vec}}. \quad (9)$$

Monte Carlo estimators are also extended to vectorial outputs as

$$\hat{\mathbf{F}}_0 = \frac{1}{M} \sum_{k=1}^M \mathbf{F}(\mathbf{x}^{(k)}) \quad (10a)$$

$$\hat{D}^{\text{vec}} = \frac{1}{M} \sum_{k=1}^M \|\mathbf{F}(\mathbf{x}^{(k)})\|_2^2 - \|\hat{\mathbf{F}}_0\|_2^2 \quad (10b)$$

$$\hat{D}_i^{\text{vec}} = \frac{1}{M} \sum_{k=1}^M \mathbf{F}(x_i^{(k)}, \mathbf{x}_{\sim i}^{(k)}) \bullet \mathbf{F}(x_i'^{(k)}, \mathbf{x}_{\sim i}'^{(k)}) - \|\hat{\mathbf{F}}_0\|_2^2 \quad (10c)$$

with $\bullet$ denoting the scalar multiplication. Though the squared Euclidean norm fits well for our purpose, we note that this choice is not obvious. In the case of a sparse output signal—which is typical, e.g., when a temporal echo is recorded in an ultrasonic testing method—other norms (such as the maximum norm) might be preferred. One may also consider an appropriate pre-processing of the signal before using the Euclidean norm.

2.2.2. Dependent variables

A method of sensitivity analysis over non-rectangular domain has recently been studied in [6]. Herein we introduce a method based on the *Rosenblatt transform* (RT) proposed in [7]. Let us denote with $\mathbf{x}^*$ a permutation of $\mathbf{x}$ and with $F_i(x_i^*|\mathbf{v})$ being the conditional cumulative distribution function of x_i^* with respect to a subset $\mathbf{v} \subseteq \mathbf{x}_{\sim i}^*$. It is known that the joint probability density function of $\mathbf{x}^*$ can be decomposed as follows:

$$p(\mathbf{x}^*) = p(x_1^*)p(x_2^*|x_1^*)p(x_3^*|x_1^*, x_2^*) \dots p(x_n^*|\mathbf{x}_{\sim n}^*). \quad (11)$$

Based on (11), Rosenblatt transform provides a bijective mapping to the unit hypercube $\mathbf{U} \sim [0, 1]^n$:

$$F_1(x_1^*) = u_1, \quad F_2(x_2^*|x_1^*) = u_2, \quad \dots, \quad F_n(x_n^*|\mathbf{x}_{\sim n}^*) = u_n. \quad (12)$$

There are $n!$ different transforms due to the permutations of $\mathbf{x}$, however, the results of interest narrow into two special cases concerning x_i . In the case of $x_i = x_1^*$, the transform is performed via the marginal distribution function of x_i , hence contains the effect from the constraints as well, while in the case of $x_i = x_n^*$, these effects are excluded due to the fact that F_i is conditional to all other variables. Therefore the authors in [7] call the 1st order Sobol' indices of x_i gained from these special transforms as full and individual indices, respectively. These indices characterize the uncertainty contribution of dependent variables.

3. Surrogate Models

Performing sensitivity analysis usually requires the modelling of the examined configuration in many settings of the input parameters in order to get the proper number of samples to Monte Carlo simulation. At each sample the electromagnetic model is evaluated (e.g., finite-element model, integral equation model), which might lead to a very long simulation. To reduce the computational burden due to the "curse of dimensionality", surrogate models are used as low-cost approximations of the true simulator, usually constructed as a linear combination of a set of orthonormal multivariate basis functions. In general a set of orthonormal univariate basis functions $\Psi(\mathbf{x}) = \{\Psi_1(x), \Psi_2(x), \dots, \Psi_l(x)\}$ is created in the first step, then the multivariate ones are built as their tensor product:

$$\Phi(\mathbf{x}) = \Psi(x_1) \otimes \Psi(x_2) \otimes \dots \otimes \Psi(x_n). \quad (13)$$

This set can be truncated to a set of lower cardinality, yielding the surrogate model as

$$f(\mathbf{x}) \approx \hat{f}(\mathbf{x}) = \sum_{i \in \mathcal{A}} \alpha_i \Phi_i(\mathbf{x}) \quad (14)$$

with $\mathcal{A}$ being an index set. Basically, the differences between the numerous data-fit models are the strategy behind the construction of the univariate basis, the truncation scheme and the estimation of the coefficients. Herein two commonly used methods are briefly summarized: the Polynomial Chaos Expansion (PCE) and the Sparse Grid (SG) interpolation.

Polynomial Chaos Expansion. The PCE provides approximation in a stochastic framework of $f(\mathbf{x})$ by choosing the basis functions to be orthonormal with respect to the joint probability density function of $\mathbf{x}$, e.g., Legendre-polynomials for variables uniformly distributed in $] -1, 1[$, Hermite-polynomials for Gaussian distributions, etc. [4] The truncation might be performed by giving a limit on the highest occurring polynomial degree. The coefficients are traditionally calculated from random input and output samples by the ordinary least-squares method:

$$\hat{\alpha} \approx \arg \min \mathbb{E} \left[\left(f(\mathbf{x}) - \sum_{i \in \mathcal{A}} \alpha_i \Phi_i(\mathbf{x}) \right)^2 \right]. \quad (15)$$

There is strong link between Sobol' decomposition and PCE due to the uniqueness of the former one and the orthogonality of the basis functions. The variance of $f(\mathbf{x})$ is

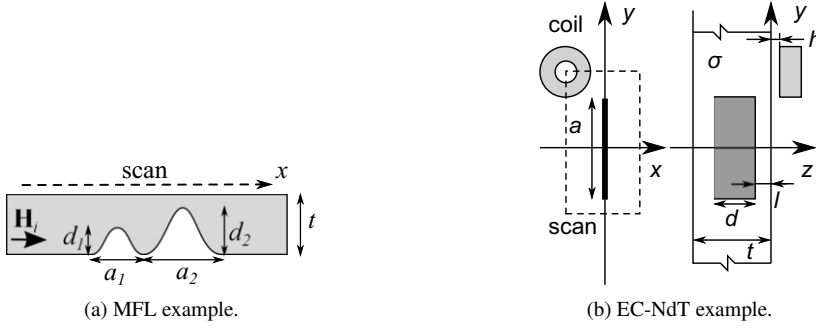


Figure 1. Sketch of the configurations. MFL parameters: $a_{1,2} \in [0.6, 1.2]$ mm and $d_{1,2} \in [0.2, 1.6]$ mm. EC-NdT parameters: $t = (1.25 \pm 0.01)$ mm, $\sigma = (1 \pm 0.01)$ MS/m, $h = (0.5 \pm 0.05)$ mm, $a \in [2, 10]$ mm, $d \in [0.125, 1]$ mm and $l \in [0.125, 1.125]$ mm.

partitioned into the sum of the square of the coefficients, providing a convenient way to evaluate the sub-variances and the Sobol' indices.

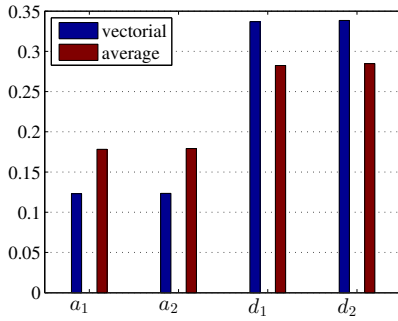
Sparse grid interpolation. SG interpolation (detailed in [1], [8]) is based on the evaluation of the original model at specific points, calling them the supporting nodes. The basis functions have a hierarchical structure of level-by-level, each of them belongs to a supporting node $\mathbf{x}_l^{(i)}$. Their tensor product results in a multivariate basis that can be truncated by, e.g., a linear constraint on the sum of the levels: $\sum_{i=1}^n l_i = l$. In the case of linear ("hat") basis functions, the interpolant at depth d equals to the sum of interpolant from the previous depth and the linear combination of the multivariate functions at level d :

$$\hat{f}(\mathbf{x}) \approx \hat{f}_d(\mathbf{x}) = \hat{f}_{d-1}(\mathbf{x}) + \sum_{i=1}^{m_d} \Phi_d^{(i)}(\mathbf{x}) \underbrace{\left[f(\mathbf{x}_d^{(i)}) - \hat{f}_{d-1}(\mathbf{x}_d^{(i)}) \right]}_{v_d^{(i)}} \quad (16)$$

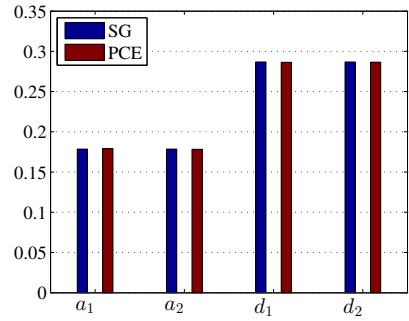
with $\hat{f}_0(\mathbf{x}) = \Phi_0^{(1)}(\mathbf{x})f(\mathbf{x}_0^{(1)})$ and $\mathbf{x}_d^{(i)}$ denoting the vector of 1-D nodes belonging to the 1-D functions of $\Phi_d^{(i)}$. The coefficients $v_d^{(i)}$ are equal to the difference between $f(\mathbf{x})$ and $\hat{f}_{d-1}(\mathbf{x})$ at the nodes $\mathbf{x}_d^{(i)}$. The SG provides a surrogate model to generate Monte Carlo samples, however, it can be used as numerical quadrature to directly evaluate the integrals in (2), similarly to PCE [9].

4. Numerical examples

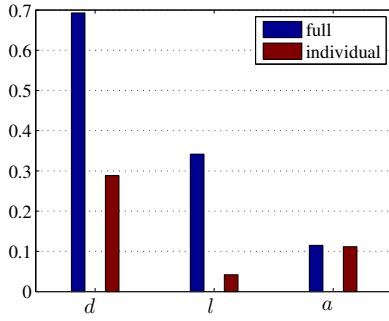
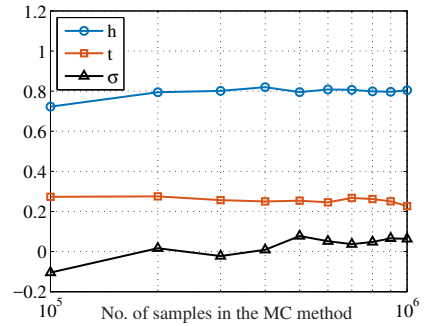
Magnetic Flux Leakage NdT (4-parameter model). A ferromagnetic plate of thickness $t = 2$ mm and $\mu_r = 100$ is corrupted by grooves infinitely perpendicular to the x -axis. The grooves are described by 4 parameter-cubic splines, with parameters uniformly distributed as given in the caption of Fig. 1. A homogeneous $H_{x0} = 1$ A/m magnetic field is imposed in the x direction. The distortion field (ΔH_x) is measured at 51 equidistant points on the top of the surface. The depths of the grooves were found to be more important parameters compared to their widths (Fig. 2a). Good correlation between the vectorial and



(a) 1st order vectorial and average Sobol' indices.



(b) Comparison of the SG and PCE based direct approaches by the average indices.

Figure 2. Results of the MFL example.(a) Full and individual 1st order indices of the defect parameters while h, t, σ are kept at their mean value.(b) Convergence of the setup parameters while d, l, a are kept at their mean value.**Figure 3.** Results of the EC-NdT example.

averaging method can be observed. Direct methods equally resulted in the same outcome as MC-based calculation (Fig. 2b).

Eddy-Current NdT (3+3 parameter model). An infinite, non-ferromagnetic, conductive plate with a thickness t and conductivity σ is investigated as shown in Fig. 1b. The plate includes an ideal crack described by 3 parameters from which d and l are dependent ones due to the $d + l \leq t$ constraint. A coil with time-harmonic excitation of 150 kHz is scanning over the surface at a lift-off h . The change of its impedance is measured at 297 test points of the grid $(x, y) \in \{-2 : 0.5 : 2\} \times \{-8 : 0.5 : 8\}$ mm. A single model evaluation needs 15 seconds by the integral equation simulation [10], thus a SG model was built to reduce computation time. A cross-validation was also performed to ensure its accuracy; RMS-error of 4 % with depth=5, below 0.1 % with depth=7 was achieved. Due to the complex nature of the output vector, it had to be transformed to a 594-element real-valued vector as $\mathbf{F}(\mathbf{x}) \Rightarrow [\text{Re}\{\mathbf{F}(\mathbf{x})\}; \text{Im}\{\mathbf{F}(\mathbf{x})\}]$. The sensitivity analysis was performed by dividing the parameters into two groups: defect parameters (a, d, l) and parameters of the measurement setup (h, t, σ).

The depth of the crack was repeatedly found to be the most important defect parameter as both its individual and full index are the highest ones (Fig. 3a). The lift-off h exceeds out of the setup parameters, while the conductivity has almost no effect. The convergence study confirmed that the number of required samples need to be close to the order of millions (Fig. 3b).

5. Summary

The proposed extension of the Sobol' indices is found to be an appropriate technique to characterize the effect of uncertain parameters on a complete line/surface scan. To some extent, dependent uncertain input parameters can also be treated. The presented tools are shown to apply well to various NdT examples, the conclusions coincide with the physical expectations. Future work will include further reduction of the required sample number by means of PCE coefficients.

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Depth Crack Evaluation Using Multi-Harmonic Excitation and GMR Sensors

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Abstract. This paper reports an experimental study to detect and distinguish surface and sub-surface cracks in non-ferromagnetic materials using eddy current testing with a multi-harmonic excitation signal. A minimization of the crest factor of the excitation signal was done using a Shapiro-Rudin phase sequence. The chosen range of frequencies was based on a corresponding range of standard depths of penetration (SDP) in order to ensure that the eddy currents penetrate more or less an aluminum plate with 4 mm of total thickness. Experimental tests were realized on surface and sub-surface cracks with depths of 0.5, 1, 1.5 and 2 mm. The results obtained by the ECT probe are analyzed in the frequency domain.

Keywords. Eddy current, multi-harmonic signal, Shapiro-Rudin sequence, crack inspection.

1. Introduction

Nowadays non-destructive testing and evaluation (NDT&E) plays an important role to monitor the material condition of industry components. To date, several methods of NDT&E have been developed [1].

In the case of metallic structures, eddy current testing (ECT) is a well-known method used to detect, locate and geometrically characterize the cracks [2-3]. In the last decades, the research activity has been focused on the development of new eddy current testing approaches capable to provide more information about the crack presence in metallic structures [4-5].

Due to the nature of the eddy current phenomena, the eddy current density in a conductor is maximum at the material surface and it is attenuated with increasing depth. For a given material, the standard penetration depth (δ) of the eddy currents changes as a function of the testing frequency (f). A single sinusoidal current excitation with a fixed frequency can be imposed to a coil to induce eddy currents in the metallic surface [6]. The use of a more complex excitation current signal (such as a multi-frequency signal) allows the penetration of eddy currents in the material at different levels providing a richer spectral content [7].

For this purpose, this paper presents an experimental study performed to distinguish surface and sub-surface cracks using a multi-harmonic excitation signal

with equal harmonic amplitudes and phases in a Shapiro-Rudin sequence. The obtained results are analyzed in the frequency domain.

2. Measurement System

The measurement system includes an ECT-based probe with a planar coil and a GMR sensor, a x-y positioning system, a function generator, a power amplifier and a multifunction data acquisition (DAQ) board, as depicted in figure 1. The planar coil was carefully positioned to induce eddy currents perpendicular to the crack line, and generates a magnetic excitation field parallel to the crack. The GMR sensing axis was fixed parallel to the crack measuring the component of the magnetic field on that direction (B_x). This field component contains the strong excitation magnetic field and the field perturbation originated by the presence of the crack. A special care must be taken in order to avoid the field sensor saturation. This component was chosen because it is much smoother than the other two components, with a maximum value over the crack central point. An instrumentation amplifier with $G=27$ dB was used to amplify the GMR output voltage to a suitable range of the DAQ board. The instrumentation amplifier and the GMR sensor were powered by a power supply of ± 12 V. A PXI-1036 Chassis from National Instruments was used to control the positioning system by RS232 interface and the AFG3012 function generator from Tektronix by GPIB interface. The TS200 power amplifier was used to amplify the multi-harmonic signal produced by the function generator to impose an excitation RMS current of 800 mA in the coil. The NI-PXI-6251 DAQ board inserted in the PXI Chassis was used to monitor the excitation current through a sampling resistor (R) connected in series with the coil and to measure the magnetic field from the GMR sensor. A Matlab program running in a desktop personal computer controls the complete system. The multi-frequency signal was created in the Matlab program and stored in the function generator when the scan begins.

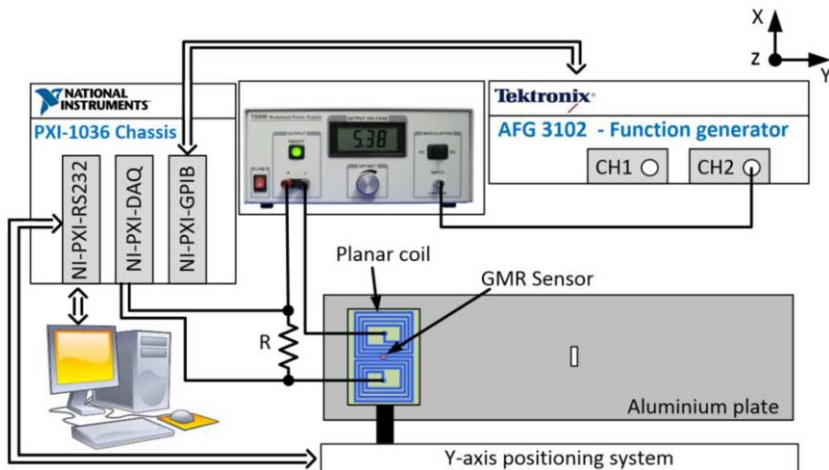


Figure 1. Measurement System.

2.1. Multi-tone signal

A multi-tone voltage signal was imposed to the coil to obtain a wide group of frequencies that allows the eddy currents to penetrate the material into different layers. Sinusoidal tones were combined in the following form:

$$s(t) = \sum_{k=1}^N A_k \cos(2\pi k f t + \alpha_k) \quad (1)$$

where k represents the number of the harmonic tone, A_k is the imposed current amplitude of each sinusoid, f is the fundamental frequency, α_k is the phase of each sinusoid, and N is the number of tones. All tones were generated with equal current amplitudes ($A_{kef} = A_{ef}$) which means that the root mean square value of the current signal $s(t)$ is given by:

$$S_{ef} = \sqrt{\sum_{k=1}^N A_{kef}^2} = \sqrt{N \times A_{ef}^2} = A_{ef} \sqrt{N} \quad (2)$$

Considering a phase value equal to zero ($\alpha_k = 0$) in each sinusoid of the multi-frequency signal, a strong amplitude peak value will be concentrated at the instants $t = kT_0$, T_0 being the period of the fundamental frequency. If this amplitude peak value is too high, problems related to the saturation of the GMR sensor, to the instrumentation amplifier and even to the signal generator will arise. Figure 2(a) shows the distribution of the amplitudes of a multi-harmonic signal (with 128 tones) computed with zero phases (for each sinusoid). In order to reduce the amplitude peak of the generated multi-tone signal, the α_k value of each sinusoid was carefully chosen. The Shapiro-Rudin sequence was used to determine the α_k values that allow the minimization of the crest factor of the excitation signal [8]. To form the Shapiro-Rudin sequence, the number of tones was considered as a power of 2, and the α_k value of each sinusoid equal to zero or equal to π . Figure 2(b) shows the distribution of the amplitudes of a multi-harmonic signal (with 128 tones) computed with the Shapiro-Rudin sequence.

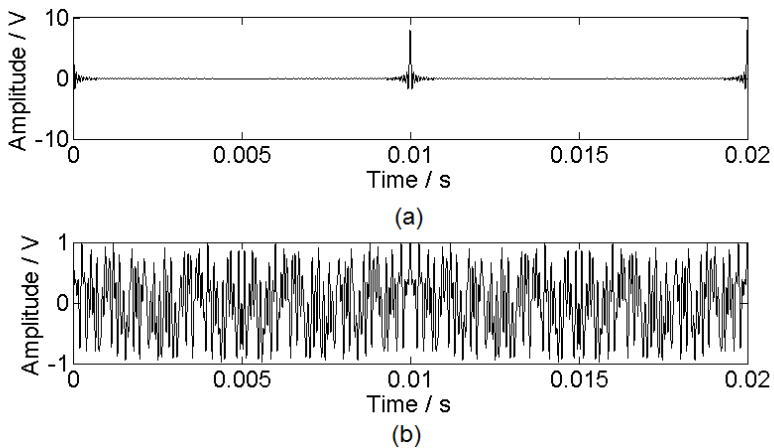


Figure 2. Representation of the amplitudes distribution of a multi-harmonic signal with 128 tones: (a) with zero phase for each tone; (b) with optimized phase choice for each tone using the Shapiro-Rudin sequence.

3. Experimental Scanning Test

The experimental tests were performed in an aluminum 1050 alloy plate widely used in the metallic industry. A multi-harmonic signal with 128 tones (between 100 Hz and 12.8 kHz) was imposed to the coil to allow the eddy currents to penetrate the aluminum plate with 4 mm of thickness in different layers. For the aluminum under test, this range of frequency ensures that the eddy current penetrates at deeper levels with a standard depth of penetration equal to 0.7 mm for $f=12.8$ kHz and equal to 8 mm for $f=100$ Hz. The first experimental tests were done to evaluate a surface crack and a sub-surface crack with 8 mm of length and 2 mm of depth. This means that the first tests were performed for cracks with 50% of the thickness of the sample plate under test. Figure 3 depicts the scan tests that were performed. The y-axis positioning system moved the ECT probe with steps of 1 mm in a straight line of 30 mm above the center region of each crack.

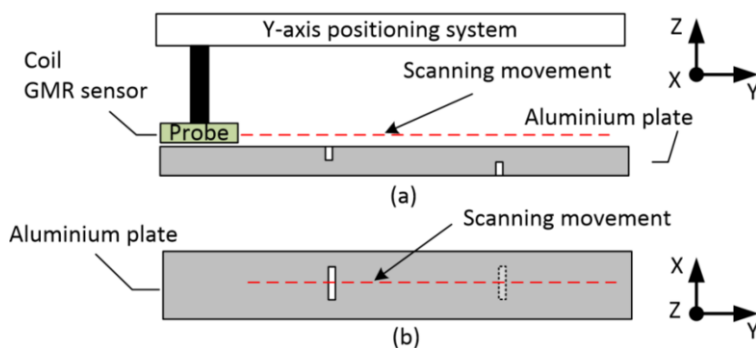


Figure 3. Representation of the performed scan: (a) Side view; (b) Top view.

The magnitude of the multi-frequency current signal was monitored by spectral analysis of the voltage at the sampling resistor R connected in series with the planar coil. A constant spectral current amplitude of 70.7 mA was imposed to the coil, resulting in a total amplitude current of 800 mA. Figure 4 depicts the magnitude of the measured magnetic field component B_x obtained by the GMR sensor for the case with the probe closer to the aluminum plate in a region without crack.

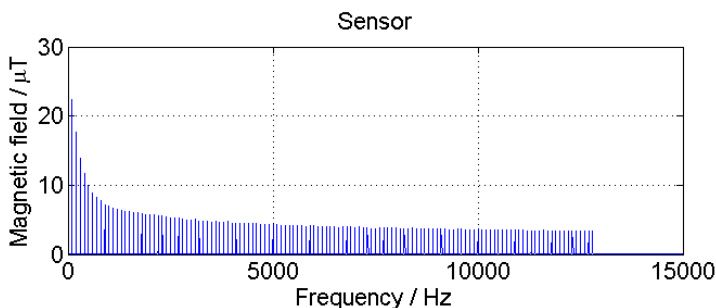


Figure 4. Magnitude Spectrum of the GMR Sensor response when the probe is closer to the aluminum plate in a region without crack.

From figure 4, it is possible to say that the magnetic field response decreases when the frequency increases. As expected, this is due to the induction of a higher eddy current density region at the surface of the plate when the testing frequency increases, which causes a decrease in the resulting B_x component.

Figure 5, depicts the magnitude of the perturbed magnetic field component B_x obtained in a series of points above the two cracks region and along a straight line crossing the crack perpendicularly. Figure 5(a) depicts the spectral content obtained for the surface crack with 8 mm of length and 2 mm of depth, while figure 5(b) depicts the spectral content obtained for the sub-surface crack with the same dimensions, but located on the other plate side. Note that the graphs were plotted after removing the baseline obtained in figure 4 in order to only obtain the perturbed magnetic field originated by the presence of the cracks.

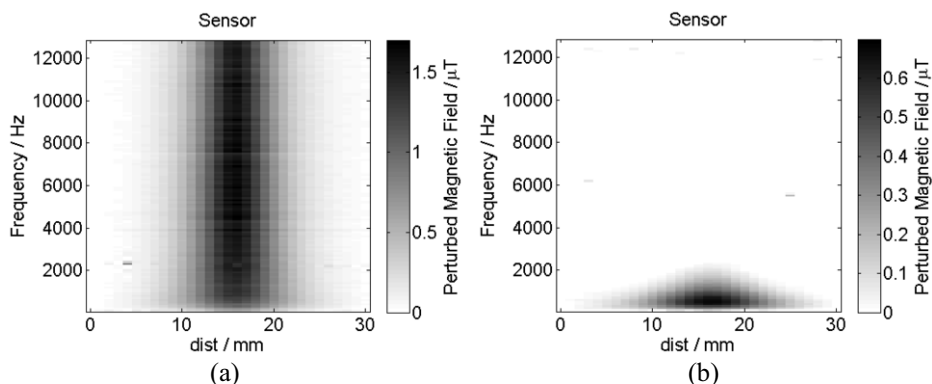


Figure 5. Perturbed magnetic field obtained from the GMR Sensor in a straight line of 30 mm above the center region of each crack: (a) Surface crack; (b) Sub-surface crack.

From figure 5, it is possible to observe that both perturbed magnetic field maps contain information about the dimensions of the cracks and both perturbed fields allow the classification of the cracks as superficial or sub-superficial. From figure 5(a), it can be seen that the value of the perturbed magnitude of B_x is approximately constant in the crack region for the group of 128 tones. This means that the perturbed magnetic field occurs for all the frequencies. From figure 5(b), it is possible to observe that the value of the perturbed magnitude of B_x is dominant in the lower frequency of the spectrum. The peak values of the perturbed magnetic field maps provide the y-axis position of the two cracks. In the y-position of the crack, the spectral magnitude of the perturbed magnetic field contains important information about the dimensions of the cracks. The high number of tones (128) was chosen to obtain a large magnitude spectrum with the information retrieved from the perturbed eddy currents at different layers. However, from figure 5, it is possible to observe that a range of lower number of tones could be used to distinguish the tested surface and sub-surface cracks. This means that the total current value of the excitation signals could be reduced using a lower number of tones. For this purpose, a new multi-harmonic signal with 32 tones was built and experimentally tested for the inspection of surface and sub-surface cracks with depths (d) of 0.5, 1, 1.5 and 2 mm. The experiments were performed in the same aluminum plate with a thickness of 4 mm.

Figure 6(a) shows the spectral magnitude of the perturbed magnetic field in the y-position of each surface crack. Figure 6(b) shows the spectral magnitude of the perturbed magnetic field in the y-position of each sub-surface crack.

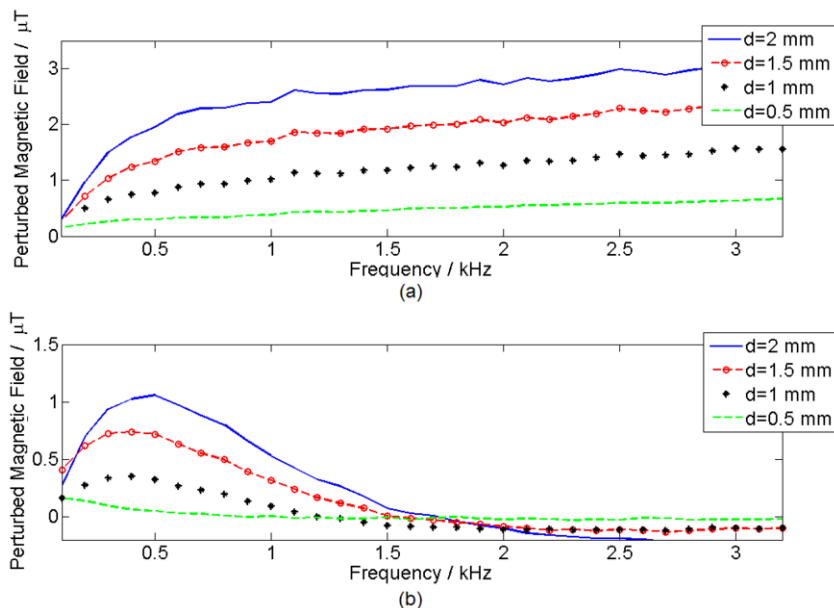


Figure 6. Magnitude of the perturbed magnetic field (B_x): (a) Surface Cracks with different depths ($d=2$ mm, $d=1.5$ mm, $d=1$ mm, $d=0.5$ mm); (b) Sub-surface cracks with different depths ($d=2$ mm, $d=1.5$ mm, $d=1$ mm, $d=0.5$ mm);

From the results, it is evident that the depth of a crack is correlated with the curve of the spectral magnitude of B_x in the y-position of the crack. For both cases (surface and sub-surface cracks), the peak values of the spectral magnitude of B_x increase with the increasing depth of a crack. For sub-surface cracks, the values of B_x are dominant in the lower frequency of the spectrum. For surface cracks, the perturbed magnetic field of B_x occurs for all frequencies of the spectrum, and a small decrease of the perturbed part occurs in the lower frequencies because the eddy currents are forced to flow at deeper levels, originating a lesser perturbation of the magnetic field due to the presence of a surface crack.

4. Conclusions

The constructed multi-frequency signal allowed the surface and the sub-surface detection of cracks with 8 mm in length and 0.5, 1, 1.5 and 2 mm of depth in an aluminum plate with 4 mm thickness. For this kind of multi-harmonic excitation, it is important to reduce the amplitude peak of the generated signal in order to decrease the possibility to saturate the output of the GMR sensor and the output of the instrumentation amplifier. The Shapiro-Rudin approach was well suited to limit the maximum instantaneous amplitude of the excitation signal. In the frequency domain,

the results obtained by the GMR sensor clearly show the difference in the perturbed magnetic field B_x between a surface and a sub-surface crack of equal dimension. The number of tones and the chosen frequencies of test should be optimized for the thickness of the metallic plate under test. The final results also show that the crack depth is directly related with the peak value of B_x in the central position of the cracks.

Acknowledgment

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Advances in Modelling the ECT of U-Bend Steam Generator Tubes Based on the Boundary Element Method

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Abstract. The modelling of the eddy current testing of U-bend steam generator tubes is a valuable tool for interpreting measured signals due to geometric distortions that can arise during the bending process. Especially, the decentred trajectory of the probe into the bending must be taken into account to ensure the accuracy of the simulated signal. As the exact trajectory is a priori not known, we propose an efficient way to build a database of the simulated signal for all possible positions of the probe into the tube, based on an adaptation of the boundary element method that is developed at CEA LIST.

Keywords. Eddy Current Testing, Modelling, Boundary Element Method

1. Motivation

We are interested in modelling the eddy current testing (ECT) of U-bend tubes in steam generators that are part of the second barrier containment of radioactive materials in nuclear pressurized water reactors. In addition to the curvature of the tube, whose modelling has already been successfully achieved for example in [1] and recently used in [2] in complex configurations, the bending process may cause a variation in thickness as well as the distortion of the inner wall of the tube [3-5]. It also affects the movement of the probe that is shifted and may be tilted in the bending. The detection and characterization of defects in this zone requires a signal analysis due to these phenomena and thus their modelling. Numerical simulation is challenging as the method used must be sufficiently flexible to adjust to small geometric distortions and accurate enough so as the digital noise does not disturb their low signal variations. As part of a collaboration between CEA LIST and IRSN that is, the French public expert in nuclear and radiological risks, we use an integral formulation [6] handled by the boundary element method (BEM) [7-9] to simulate the effects of the bending for ECT.

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A synthetic description of the BEM and of its specific use for this study are given in sections 2.1 and 2.2, respectively. Recently, these numerical tools made it possible to observe the sensitivity of the probe response to its trajectory in the bending [10].

A numerical assistance strategy for determining this trajectory is therefore proposed in section 3.1. It is based upon the use of a database of coil impedance variation for all mechanically admissible positions of the probe in the bending. The main originality of this contribution lies in this modelling strategy which is made viable thanks to the previously introduced BEM. An X-ray imaging of the probe in the bending revealed the distortion of the inspected tube as well as the effective position of the probe and validates, yet partially, the proposed strategy in section 3.2.

Once the probe location is established, the ECT response of a defect can be simulated, allowing to observe the effect of the bending onto its signature. Section 4 shows that this perturbation remains secondary by comparison with the trajectory of the bobbin coil.

2. Advanced Boundary Element Method

2.1. Boundary Element Method

The 3D problem can be expressed as an equivalent transmission problem at the interface between homogeneous parts of the medium by a suitable integral representation form [6]. The goal then becomes to compute the electric ($\mathbf{J} = \mathbf{n} \times \mathbf{H}$) and magnetic ($\mathbf{M} = \mathbf{E} \times \mathbf{n}$) surface current densities ($\mathbf{n}$ stands for the normal at the interface, $\mathbf{E}$ and $\mathbf{H}$ are the electric and magnetic fields). A Galerkin discretization of the variational form then enables us to approximate the physical solution by solving a linear system.

Compared to the finite element method, the number of unknowns within the boundary element method is dramatically reduced (but the matrix is full with complex filling) and therefore the use of a direct solver becomes possible. A direct solver is far more effective than an iterative one when we face a system that is ill-conditioned or presenting many RHS (right hand sides), provided that the number of unknowns remains limited (a few tens of thousands). Hence it is generally well suited to ECT applications for which we can count thousands of RHS (positions of the coil) in a restricted computational domain, due to the localized support of the source and to negligible propagation effects.

A higher-order approximation [7] (i.e. increasing the polynomial degree of the expansion) further reduces the number of unknowns for a given precision, therefore making more accessible the direct solver. In practical terms, the increase of the order of the mesh allows the use of a small number of surface elements with curved edges that are particularly effective for modelling smooth geometric distortions.

2.2. Adapted BEM

Such a high order boundary element method is currently implemented at CEA LIST. Firstly developed for magneto-static applications such as the modelling of 3D ferrite cores [11], this approach has been expanded to low frequency electromagnetic applications [8] and especially to the quasi-static eddy current regime [9]. In what follows as in [12] we use Hdiv-conforming boundary elements similar to [7] with a quasi-

Helmholtz decomposition (see e.g. [13]) of the integral boundary formulation presented in [8] without the multi-step option.

The quality of the discrete mesh is crucial as low changes of the curvature may lead to high numerical noise with regards to the slight variations of signal we are looking for. As a first step, meshing is performed on a portion of the straight tube for nominal parameters (external diameter of 19.05 mm, thickness of 1.09 mm and conductivity of 1 MS/m). This mesh is composed of quadrilateral elements with curved edges, whose size increases as the element is closer from the truncation. Then, for a given axial location of the probe in the bending, corresponding distortions are applied to the straight mesh to obtain the deformed one (Figure 1). Finally the probe is shifted and tilted as wanted.

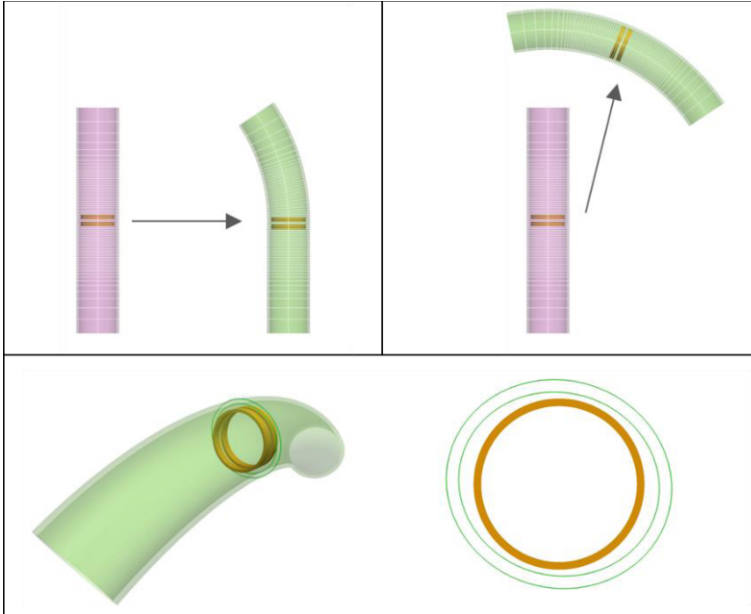


Figure 1. Mesh of the straight (purple) and of the deformed (green) SG tubes for two locations of the probe (orange), considering an ovalization and a torsion (bottom).

The small number of high-order basis functions (here 2×7424 1st and 2nd mixed-order basis functions) allows us to use a LU-based direct solver, so that many tilt and shift variations can be treated in a single calculation at a fixed axial position of the probe. This has facilitated the achievement of parametric studies to identify the most influent inputs, namely the distortion (such as ovalization) of the inner tube wall and the probe path for a bobbin coil [10]. As a matter of fact, these parameters vary depending on the tubes and the probes and are not explicitly known.

3. Proposed Strategy

3.1. Simulation tools to help determine the path of the probe

We therefore focused on a simulation strategy that can assist NDT practitioners in describing the probe position. This strategy first consists in generating (by forward

simulation) a database of the coil impedance variation of all the mechanically admissible positions of the probe in the tube.

The plastic body of the probe is first introduced into the geometric model. Shown in Figure 2.1, this body is represented by a cylinder whose ends are two half ellipses. The envelope of the mechanically possible positions is then determined by a landing algorithm of this probe body onto the inner wall of the mesh. Illustrated in Figure 2.2 for a fixed axial position, this (2D) section of the (3D) envelope is read from bottom to top for the shift and from left to right for the tilt. The signal is then calculated on a representative basis. This is illustrated Figure 2.3 on a complete tube for which we consider here a gradual ovalization from the straight part to the top of the bending (the tip), with a sensor operated in absolute mode. These impedance variations are not calibrated and are given for information purposes only.

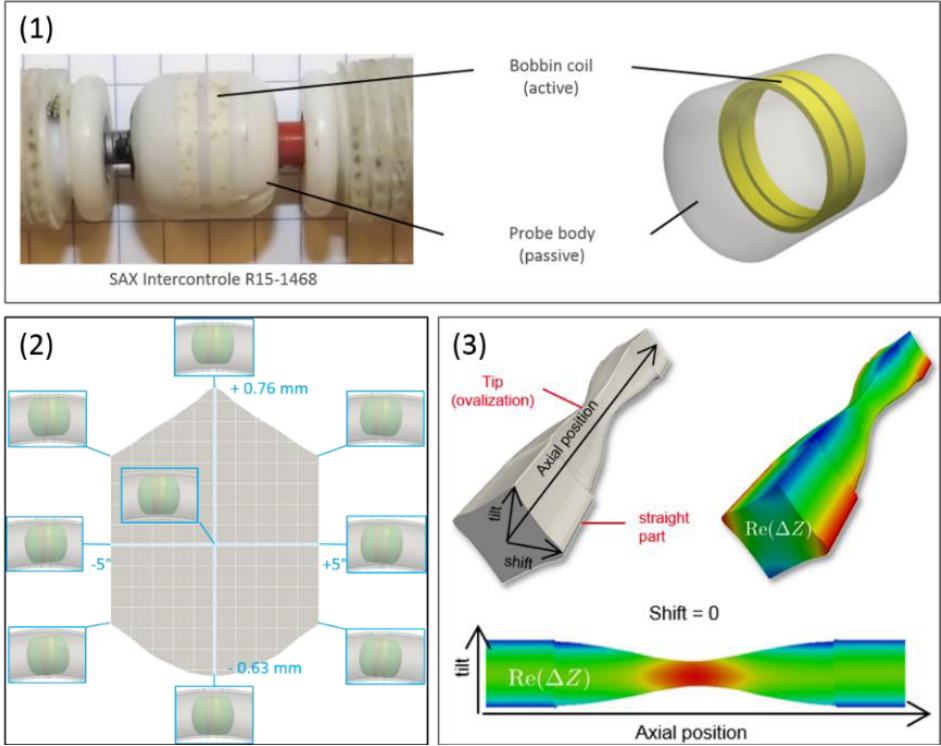


Figure 2. Geometric modelling of the plastic body of the probe (1), 2D envelope of the admissible positions of the probe at a given axial position in the bending (2), 3D envelope for the complete U-bend tube and corresponding impedance variation of the signal (3).

Realized as a pre-processing step (offline), building such a databases requires 1 to 2 hours of calculation on a standard PC. This computation time is directly related to the deformation profile of the tube: fast deformations with regards to the axial displacement of the probe would necessitate increasing the number of forward unit simulations (i.e. fixed axial positions) and would therefore generate an additional cost of the order of the hour.

Once this database is available, the second step would be to simulate (on line) the impedance variation signal for a large number of parametric trajectories of the probe and to rely on acquisitions to refine this parametrization. However such a parametrization

should be based on few trajectory observations that are hardly available. An alternative could be to estimate the trajectory using Kalman filter.

3.2. Adjustment of geometric parameters through X-ray inspection for validation

X-ray imaging of probe was performed at the top of the bending to set the geometric inputs of the model (Figure 3.1 and Figure 4.1). Figure 4.1, inside the tube, the red lines delimit the boundary of the coils (external diameter of 14.36 mm, thickness of 0.67 mm) while the green line corresponds to the boundary of the plastic body of the probe (diameter of 15.00 mm). Onto the tube, red and blue lines correspond to outer and inner walls of the tube while dashed black lines correspond to a perfect cylinder (the tube with no distortion).

These inputs were taken into account in the model (Figure 4.2) and the signal was simulated for each inspection channel on the envelope of the mechanically possible positions of the probe at this axial position (Figure 4.3). Each result has then be restricted to the zone where the simulated signal is close to the acquisition (Figure 3.2). The intersection of the resulting reduced envelopes indicates the supposed position of the probe, here in contact with the extrados (Figure 4.4).

Incorrect results (shifted probe towards the intrados) were obtained before imaging, when simulating the signal on a U-bend tube with no distortion [10]. This sensitivity would justify extending the database to the main geometrical distortion inputs.

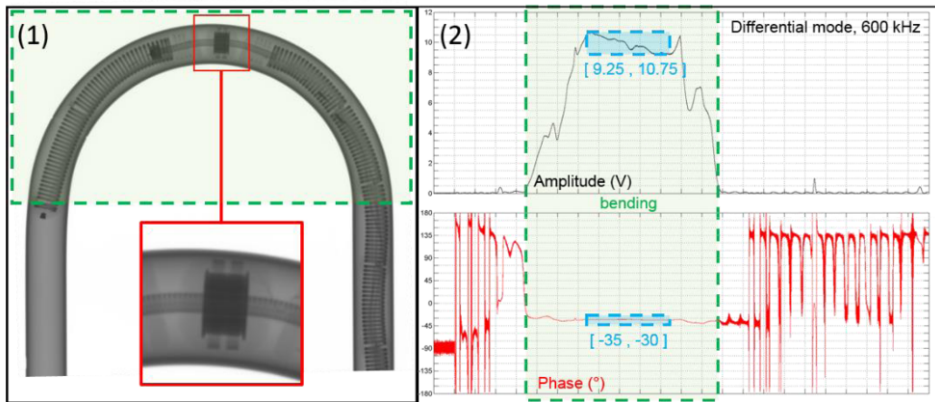


Figure 3. X-ray imaging of the probe at the tip (1) and experimental measurement for a probe operated in absolute mode at 600 kHz (2).

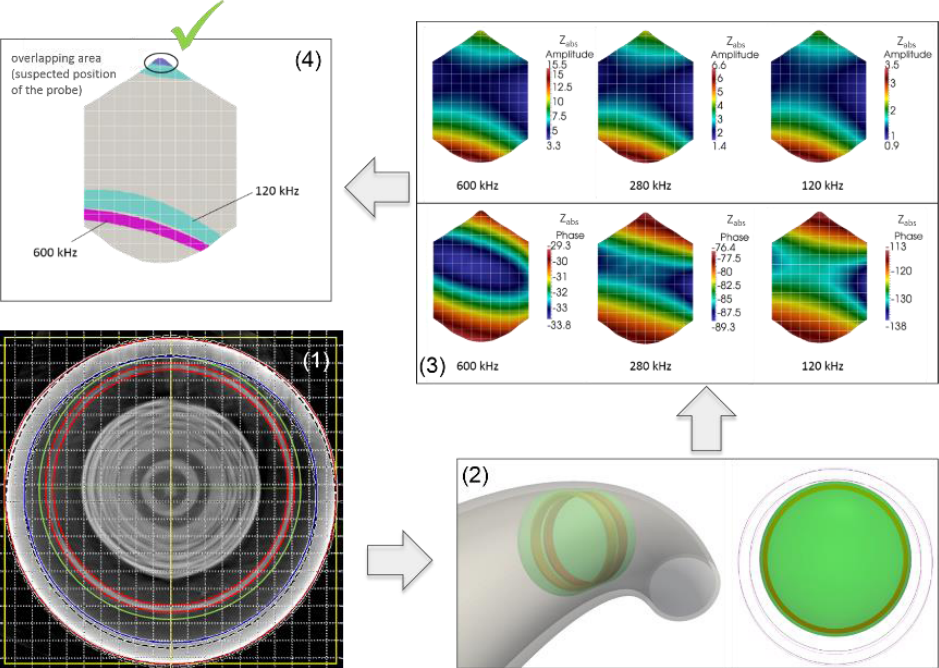


Figure 4. X-ray imaging of the probe at the top of the bending (1), corresponding geometric modelling (2), phase and amplitude of the simulated signal for the admissible positions of the probe (here operated in absolute mode) (3), and intersection of the reduced envelopes (4).

4. Simulation of the defect’s signature

Standard defects such as grooves & notches were introduced in the straight tube, before distortion. Hence their effective depth is not absolute but relative to the local thickness of the section. This process and the few simulated defects are illustrated [Figure 5](#).

Simulations were carried out for a straight tube then for a 3-inch bending, with a shifted probe at 200 μm or in contact with the extrados (lift-off = 0 mm), as well as for a centred probe in the straight tube. The defect signals were balanced (but not calibrated) with the defect-free signal for a corresponding probe position so as to focus on the possible correlation between the flaw response and the geometry signal.

It is found that the simulated signals vary only slightly with the bending, with at most a phase shift of a few degrees compared with the signals obtained in the straight part of the tube for the same probe shift, as shown in [Figure 6](#). Taking into account the tube wall distortion revealed by X-ray imaging did not alter this observation on the few defects studied.

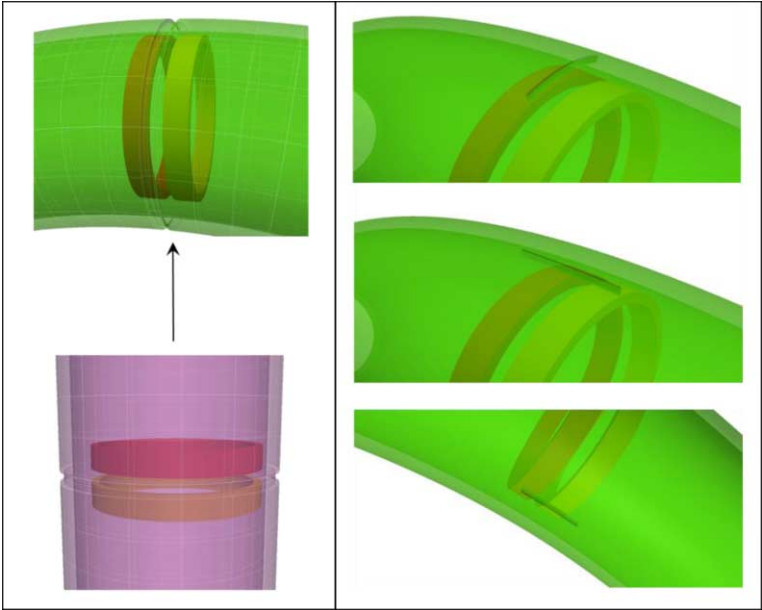


Figure 5. Geometric modelling of simple defects.

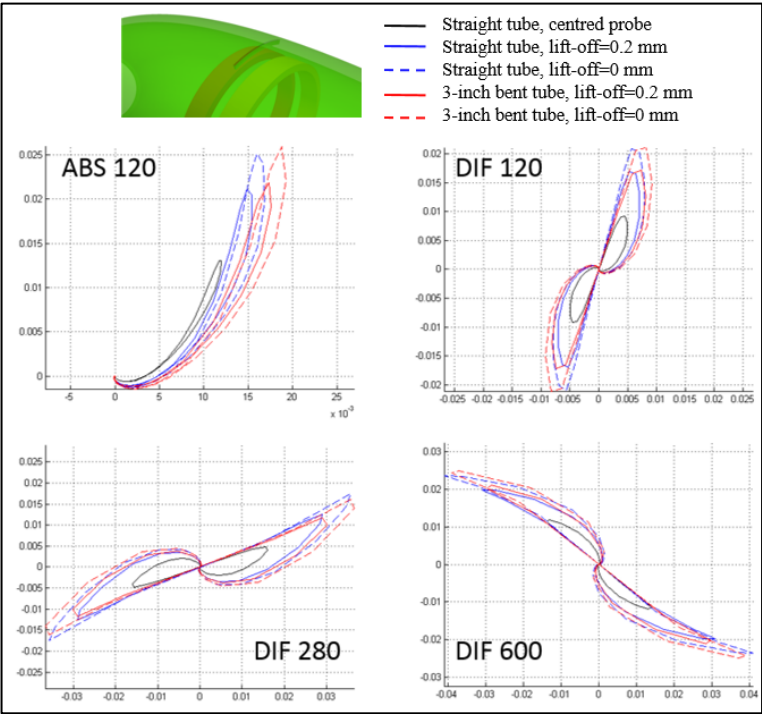


Figure 6. Simulation of the ECT signal with a circumferential notch (40% depth, 6 mm x 0.15 mm) and two bobbin coils operated in absolute mode (ABS) at 120 kHz and differential mode (DIF) at 120, 280, 600 kHz.

5. Conclusion

Modelling the eddy current testing in the bending of a steam generator tube requires good knowledge of the probe's trajectory. Since this trajectory is not known explicitly, we proposed simulation tools to assist its parametrization.

Based on the use of a database for which direct calculations are carried out using a higher-order boundary element method, this process has been partially validated by experimental measurements carried out in the framework of this study. It is found that the process is particularly sensitive to the tube wall deformation, suggesting that the database should be extended to the principal distortion inputs.

On the other hand, first simulations tend to show that the response of a defect is very sensitive to the trajectory of the bobbin coils but remains relatively insensitive to the deformation of the tube. This should motivate the search for a post-treatment of (fast) simulations made on a simplified geometry. More precautions should be taken with pancake coils, and additional geometric effects such as pilger noise should be investigated to complete and confirm these observations.

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Fast Models Dedicated to Simulation of Eddy Current Thermography

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Abstract. This communication presents the first development aiming at efficiently simulating configurations of eddy current thermography for nondestructive evaluation. The numerical method proposed here is based on the Finite Integration Technique for both electromagnetic and thermal problems. Simulation results obtained using two different materials, steel and aluminum, are compared and discussed with respect to the presence of a flaw affecting the piece under test.

Keywords. Eddy current, thermography, simulation, finite integration technique

1. Introduction

The use of thermography [1] for nondestructive testing applications had received growing attention in the last years. This is mainly due to the fact that infrared (IR) cameras have recently improved significantly in both sensitivity and spatial resolution and that this technique is particularly adapted to many applications [2] such as composites' inspection. Unlike other direct imaging techniques, it is a fast, high resolution and contactless method. Thermal testing is generally divided into two main streams: passive and active. Passive thermography is defined as measuring the temperature difference between the target material and its surroundings under different ambient temperature conditions. Active thermography uses a thermal source in order to deposit heat in the target material. Most common sources consist in lamps or lasers [3] that heat part of the piece surface. These techniques of depositing heat on the materials have potential disadvantages, e.g. the reflected heat from the material can interfere with the measured signals, causing signal-to-noise-ratio (SNR) problems. For instance, many conductive materials when used in industry are coated or painted. The heating of the workpiece may also be obtained via the application of sonic or ultrasonic energy using a welding horn, i.e. vibrothermography, thermosonics or sonic infrared [4]. In this case, however, contact between the workpiece and the ultrasonic welding horn is required, which can complicate its practical use and cause a loss of energy transmission.

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Eddy current thermography (ECT), also named as induction thermography, is an alternative to inspect metallic structures that does not suffer from the above-mentioned disadvantages. This is an emerging technology in nondestructive testing (NdT) that combines eddy current and thermography. ECT is based on electromagnetic induction and Joule effect heating. The technique uses induced eddy current to heat the sample and defect detection is based on the changes of the induced eddy current flow revealed by the thermal visualization captured by an IR camera. Induction thermography can be used to detect cracks [5], disbond, impact damage, delamination and corrosion.

This work presents a modelling approach using a two-dimensional numerical solver based on the Finite Integration Technique (FIT) [6,7]. A typical configuration, consisting of a coil located above a plate, is sketched in Figure 1. This configuration will be used in our simulations and an axial symmetry is assumed. The same numerical tool is used to solve both physical problems, namely the electromagnetic induction by the coil in the plate and the heat diffusion in the plate after excitation. Due to the large difference in time scale between the electromagnetic problem and the thermal one, a weak coupling of the two problems is possible.

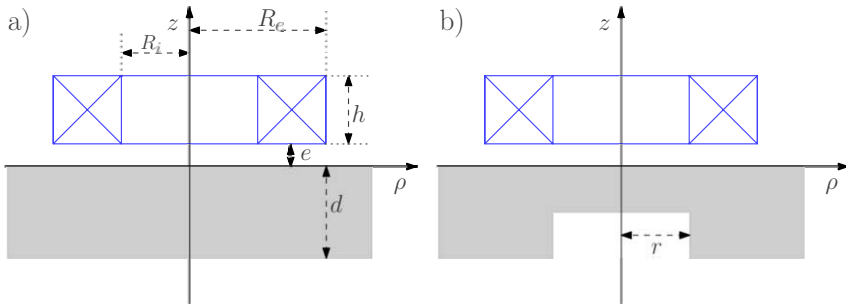


Figure 1. Schematic setup diagram. Circular coil of inner radius R_i , outer radius R_e and height h standing above a conductive plate of thickness d in a distance e . **a)** Homogeneous plate. **b)** Homogeneous plate with a axisymmetric defect of radius r .

In other words, the electromagnetic problem is first solved to calculate the time-dependent eddy current density induced in the plate, then it is converted into a heat source term by considering Joule effect. Finally, the diffusion of heat in the plate is computed with respect to time. This first development will serve as reference for further works, consisting in solving both problems with fast modal methods [8].

2. Theoretical formulation

ECT involves multi-physical interactions with electromagnetic-thermal phenomena including eddy current, Joule heating and heat conduction. Simulation of induction heating requires the ability to model multiple physical fields. Thus, modeling the generation of eddy current requires an electromagnetic solution in the workpiece, which results in a Joule heat distribution. The latter is used as a volumetric heating source in order to obtain the temperature distribution in the workpiece.

The coupling between the electromagnetic problem and the thermal one can often be further complicated by the fact that the electromagnetic properties of the workpiece are

depending on the temperature of the workpiece, which will lead to a strongly-coupled problem. This coupling of the two problems requires the electromagnetic solution to be computed based on a time/temperature updated set of materials properties. This leads to a time consuming numerical computation that can be avoided under some assumptions.

The induction heating process involves multiple time and length scales. Generally, the time scale associated with the heat transfer is much larger than the time scale associated with the electromagnetics. The time scale associated with the electromagnetic solution depends on the frequency f of the alternative current in the coil, while the time scale associated with the transient heat transfer in the workpiece is determined by its thermal properties. The length scale, for the electromagnetic problem, also depends on the frequency, as well on the magnetic permeability μ and electrical conductivity σ . The so-called *skin depth*, defined in the particular case of a half-space medium by the relation $\delta = (\pi\mu\sigma f)^{-1/2}$, illustrates the penetration of the electromagnetic field in the piece. As a consequence, the associated Joule effect is also generated in a depth range of two or three times the skin depth δ . From the numerical point of view, this implies that this particular region must be finely discretized.

2.1. The electromagnetic problem

In a typical configuration, a pulse generator emits a signal to an infrared camera and to a induction heater, which generates an excitation signal. This excitation signal is usually a sinusoidal of alternating current with high amplitude. The current is then driven into an inductive coil, which induces eddy current in the neighbouring workpiece. This phenomenon is described by the Maxwell's equations, which for the quasi-static approximation are

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t}, \quad (1)$$

$$\nabla \times \mathbf{H} = \mathbf{J}, \quad (2)$$

$$\nabla \cdot \mathbf{B} = 0, \quad (3)$$

where $\mathbf{E}$ is the electric field intensity, $\mathbf{H}$ is the magnetic field intensity, $\mathbf{B}$ is the magnetic flux density, and $\mathbf{J}$ is the current density. Excitation frequencies are typically lower than 10 MHz, consequently the displacement current term ($\partial \mathbf{D} / \partial t$) in (2) can be neglected. The above equations are combined with the following constitutive relations characterizing a linear, homogeneous and isotropic material

$$\mathbf{J} = \sigma \mathbf{E}, \quad \mathbf{B} = \mu \mathbf{H}. \quad (4)$$

For solving the above differential equations and since $\mathbf{B}$ is divergence free, it can be expressed as $\mathbf{B} = \nabla \times \mathbf{A}$ where $\mathbf{A}$ is the magnetic vector potential. Substituting $\mathbf{B}$ into (1) and using (2) as well as the constitutive relations (4), the diffusion equation for the magnetic vector potential can be derived as:

$$\nabla \times \mu^{-1} \nabla \times \mathbf{A} + \sigma \frac{\partial \mathbf{A}}{\partial t} = \mathbf{J}_s \quad (5)$$

where $\mathbf{J}_s$ is the current density driving the inductor. The choice of a Coulomb gauge, $\nabla \cdot \mathbf{A} = 0$, is made here.

The FIT method provides a discrete reformulation of Maxwell's equations on its integral form. A typical simulation task is described by a known geometry and material configuration, as well as boundary and initial conditions. In the following a rectangular cubic cell complex consisting of a material grid complex $\mathbf{M}$, a primary grid complex $\mathbf{G}$ and a dual grid complex $\tilde{\mathbf{G}}$ will be used.

The eddy current density in FIT is computed by the equation

$$\tilde{\mathbf{C}}\mathbf{M}_\mu^{-1}\mathbf{C}\hat{\mathbf{a}} - \mathbf{M}_\sigma\hat{\mathbf{a}} = \hat{\mathbf{j}} \quad (6)$$

where $\tilde{\mathbf{C}}$ and $\mathbf{C}$, contain only topological information and represent a discrete curl-operator on the primary and the dual grid $\mathbf{G}$ and $\tilde{\mathbf{G}}$, respectively, $\mathbf{M}_\mu^{-1}$, $\mathbf{M}_\sigma$, are the material matrices, $\hat{\mathbf{a}}$ is the magnetic vector potential and $\hat{\mathbf{j}}$ is the current density.

Initial conditions of the model assume a thermal equilibrium of the sample and its surroundings. In the z -direction Neumann boundary conditions are imposed. On the artificial left side boundary, at $\rho = 0$ where the axis of the symmetry is, Neumann condition is imposed too. On the right side boundary, at $\rho = \rho_e$ Dirichlet boundary condition is imposed. We suppose that the workpiece is infinite in the ρ -direction so this artificial boundary does not affect the solution within the domain of interest.

2.2. The thermal problem

Due to resistive heating from the induced eddy current, the temperature of conductive materials increases, which is known as Joule heating. It can be expressed by the equation

$$\dot{Q} = \frac{1}{\sigma} |\mathbf{J}_s|^2,$$

where the sum of generated power density $\dot{Q}$ is proportional to the square of the eddy current density. The resistive heat will diffuse as a time transient until an equilibrium state is restored between the bulk and its surface, or better saying the workpiece and the environment. The thermal part of the problem can be divided into two phases, (i) the heating phase, during which the heat is being deposited in the workpiece and (ii) the cooling phase, when the workpiece has reached a maximum temperature, the deposit of heat has stopped, and only diffusion of the heat is occurring in the plate.

Starting with the energy conservation law in integral form

$$\int_V \varrho C_p \frac{\partial T}{\partial t} dV = \int_V \dot{Q} dV - \oint_{\partial V} \mathbf{J} \cdot d\mathbf{s} \quad (7)$$

and using the Fourier's law $\mathbf{J} = -\kappa \nabla T$, the heat equation is derived as

$$-\kappa \nabla^2 T + \varrho C_p \frac{\partial T}{\partial t} = \dot{Q}, \quad (8)$$

where κ is the thermal conductivity, ϱ the density, C_p the specific heat and T the temperature.

In the FIT, the solution of the thermal problem is given by the equation

$$\tilde{\mathbf{S}}\mathbf{M}_\kappa\mathbf{G}\theta - \mathbf{M}_c\dot{\theta} = -\dot{\mathbf{q}}, \quad (9)$$

which is a discrete formulation for the heat equation, where $\tilde{\mathbf{S}}$ is the div-operator on the dual grid, $\mathbf{M}_\kappa$, $\mathbf{M}_c$ are material matrices, $\mathbf{G}$ is the discrete gradient matrix, θ the temperature and $\dot{\mathbf{q}}$ the source term.

Pairing this equation with the boundary conditions, the temperature of the workpiece and its surrounding is computed. The boundary conditions that are used here are the same as in the electromagnetic problem.

3. Simulation and results

For the simulations a circular coil is used with $e = 1$ mm of lift-off and inner radius $R_i = 11$ mm, outer radius $R_e = 84$ mm, height $h = 41$ mm and number of wire-turns $N = 408$ as in [9]. For modeling the workpiece, two different materials are used: (i) Aluminum, (ii) Steel, their respective physical parameters being given in Table 1. We are investigating here two scenarios. In the first case, we compare the behaviours of two homogeneous plates of thickness $d = 10$ mm, made of aluminum and steel, respectively. In the second case, we introduce a defect, which can be assumed to be a corrosion, at the bottom surface of a thin plate ($d = 1$ mm) of aluminum. The defect is modelled as a local change of physical properties (same as the surrounding environment). The excitation signal is considered to be a sinusoid of frequency f . Setting the frequency of the excitation signal at $f = 200$ Hz and the duration of the signal at 50 ms we probe the two different plates with 10 periods of the signal. In Figure 2, the images of the simulation for three crucial times are given. In the first two rows, for which no diffusion has been occurring, the skin depth effect is highlighted, i.e. the difference of the electromagnetic properties of the materials.

As one can expect, through the Joule effect, the penetration depth of the eddy currents in the aluminum plate is much larger than in the workpiece of steel. The difference between both rows of images is a result of the difference of the thermal properties of the materials. A corrosion has been now introduced in a thick aluminum plate. The frequency of the probed signal has been kept the same, at 200 Hz, which gives a penetration depth of 6 mm, much larger than the thickness of the plate. The defect has been modeled as a circular discontinuity in the plane (ρ, θ) of radius $r = 3$ cm or 5 cm and a thickness of 0.5 mm in the z -direction. The results of the simulation of this setup are given in Figure 3.

Since only the infrared radiation emitted by the surface of the workpiece as a function of time can be captured by a thermal camera, in Figure 4 and Figure 5 the surface temperatures are plotted. In Figure 4, the temperature of the the workpiece on the sur-

Table 1. Electromagnetic and thermal parameters of the materials.

	μ_r	σ (S/m)	κ (W/m/K)	ϱ (Kg/m ³)	C_p (J/Kg/K)
Aluminum case	1	3.5×10^7	237	2707	897
Steel	700	3.21×10^6	44.5	7850	475

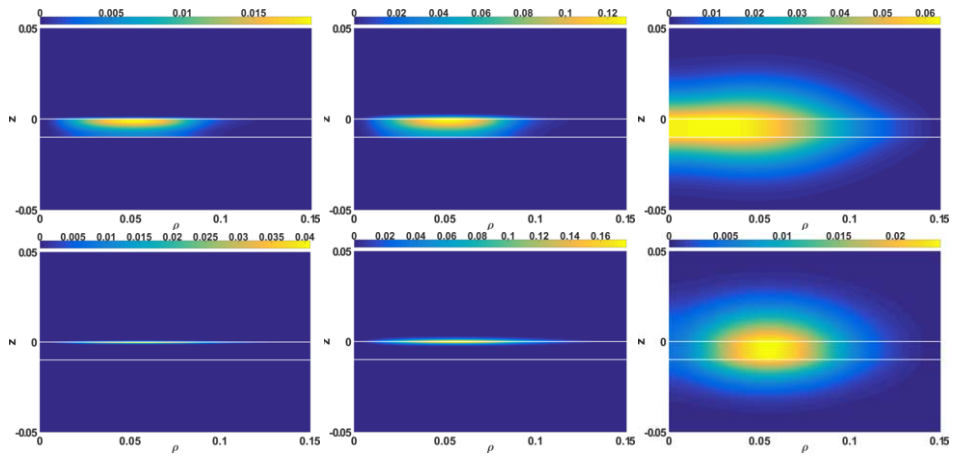


Figure 2. Temperature distribution images in the region of interest for different time and homogeneous materials without defects.

Top: Aluminum. *Left to the right:* Different observation times, $t = t_0$, $t = t_0 + 50$ ms, $t = t_0 + 5$ s.

Bottom: Steel. *Left to the right:* Different observation times, $t = t_0$, $t = t_0 + 50$ ms, $t = t_0 + 5$ s.

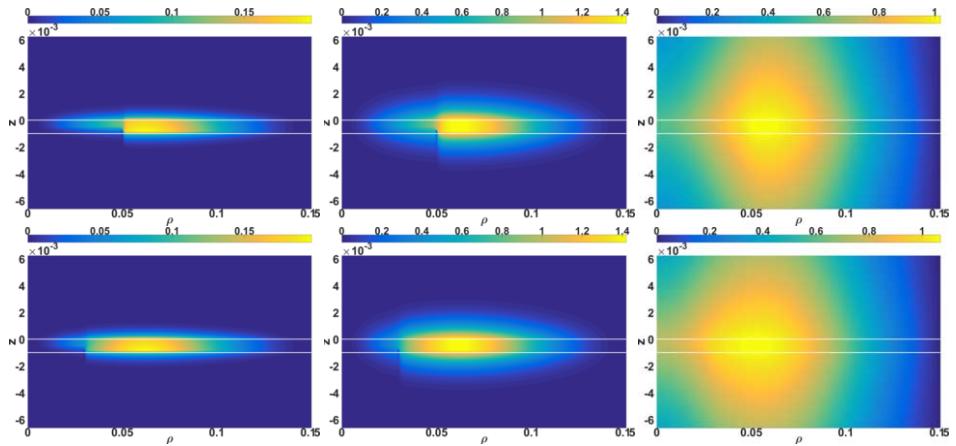


Figure 3. Temperature distribution images in the region of interest for different time and an aluminum plate with defect.

Top: Discontinuity radius: 5 cm. *Bottom:* Discontinuity radius: 3 cm.

Left to the right: Different observation times, $t = t_0$, $t = t_0 + 50$ ms, $t = t_0 + 5$ s.

face, when the radius of the corrosion is 5 cm, is plotted compared with the case of the undamaged plate. In Figure 5 the results with a corrosion of radius of 3 cm are given. Since the position of the thermal camera is not fixed, *i.e.* we can have thermal images on both sides of the plate, the difference of temperature is shown in both surfaces.

On both cases, the information about the defect is clear for early times where we are able to distinguish its edges, *i.e.* the size of the defect in the ρ -direction is well defined. When diffusion occurs, the temperature curves of the damaged and undamaged plate are close. In the case of the smaller defect, still we can see that there is a defect in the workpiece but it is impossible to locate its edges. On the other hand, when the defect's

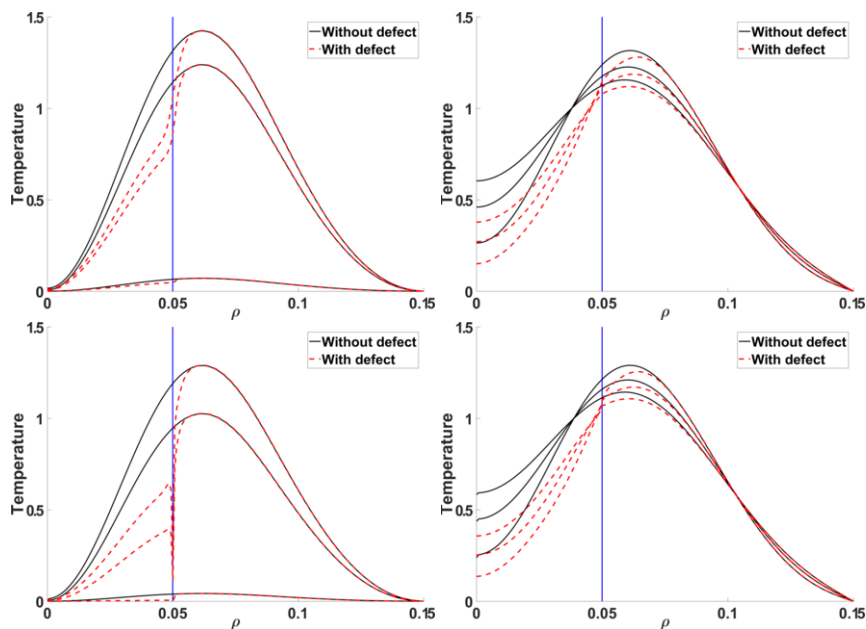


Figure 4. Comparison of the distribution of temperature at the surfaces of the plates with and without defect for different observation times. Defect's radius $r = 5$ cm. *Top:* Upper surface, *Bottom:* Bottom surface. Different observation time: *Left:* 10, 50, 70 ms *Right:* 1, 2, 3s

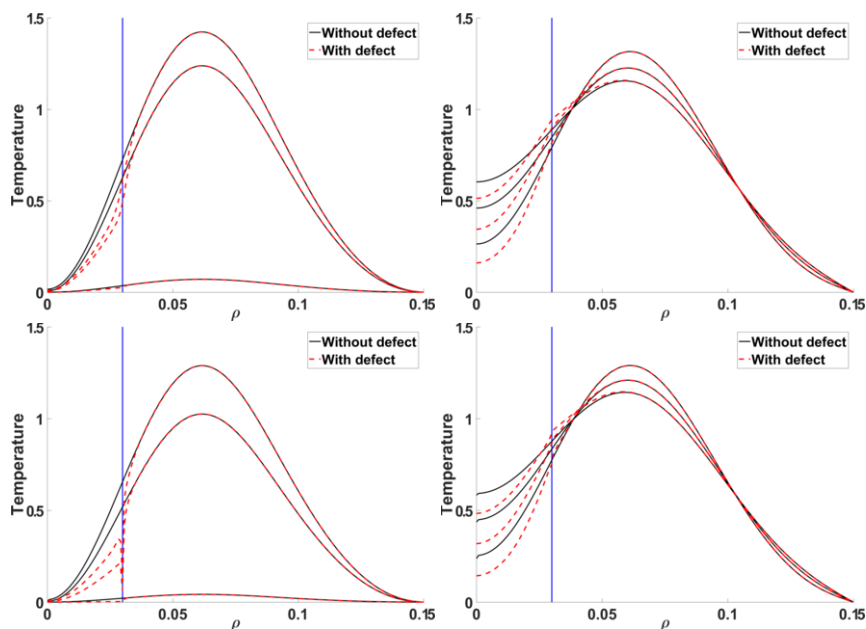


Figure 5. Comparison of the distribution of temperature at the surfaces of the plates with and without defect for different observation times. Defect's radius $r = 3$ cm. *Top:* Upper surface, *Bottom:* Bottom surface. Different observation time: *Left:* 10, 50, 70 ms *Right:* 1, 2, 3s

radius is 5 cm, even after 5 seconds the localization of the edges of the defect can be easily achieved through those curves. In general, in ECT we are interested on early times when the difference of the electromagnetic and thermal properties of the workpiece can highlight any presence of a damage on it and this is well shown by the previous results.

4. Conclusions and perspectives

To conclude, a 2D numerical solver based on the finite integration technique has been implemented and used to simulate the behaviour of different materials under inspection by means of eddy current thermography. Both electromagnetic and thermal problems are coupled in a weak way, taking advantage of the large difference between their characteristic time constants. Possible generalizations of this solver are the development of a 3D version or the investigation of strong coupling in the calculation process.

The use of such a numerical solver is important to investigate easily some aspects like effect of piece inhomogeneity or anisotropy, however it can lead to heavy calculations and complicated meshing considerations when addressing 3D configurations. For this reason, this tool will later be used in complement of fast modal methods to address 3D cases involving canonical geometries like a stratified planar medium.

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Analysis of Damage's Influences on Metal Magnetic Memory Signal

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Abstract. Metal Magnetic Memory (MMM) technique has been extensively used as a qualitative method to test the position of possible damages in ferromagnetic metal structures, a bottleneck in further development of MMM technique is the quantitative relationship between MMM signal and damage characteristic has not been fully revealed. In this paper, analysis of influences of stress and discontinuous structure on MMM signals is presented. Samples of diverse types of structural defects were made and then loaded with different series of stress. The resulting residual deformations and gradients of the magnetic field were measured. Analysis of the quantitative relationship between MMM signals and stress was conducted. The results show: 1) The magnetic field gradient above the structure-continuous zone increases with stress. Therein lies the potential to estimate the stress level point by point. 2) The second-order gradient of the magnetic field near the structure-discontinuous zone increases with the maximum strain in a similar way for different types of discontinuous structures, implying a possibility that the second-order gradient of the magnetic field may be conceive some information about the maximum stress at the edge of discontinuous structures independent of their geometry characteristics.

Keywords. Damage's influences, metal magnetic memory signal, magnetic field gradient

1. Introduction

In real world engineering, there has been a higher demand for non-destructive testing (NDT) methods to determine the state of stresses and deformations in structures. Because the motion of magnetic domains in ferromagnetic material can be influenced by stress and deformation, corresponding magnetic field signals carry information about the status of stress and deformation, some electromagnetic NDT methods, such as magnetic flux leakage (MFL), magnetic barkhausen noise (MBN) and metal magnetic memory (MMM) [1], have attracted much attention and are developing fast in recent years. Among them, MMM technique is time-saving and easily operated, it has been wildly applied in engineering to qualitatively assess the early damage, especially the micro-damage due to local stress concentration in ferromagnetic structure. Its

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capacity for quantitative detection of stress, however, is insufficient because the quantitative relationship between MMM signal and damage characteristic has not been fully revealed, which interest many scholars and arouse their research enthusiasm lately [2-5].

The physical basis of MMM technique is magneto-mechanical effect [6] under the stimulation of the Earth's magnetic field. The magneto-mechanical effect and the leakage of magnetic field caused by the discontinuity of structure are largely regarded as two of the critical factors influencing MMM signal. On the issue of the magneto-mechanical effect of MMM technique, Dong et al. discovered the magnetic gradient (the gradient of the normal component of the residual magnetic field (RMF)) shows a good linearity with evenly distributed elastic stress [7]. Wang et al. proposed a magnetic charge model which established the relationship between stresses and the MMM signal for unevenly distributed stresses [8], whereas whose validity still needs to be tested by further studies.

In addition to the magneto-mechanical effect, another important factor that affects MMM signals is the discontinuous structure which causes the emergence of stress concentration in most case, and promotes each other. Discontinuity cuts off the magnetic circuit within the ferromagnetic structure, the insertion of air increases the intensity of magnetic resistance between the parts of the discontinuous structure, ultimately, affects the tested MMM signals. Roskosz et al. compared the distributions of RMF signals with stress distribution, finding that there are correlations between the discontinuous structure and the RMF signals [9-10], yet the quantitative relationship between them is not clear.

Based on the above background, this paper aims to present some quantitative analysis on how those two factors influence the RMF signals, and further to explore the NDT characterizations of MMM technique to stress concentration.

2. Experimental

The samples are 3-mm thick flat bars made of Q195 steel with a yield stress of 195Mpa. The samples are of different shapes and sizes, type-A, B and C. As shown in Fig.1, type-A sample is made without any artificial defect, type-B sample has a circular hole of 10-mm in diameter in the middle, type-C sample has two 2-mm wide notches on both edges.

The samples are demagnetized by a degauss device and then loaded on a tensile testing machine MST 880. Fourteen samples were tested in total, including five type-A, three type-B and six type-C. Numbers attached on each sample indicate the type of sample and the degree of nominal stress applied, greater number means more stress. Fig.2 shows the stress (σ)-strain (ϵ) curve of A5 sample. The applied nominal stresses are given in Table 1. Each sample is loaded into the plastic deformation range.

According to the knowledge of mechanics of materials, the increase of applied stress during the range of plastic deformation for the tested materials will lead to a faster increase in plastic strain, thus the level of plastic strain is used to quantify the intensity of applied stress hereafter.

During loading, changes of surface deformations within the measuring fields (as shown in Fig.1) of samples are recorded by an ARAMIS V5.4.1 deformation testing device. After loading, samples are removed from the tensile machine, magnetic gradients of the RMF in the measuring fields are scanned by a Bartington Mag-01H

flux gate magnetometer, surface plastic residual strains in the measuring fields are computed by the ARAMIS V5.4.1 deformation testing system. Here, the magnetic gradient (dB/dY) denoted by

$$dB/dY = \frac{\partial B_n}{\partial Y} \tag{1}$$

Where, B_n is the normal component of RMF, Y the coordinate of the testing position along the Y axis of the sample (as shown in Fig. 1).

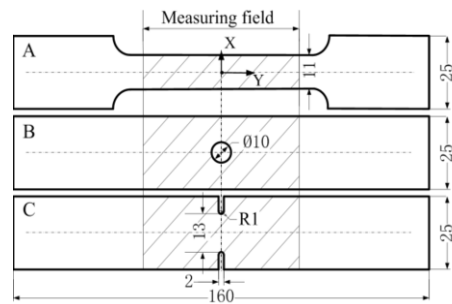


Fig.1. Shapes and sizes of samples

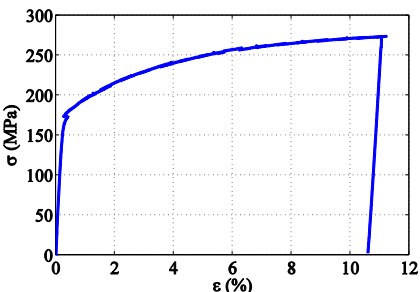


Fig.2. Stress (σ)-strain (ε) curve of A5 sample

Table 1. Applied nominal stresses to samples

Sample	Sample number; Nominal stress (MPa)					
Type A	A1:220.9	A2:248.6	A3:260.0	A4:266.2	A5:273.1	
Type B	B1:200.9	B2:234.2	B3:237.6			
Type C	C1:267.6	C2:297.8	C3:333.1	C4:347.0	C5:351.8	C6:344.2

3. Results

Because the distributions of residual plastic strain for the same-type of samples are similar but with different intensities, Fig.3 just shows the residual strain (ε) distributions of samples of A5, B3 and C5. For the same reason, Fig.4 shows the magnetic gradient (dB/dY) distributions of samples of A5, B3 and C6. Comparing the results displayed in Fig.4 with those of Fig.3, the magnetic gradient distributions look very similar to their strain distributions for each type of sample. The magnetic gradient distributes almost evenly for type-A samples with evenly distributed stress. For type-B and C samples with discontinuous structure, there are stress concentration zones (SCZ) appearing near the hole of type-B sample and the notches of type-C sample, the intensities of magnetic gradient above SCZ are obviously stronger. Especially, above the structure-discontinuous zone, i.e., the hole of type-B samples and the notches of type-C samples, the magnetic gradient intensities reach the maximum value, just like the results shown in Fig.5.

To explore the quantitative relationships of magnetic gradient with stress, 14 sets of dB/dY data on midlines of Y=0 of samples are picked and shown in Fig.6. For different type of samples, there has been a common phenomenon that the magnetic gradient where Y=0 always increases with the applied stress.

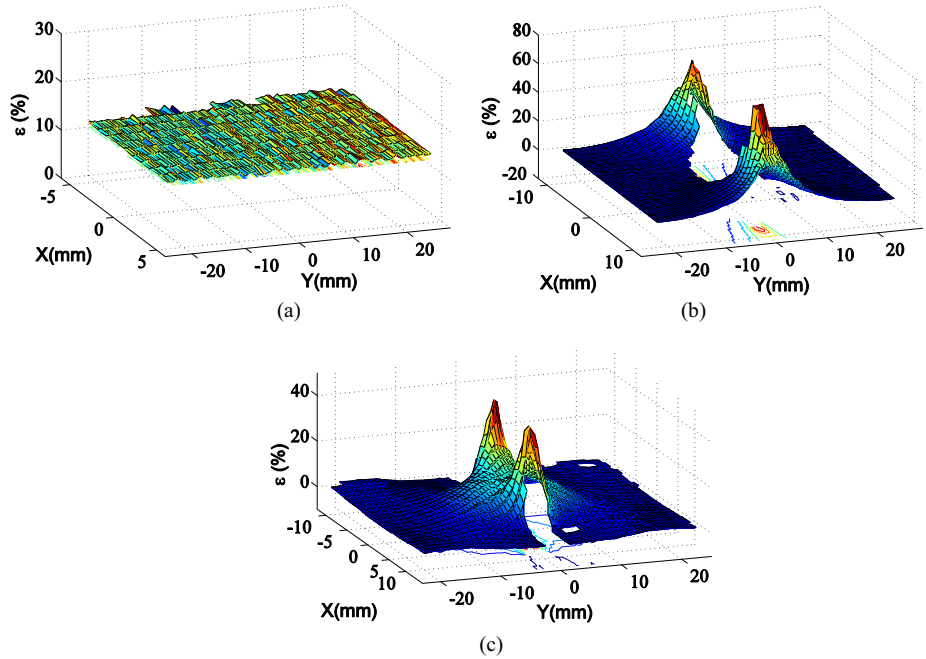


Fig.3. Strain distributions of samples of (a) A5, (b) B3 and (c) C5

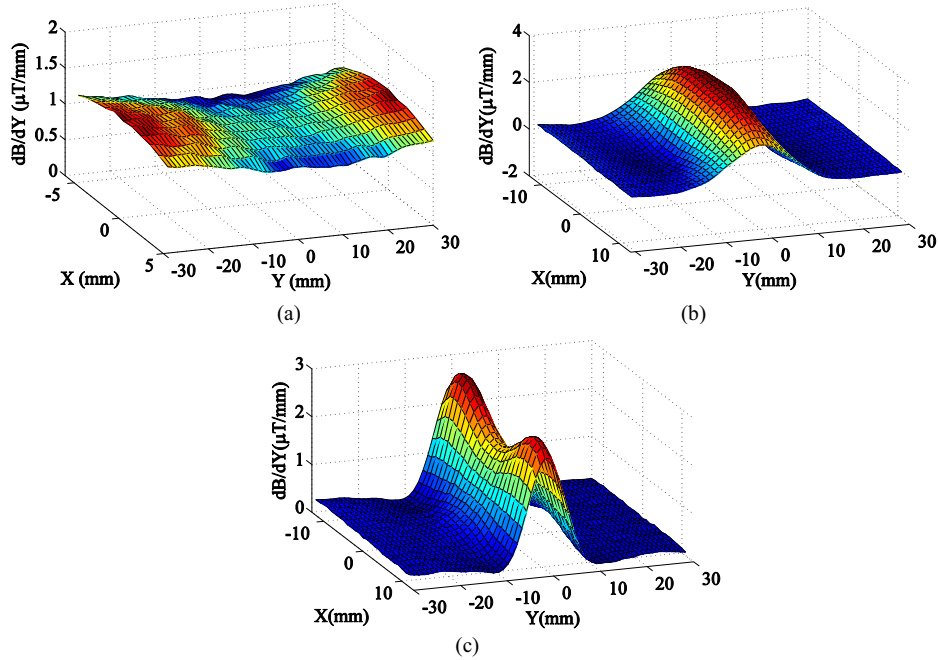


Fig.4. Magnetic gradient distributions of samples of (a) A5, (b) B3 and (c) C6

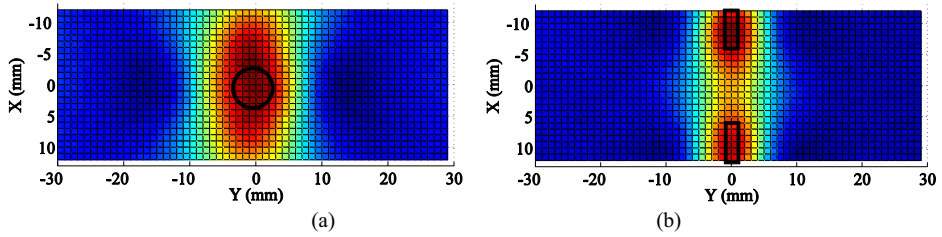


Fig.5. Magnetic gradient distributions above the structure-discontinuous zones of (a) B3, (b) C6 samples

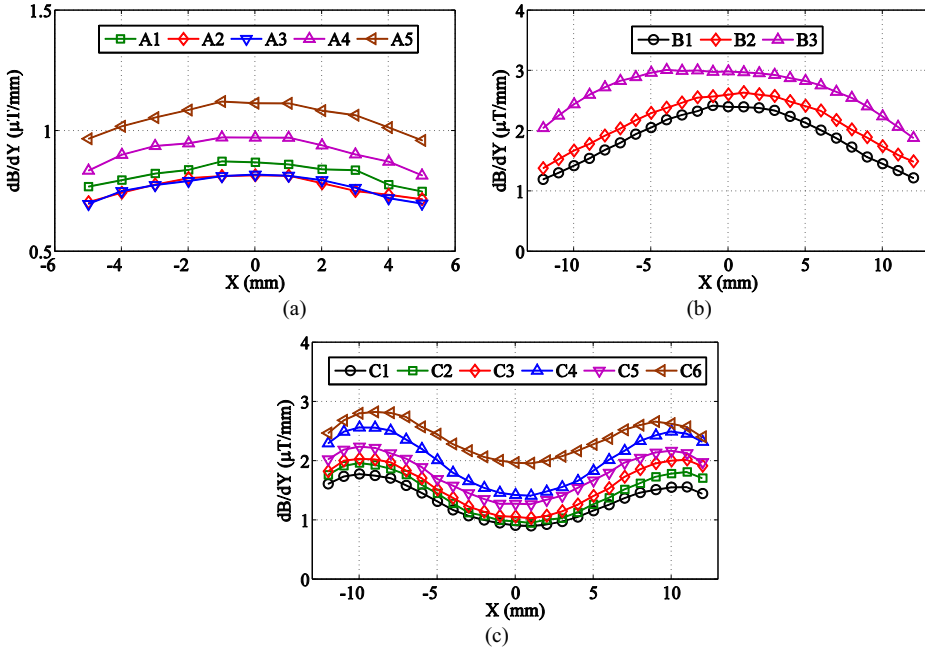


Fig.6. Magnetic gradients where $Y=0$ for samples: (a) Type-A; (b) Type-B; (c) Type-C.

4. Analysis

Analyze the results shown in Fig.3 to Fig.6, the magnetic gradient is stronger where the stress is large as shown in Fig.4, the magnetic gradient increases with the applied stress for each type of sample as shown in Fig.6, it is easy to conclude that the increase of stress leads to the increase of magnetic gradient. But the problem is not that simple when the effect of the discontinuous structure is considered. It is generally known that the discontinuous structure plays an important role in influencing the distribution of MFL signal for the MFL technique. Discontinuous structure breaks the magnetic circuit within the ferromagnetic structure, resulting in the leakage of the magnetic field into the air and thus affects the MFL signal. For MMM technique, it is the same case. Therefore, the RMF signal of MMM technique is affected not only by the stress concentration, but also by the discontinuous structure. In summary, the quantitative relationship between RMF and stress should be analyzed from two aspects:

1) for structure-continuous zone to test the effect of stress; 2) for structure-discontinuous zone to test the effect of the discontinuous structure.

For the structure-continuous zone, to test whether the relationship between the magnetic gradient and stress is the same for different types of sample, points at the center of type-A and C samples are used for the continuity of their structures. The measured magnetic gradients and strains at the center points of type-A and C samples are singled out, the relationship between the magnetic gradient and strain of the center points of type-A and C samples for structure-continuous zone is plotted in Fig.7. The figure shows that the relationship of magnetic gradient with stress is almost the same under two different conditions: evenly distributed stress of type-A, and concentrated distributed stress of type-C. The results give a further confirmation that the magnetic gradient increases with stress disregarding the form of stress distribution, and may be used to estimate the stress level point by point.

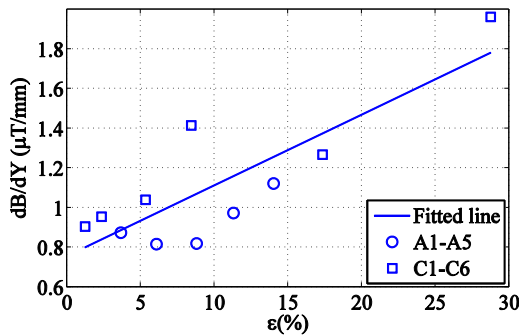


Fig.7. Relationships between magnetic gradient and strain for structure-continuous zone

For the structure-discontinuous zone, based on the previous conclusion that the magnetic gradient increases with stress point by point, it is easy to deduce that the magnetic gradient will change a lot near the edge of the discontinuous structure because where the stress changes to zero. To test this idea, the second derivatives of the magnetic field versus Y position (d^2B/dY^2) of type-B and C samples are calculated and shown in Fig.8 and Fig.9.

$$d^2B/dY^2 = \frac{\partial^2 B_n}{\partial Y^2} \quad (2)$$

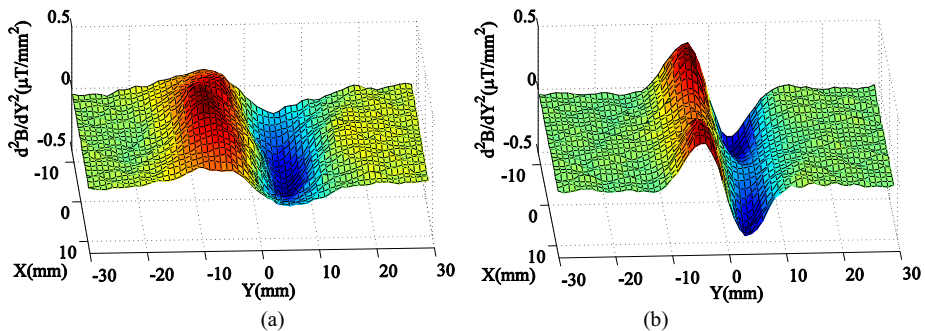


Fig.8. Second derivatives of magnetic field of (a) B3 and (b) C6 samples

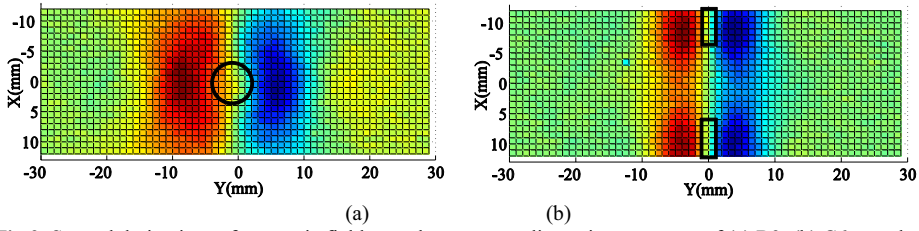


Fig.9. Second derivatives of magnetic field near the structure-discontinuous zones of (a) B3, (b) C6 samples

The results show that the d^2B/dY^2 sure has changed a lot near the edge of the discontinuous structure. To test whether the change rule of d^2B/dY^2 with stress is the same for different types of discontinuous structure, the peak-to-peak value of d^2B/dY^2 (amplitude of d^2B/dY^2) and the maximum strain (used to quantify the degree of stress concentration) for all type-B and C samples are picked out. The relationship between amplitude of d^2B/dY^2 and maximum strain for different discontinuous structures is shown in Fig.10. It shows that the relationships of d^2B/dY^2 with maximum strain for different types of discontinuous structure change with similar rules, which implies a possibility that the RMF characteristic: d^2B/dY^2 may be qualified as a unified quantitative criterion to estimate the maximum stress near the discontinuous structure independent of its geometry.

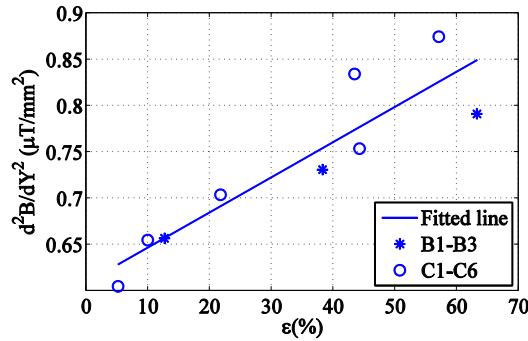


Fig.10. Relationships between second derivative of magnetic field and maximum strain for the structure-discontinuous zones

5. Conclusion

Stress and discontinuous structure are two of the principal factors influencing RMF signal of MMM technique, to get characteristic parameters from RMF signals to quantitatively detect the stress degree, the quantitative relationship between RMF signal and stress, which is quantified by the strain, are studied in this paper. First, it has been found that the magnetic gradient distributes stronger where the strain is larger, the magnetic gradient increases with the applied stress for each type of samples, and the magnetic gradient above structure-continuous zone increases with stress in a similar way for different stress distribution. These results suggest that the magnetic gradient increases with stress disregarding the form of stress distribution, which may be used to assess the stress level point by point. Second, based on the above conclusion, a RMF characteristic: second derivative of magnetic field has been presented. It has been observed that the rules of how the second derivative of magnetic field changes with

maximum strain for different types of discontinuous structure are similar, which implies the second derivative of magnetic field may be qualified as a unified quantitative criterion to estimate the maximum stress in a discontinuous structure.

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Fast Characterization of Multiple Cracks in Conductive Media Based on Adaptive Feature Extraction and SVR

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Abstract. This paper describes real-time cracks characterization and localization inside a Structure Under Test (SUT) by exploiting Learning by Examples (LBE) strategy in the context of Eddy Current Testing (ECT). Within the framework of LBE, an optimal training set has been generated in offline phase by adopting Partial Least Squares (PLS) feature extraction combined with a customized version of output space filling (OSF). Support Vector Regression (SVR) algorithm is utilized for developing an accurate model based on the training set and subsequently real-time inversion (online phase) has been performed on unknown test set. The robustness of the proposed PLS-OSF/SVR approach is numerically assessed in presence of synthetic noisy test set.

Keywords. Eddy Current Testing, Real time inversion, Learning By Examples, Partial Least Squares, Output Space Filling, Support Vector Regression.

Introduction

Eddy Current Testing (ECT) is a popular electromagnetic nondestructive (E-NDT) inspection method that is applied in industries for defect or crack detection, localization and characterization in order to assess the health of the Structure Under Test (SUT). Performance of iterative inversion algorithms, consisting in minimizing a cost function related to the discrepancy between simulation and observation, used for this characterization are affected by ill-posedness and non-linear nature of the problem. In addition, their iterative nature often make them computationally expensive and some of them can be trapped in area realizing local minima of the cost function under minimization [1,2].

Among different non-iterative inverse solutions [3]-[7], Learning by Examples (LBE) strategies have already been presented in [5]-[7] within the context of ECT. In principle, they consist first, in a so-called offline phase, in the generation of a training set of inputs/outputs (I/O) pairs (e.g., ECT signal and the defect parameters), from which an accurate model is built. The available training model from offline phase is then used to predict the output associated to an unknown test sample during online phase. In [5], dif-

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ferent learning algorithms have been studied for crack characterization in a plate structure by adopting standard GRID sampling (i.e., full factorial) approach. The performance of learning algorithms depends on the particular choice of training set generation. Therefore, it becomes essential to select optimal set of training samples of defect parameters and the associate ECT signals for building accurate training model. However, increasing the number of ECT signals (e.g., the number of inspected points from where impedance variation information are collected), requires to increase the number of training samples for optimal model development by a learning algorithm during training phase. This is also known as curse of dimensionality problem for learning algorithms. Higher number of training samples increases the training time as well as the complexity for the learning algorithm to built optimal training model. In [6,7], Partial Least Squares (PLS) feature extraction [8] has been adopted in order to reduce the high dimensional feature space and overcome these issues. In addition, with the increase of number of crack parameters to be inverted, the requirement of higher number training samples by GRID sampling is also increased [5]. The necessity of suitable distribution of crack parameters in the parameter space and associate extracted features in the feature space becomes crucial issue. Hence, an adaptive sampling strategy by utilizing PLS with Output Space Filling (OSF) [9] is also carried out. PLS feature extraction is employed to reduce the actual ECT signal space to the extracted feature space having a much lower cardinality (dimension). The goal of the OSF sampling strategy is to uniformly explore the extracted feature space to have enough information for suitable training set generation. However, the effectiveness of PLS-OSF sampling was not intensively explored in case of high dimensional problem (i.e., the number of parameter to be inverted).

This paper describes applicability of PLS-OSF sampling approach combining Support Vector Regression (SVR) [10] for multi-cracks characterization and localization problem having high dimensional parameter space (Fig. 1). SVR has been utilized for training model construction from the obtained training set and quasi real time inversion during offline and online phases, respectively. The robustness of PLS-OSF/SVR is numerically assessed on noisy test sets and compared with standard approach (e.g., GRID/SVR) [5].

1. Mathematical formulation of forward and inverse problem

Let us consider a conductive and non-ferromagnetic plate having conductivity σ , permeability $\mu_r = 1$, is inspected by a single coil probe working in absolute mode. For the sake of simplicity, assume that the plate is affected by a rectangular crack having geometrical region Ω within the SUT (i.e., $\sigma(\mathbf{r}) \neq \sigma \forall \mathbf{r} \in \Omega$). The crack Ω can be characterized by a finite set of Q descriptors (e.g., length, height, width etc.). The coil measures the ECT signal Ψ_k from k -th ($k = 1, \dots, K$) scanning position by means of impedance variation with respect to the flawless region. Ψ_k can be expressed by the integral form [11,12] in Eq. 1.

$$\Psi_k = \Psi_k(\mathbf{r}_k) = \frac{1}{I^2} \int_{\Omega} \mathbf{E}^{inc}(\mathbf{r}|\mathbf{r}_k) \cdot \rho(\mathbf{r}|\mathbf{r}_k) d\mathbf{r}. \quad (1)$$

$\mathbf{E}^{inc}(\mathbf{r}|\mathbf{r}_k)$ is the incident field generated at position $\mathbf{r}$ in the unflawed plate. $\mathbf{r}_k = (x_k, y_k)$ represents the k -th coil position within the plate and I is the flowing current inside the

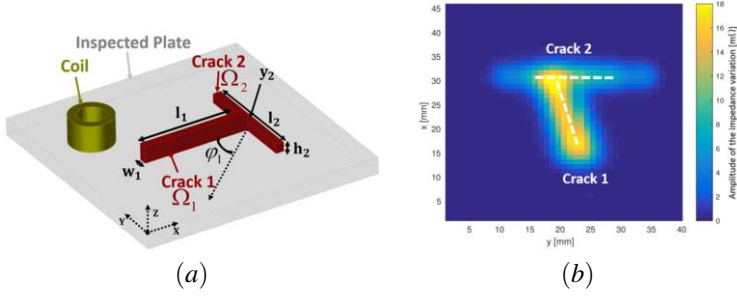


Figure 1. Example of (a) studied plate geometry and (b) amplitude of coil impedance variation.

coil. The unknown current dipole density $\rho(\mathbf{r}|\mathbf{r}_k)$ represents the perturbation of the eddy current (Eq. 2), while $\mathbf{E}^{Tot}(\mathbf{r}|\mathbf{r}_k)$ is the generated total field in presence of the crack [11].

$$\rho(\mathbf{r}|\mathbf{r}_k) = [\sigma(\mathbf{r}) - \sigma]\mathbf{E}^{Tot}(\mathbf{r}|\mathbf{r}_k) \quad (2)$$

An implementation of this integral formulation in the CIVA software [13] is used as a forward operator $\Phi\{\cdot\}$ for generating ECT signal vector, $\underline{\Psi} = (\Psi_k; k = 1, \dots, K)$ for the set of defect descriptors vector $\underline{p} = (p_q; q = 1, \dots, Q)$. Whereas $\Psi_k = \Phi\{\mathbf{r}_k, \underline{p}\}$ satisfies the relationship of Eq. 1. Within the scope of LBE, an (unknown) inverse operator $\hat{\Phi}^{-1}\{\cdot\}$ can be formulated by suitably processing the ECT signal $\underline{\Psi}$ and the corresponding crack parameters $\underline{p}$ through a set of N I/O pairs (i.e., training set). The following steps describe the iterative procedure for obtaining suitable I/O pairs to build optimal training model.

Initialization- Generate N_0 number of initial samples by using uniform GRID sampling within the crack parameter space. A matrix of crack parameters $\underline{p} = (\underline{p}^{(n)}; n = 1, \dots, N_0)$ having $(N_0 \times Q)$ dimension is formed, where $\underline{p}^{(n)}$ is the n -th row of $\underline{p}$. By using $\Phi\{\cdot\}$ generate ECT coil signals and fill the $(N_0 \times F)$ feature matrix $\underline{\Psi}$ which n -th row is represented by $\underline{\Psi}^{(n)}$. Due to the complex nature of the ECT signals, $\underline{\Psi}^{(n)}$ is represented by $\underline{\Psi}^{(n)} = \left\{ \left(\Re\{\Psi_k^{(n)}\}; \Im\{\Psi_k^{(n)}\} \right); k = 1, \dots, K \right\}$ of $F = 2K$ ECT features.

Feature extraction- Build the $(N_0 \times F)$ matrix $\underline{\Psi}'$ by subtracting to each f -th column of $\underline{\Psi}$ ($f = 1, \dots, F$) its mean value μ_f and compute the $(N_0 \times Q)$ matrix $\underline{p}'$ by subtracting to each q -th column of $\underline{p}$ its mean value μ_q ($q = 1, \dots, Q$). Apply the PLS algorithm to linearly decompose $\underline{\Psi}'$ and $\underline{p}'$ as follows

$$\underline{\Psi}' = \underline{T} \times \underline{S} + \underline{Y}; \quad \underline{p}' = \underline{U} \times \underline{Z} + \underline{G} \quad (3)$$

In Eq. (3), $\underline{T} = (\underline{T}^{(n)}; n = 1, \dots, N_0)$ is the $(N_0 \times J)$ matrix of Ψ -scores where $\underline{T}^{(n)} = (T_j^{(n)}; j = 1, \dots, J)$. It is obtained from $\underline{\Psi}'$ by calculating $(F \times J)$ weight matrix $\underline{W}$ (i.e., $\underline{T} = \underline{\Psi}' \times \underline{W}$) using SIMPLS [8] algorithm. J is the number of extracted features ($J < F$). $\underline{Y}$ and $\underline{G}$ contain the $(N_0 \times F)$ and $(N_0 \times Q)$ residuals of the linear decomposition while $\underline{S}$ and $\underline{Z}$ are the $(J \times F)$ and $(J \times Q)$ matrices of loadings. The decomposition in Eq. (3) aims at maximizing the co-variance between the corresponding columns of $\underline{T}$ [i.e., $(\bar{T}_j; j = 1, \dots, J)$] and of the $(N_0 \times J)$ matrix of p -scores $\underline{U}$ [i.e., $(\bar{U}_j; j = 1, \dots, J)$]. This

guarantees all the information about the ECT signal embedded inside $\underline{\Psi}'$ (i.e., inside $\underline{\Psi}$) is compressed into $\underline{T}$. Assign the number of training sample $N_{iterative} = N_0$ and construct an initial training set $\hat{D}_{N_{iterative}} = \left\{ \left(\underline{T}^{(n)}; \underline{p}^{(n)} \right); n = 1, \dots, N_{iterative} \right\}$ for the adaptive step. **Adaptive sampling-** By using Latin Hypercube Sampling (LHS), select V candidate samples within the parameter space by $\underline{p}_{cand}^{(v)} = \left(p_{cand,q}^{(v)}; q = 1, \dots, Q \right)$, $v = 1, \dots, V$. An estimation of the J -dimensional set of extracted features corresponding to each v -th candidate $\tilde{\underline{T}}_{cand}^{(v)}$ is retrieved by applying a multi-dimensional linear interpolator on $\hat{D}_{N_{iterative}}$. Select the required candidate sample $v = v^*$ through maximizing the minimum distance between the obtained $\tilde{\underline{T}}_{cand}^{(v^*)}$ and the $\underline{T}^{(n)} (n = 1, \dots, N_{iterative})$ sets in $\hat{D}_{N_{iterative}}$ [i.e., $v^* = \arg \left(\max_v \left\{ \min_n d_{vn} \right\} \right)$]. $d_{vn} = \sqrt{\sum_{j=1}^J \left(\tilde{T}_{cand,j}^{(v)} - T_j^{(n)} \right)^2}$ is the Euclidean distance between $\tilde{\underline{T}}_{cand}^{(v)}$ ($v = 1, \dots, V$) and $\underline{T}^{(n)}$. By utilizing $\Phi\{\cdot\}$, the set of F ECT fea-

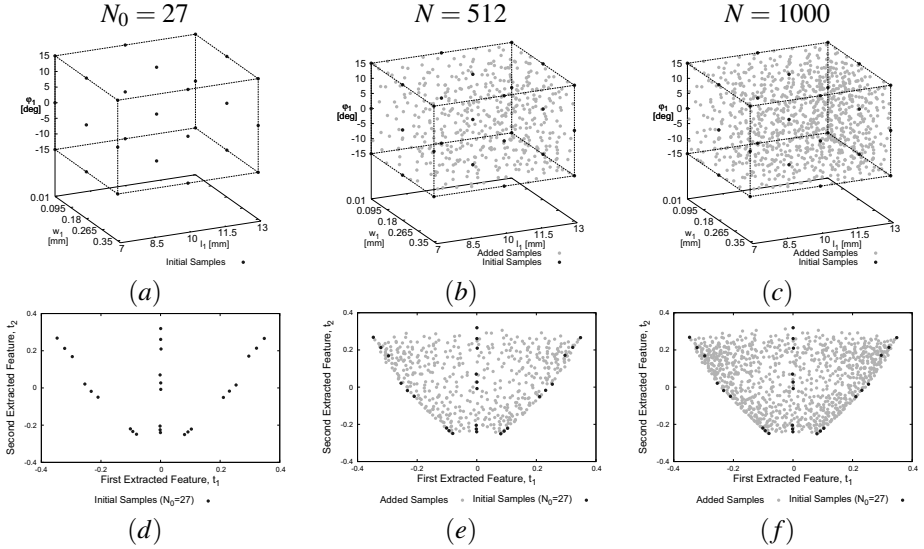


Figure 2. Disposition of the training samples generated by PLS-OSF in (a)(b)(c) the crack parameter space and (d)(e)(f) the extracted feature space, for (a)(d) $N_0 = 27$, (b)(e) $N = 512$ and (c)(f) $N = 1000$.

tures $\underline{\Psi}_{cand}^{(v^*)} = \left\{ \left(\Re \left\{ \Psi_{cand,k}^{(v^*)} \right\}; \Im \left\{ \Psi_{cand,k}^{(v^*)} \right\} \right); k = 1, \dots, K \right\}$ associated to the selected candidate sample v^* is computed. Finally, the set of extracted features is retrieved by $\underline{T}_{cand}^{(v^*)} = \left(\underline{\Psi}_{cand}^{(v^*)} \right)' \times \underline{W}$, where $\left(\underline{\Psi}_{cand}^{(v^*)} \right)'$ is obtained by subtracting to each f -th element of $\underline{\Psi}_{cand}^{(v^*)}$ ($f = 1, \dots, F$) its mean value μ_f . Finally, the training set is updated with $\hat{D}_{N_{iterative}+1} = \hat{D}_{N_{iterative}} \cup \left\{ \underline{T}_{cand}^{(v^*)}; \underline{p}_{cand}^{(v^*)} \right\}$ and $N_{iterative} = N_{iterative} + 1$. This is the customized version of OSF algorithm [9] (e.g., candidate parameters are chosen such that the extracted features are uniformly distributed).

Stop criterion- The adaptive sampling step adds new sample iteratively until $N_{iterative} = N$ (N is desired/feasible training size).

At this stage, SVR has been utilized to train separately q -th set of I/O pairs $\hat{D}_{N,q} = \left\{ \left(\underline{T}^{(n)}; p_q^{(n)} \right); n = 1, \dots, N \right\}$ on the generated training set for each q -th parameter ($q = 1, \dots, Q$) of the crack. Moreover, a test set $\underline{\Psi}_{test}$ of F ECT features associated to a previously-unseen crack parameter configuration $\underline{p}_{test}$ is projected through $\underline{W}$ onto the J -dimensional PLS-extracted features space [i.e., $\underline{T}_{test} = \underline{\Psi}'_{test} \times \underline{W}$]. $\underline{\Psi}'_{test}$ is obtained by subtracting to each f -th element of $\underline{\Psi}_{test}$ ($f = 1, \dots, F$) its mean value μ_f . Finally, $\underline{T}_{test}$ is given as input to the corresponding trained model in order to estimate the q -th parameter of the crack, $\tilde{p}_{test,q}$ ($q = 1, \dots, Q$). Figure 2 represents the exploration of adaptive training samples generation in the parameter space, as well as in extracted feature space. For visualization purpose, the training set of three dimensional parameter space with the corresponding two dimensional extracted features space are shown by varying the parameters of Crack 1 and fixing Crack 2 parameters.

2. Training and test set configuration

This section describes the numerical configuration of the treated problem for performing inversion. A homogeneous plate of thickness 1.27 mm, conductivity 1.02 MS/m and permeability $\mu_r = 1$ is affected by two cracks, as shown in Fig. 1. Crack 1 is a surface breaking crack Ω_1 of 1.27 mm height, while Crack 2 is a subsurface crack Ω_2 having fixed width 0.19 mm. Ω_1 and Ω_2 are characterized by total $Q = 6$ descriptors $\underline{p} = (l_1, \phi_1, w_1, l_2, h_2, y_2)$. l_1, ϕ_1, w_1 represent the length, angular displacement and width of Crack 1 respectively and Crack 2 is characterized by length l_2 , height h_2 and axial position y_2 along y axis. The plate is inspected by a single coil working in absolute mode at a frequency of $f = 500$ kHz. The coil collects the ECT signal from 46 and 41 positions along X and Y directions, respectively with a step size of 0.5 mm through a raster scan. More details on coil parameters can be found in [14]. An efficient training set has been built by CIVA [13]. To partially consider noise effects, Additive White Gaussian Noise (AWGN) with a signal-to-noise ratio (SNR) is applied for blurring ECT signal, where $SNR = 10 \log_{10} \left\{ \frac{\sum_{k=1}^K |\Psi_k|^2}{\sum_{k=1}^K |\mathfrak{N}_k|^2} \right\}$. $\mathfrak{N}_k$ is the complex valued additive noise (random number), which corrupts the ECT signal Ψ_k at k -th inspected point. The performance of the inversion scheme is analyzed by computing the normalized mean error (NME), expressed as $NME_q = \frac{1}{M} \sum_{m=1}^M \frac{|p_q^{(m)} - \tilde{p}_q^{(m)}|}{|p_q^{(m)}|}$. $p_q^{(m)}$ and $\tilde{p}_q^{(m)}$ are the actual and predicted q -th parameter of the m -th ($m = 1, \dots, M$) unknown test sample, respectively. Six training sets for six different trained models have been created by changing the parameters of both cracks within the range $l_1 \in [7, 13]$ mm, $\phi_1 = [-15, 15]$ deg, $w_1 = [0.01, 0.35]$ mm, $l_2 \in [7, 13]$ mm, $h_2 = [0.254, 1.016]$ mm and $y_2 = [24, 31]$ mm. The sets of $N = 4096$ training samples have been built by PLS-OSF sampling methods within the selected cracks parameter ranges and the corresponding ECT signals (i.e., impedance variation signal) are collected from $K = 46 \times 41 = 1886$ number of inspected points. $N_0 = 729$ and $N = 4096$ are chosen for the appropriate performance comparison between PLS-OSF/SVR and GRID/SVR (i.e., complete $3^6 = 729$ and $4^6 = 4096$ factorial design) strategies. The details of the GRID/SVR can be retrieved from [5]. Each of these samples is associated to $F = 2K = 3772$ ECT features (i.e., applied for GRID/SVR) which have been reduced to $J = 15$ extracted features by PLS feature extraction for PLS-OSF/SVR

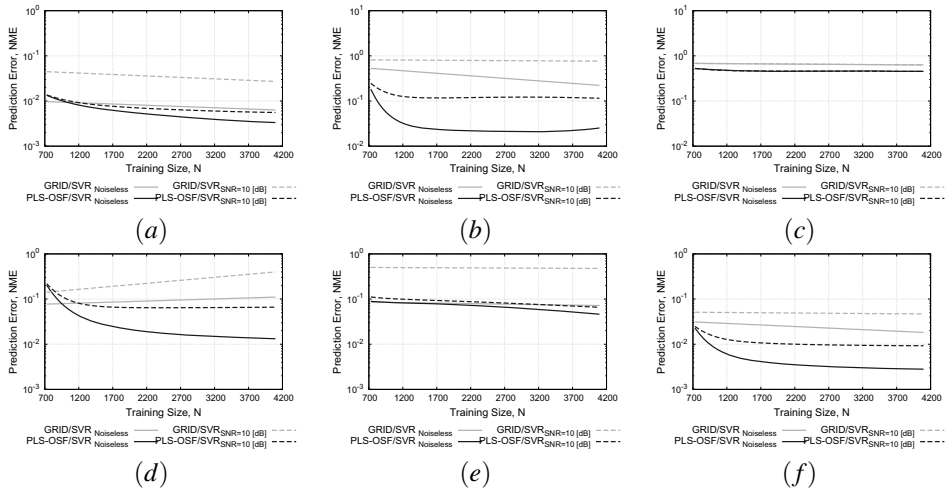


Figure 3. Training size, N versus prediction error, NME at (a) $p_1 = l_1$, (b) $p_2 = \phi_1$, (c) $p_3 = w_1$, (d) $p_4 = l_2$, (e) $p_5 = h_2$ and (f) $p_6 = y_2$ for noisy ($SNR = 10$ [dB]) and noiseless test sets by PLS-OSF/SVR and GRID/SVR.

strategy. $V = 300$ candidate samples have been chosen randomly by *LHS* during each iteration of adaptive loop. An unknown test set of $M = 1000$ samples for 6 crack parameters has been generated by using *LHS* design. As for the training set, the same number of ECT features (i.e., $F = 2K = 3772$) is also considered for each test sample for treating training model obtained by GRID/SVR. In case of PLS-OSF/SVR approach, $J = 15$ features are extracted by projecting the test set into extracted feature space (e.g., the one obtained from training set). The hyper parameters (e.g., optimum values of C and γ parameters) of SVR are tuned by exploiting 5-fold cross-validation during the training phase on the different training sets obtained from different training strategy (i.e., PLS-OSF/SVR and GRID/SVR). For utilizing, ϵ -SVR, ϵ is kept fixed at 10^{-1} . Different training models (offline phase) are obtained separately by different sampling strategies with the corresponding SVR parameters. Finally, the estimation of crack parameters on the unknown test set is evaluated (online phase) by the trained models.

3. Results and discussion

Having a six dimensional parametric problem at hand requires to generate high number of training samples, which is difficult when using standard GRID approach. Applying PLS features extraction, ensures most significant information are projected to a much reduced extracted feature space from actual ECT features. By uniformly distributing the extracted feature space, redundant and unnecessary samples for different configuration of cracks parameters are avoided. As a consequence, more accurate models are obtained from smaller training sets. Figure 3 depicts the inversion performance comparison between PLS-OSF/SVR and GRID/SVR approaches for noiseless and noisy (i.e., $SNR=10$ dB) test sets. Whereas, GRID/SVR shows higher prediction error for estimating almost all the parameters of Crack 1 and Crack 2 than PLS-OSF/SVR. PLS-OSF/SVR has shown much improved prediction accuracy on noiseless as well as noisy test sets for l_1 , ϕ_1 and w_1 estimation once N is larger than 1000 samples. ECT signal does not change

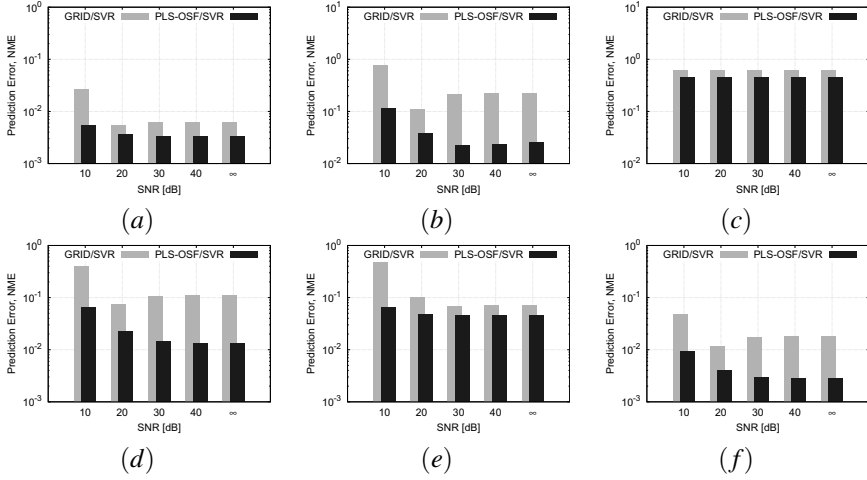


Figure 4. Signal to Noise Ratio, SNR versus Prediction error, NME at (a) $p_1 = l_1$, (b) $p_2 = \phi_1$, (c) $p_3 = w_1$, (d) $p_4 = l_2$, (e) $p_5 = h_2$ and (f) $p_6 = y_2$ for training size, $N = 4096$ through PLS-OSF/SVR and GRID/SVR.

significantly due to the variation of crack width w_1 (the probe diameter is larger than the gap size on this test case). Thus, crack width estimation shows higher prediction error among all the cracks parameters for both GRID/SVR and PLS-OSF/SVR approaches. From Fig. 1(b), we can see that impedance variation signals for the presence of Crack 1 (surface breaking crack) is much stronger than the one due to Crack 2 (subsurface crack). Hence, ECT signals due to the presence of Crack 2 are more prone to be affected by the noise. Moreover, due to the crack parametric definition, we are varying the axial position y_2 while varying the length l_2 . This makes the studied test case more complex to estimate l_2 . Even when increasing N , the prediction accuracy of l_2 estimation by GRID/SVR approach is not improved. By adopting PLS-OSF, the extracted feature space is filled in such a way that the contribution of each parameter is retained and this phenomenon reflects higher prediction accuracy for l_2 , h_2 and y_2 estimation than GRID/SVR approach. Prediction accuracy of different crack parameters, in terms of error (NME), are also shown in Fig. 4 for different SNR for $N = 4096$ samples. It is also visible that PLS-OSF/SVR has higher prediction accuracy than GRID/SVR for predicting all crack parameters on both noisy and noiseless test sets. Moreover, PLS-OSF/SVR takes $\sim 0.05s$ for predicting $M = 1000$ test samples where GRID/SVR takes $\sim 6s$.

4. Conclusions

In this paper, we have shown quasi real-time multi-cracks characterization and localization by combining an adaptive sampling strategy through PLS feature extraction with OSF and SVR. PLS-OSF/SVR has shown higher prediction accuracy for performing inversion on the studied problem having six parameters for a training set of $N \geq 1000$ samples than GRID/SVR [5]. One of the significant advantages of PLS-OSF strategy is to project the high dimensional ECT signals to a different and much lower dimensional space by PLS. However, the number of parameters to be inverted depends on the com-

plexity of the problem, the suitable set of ECT features and the training set size N . Indeed, provided a problem with a certain degree of complexity, increasing the number of dimensions leads to an (more than proportional) increase of the training set size (i.e., N). This fact can lead to a cumbersome and very time consuming offline phase, where hundreds of thousand forward calls could be required. Moreover, PLS-OSF/SVR can be applied for dealing with parallel cracks placed in a reasonable distance ($\geq 10\mu\text{m}$). In case of smaller distance, ECT signal variation due to the presence of one crack approaching the other would be very minor. This makes the inversion as ill-posed problem which does not have any unique solution.

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Thickness Measurements Using Extremely Low Frequency Eddy Current Testing via TMR Sensors Operated with AC Modulation

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Abstract. Conventionally, eddy current testing (ECT) has been used for the surface crack detection. However, the detection of underside and internal corrosion was difficult using this technique. Therefore, we developed an underside and internal corrosion detection technique based on extremely low frequency ECT (ELECT). To use the TMR sensor for ELECT, a TMR sensor operated with AC modulation was applied as a magnetic sensor. Then, we developed the ELECT system using a TMR sensor operated with AC modulation, and succeeded in measuring the thickness of thick steel plate.

Keywords. ELECT, TMR sensor, AC modulation, SAM

1. Introduction

Among the non-destructive testing (NDT) methods, eddy current testing (ECT) has advantages such as high speed, high precision, and robustness. Although conventional ECT has been used for surface crack detection, it is difficult to apply for deep crack detection. To obtain more information depending on the skin depth, measurements must be carried out with an extremely low frequency magnetic field. Therefore, our group developed ELECT, an extremely low frequency eddy current testing system that detects underside and inside defects as well as surface defects in steel materials [1-3]. With conventional ECT methods that use pick up coils to detect high frequency magnetic fields, there are often issues when skin depth is shallow. Therefore, instead of the pick-up coil, ELECT uses a magnetic sensor that can carry out high-sensitivity measurements, even at low frequency magnetic fields. There are various types of magnetic sensors such as hall sensors, magneto resistive (MR) sensors, and superconducting quantum interference devices (SQUID). Among them, the MR sensor uses a material whose resistance changes according to the applied magnetic field. According to the difference in the thin film structures, there are anisotropic magneto resistive (AMR) sensors, giant magneto resistive (GMR) sensors, and tunnel magneto resistive (TMR) sensors. Among them, the TMR sensor is characterized by small size and high sensitivity. However, it has even numbered characteristics. TMR sensors cannot distinguish the polarity of magnetic fields, and can only be operated in linear regions for use in NDT methods. In general, to solve these problems, a DC biased magnetic field generated by a permanent magnet or a direct current is applied to the TMR sensor, and the TMR sensor is operated in a linear region.

However, it is difficult to arrange permanent magnets to apply an appropriate magnetic field and obtain stability with DC magnetic fields. For these reasons, an AC modulation method was developed as a new driving method for TMR sensors [4]. In this study, we developed the ELECT system using a TMR sensor operated with AC modulation to evaluate the thickness of steel plates using the new system.

2. ELECT system

2.1. TMR sensor magnetic probe

In this study, a nanogranular in-gap magnetic sensor was used as the TMR sensor [5]. Metal nanogranules are interspersed in the insulating material. The magnetic spins of these nanogranules have random directions if there is no external magnetic field, so that adjacent magnetic spins are antiparallel and the tunnel current is small. However, parallel magnetic spins increase by increasing the external magnetic field and the tunnel current.

The TMR sensor used in this study had the half bridge structure of a high sensitivity element and a low sensitivity element (resistances of both elements were 100 k Ω). The drive voltage was set to 5 V, and the output voltage was obtained at the connection point of both elements. The output from the TMR sensor was amplified by an IC amplifier that can amplify 10 times. In the magnetic sensor probe, a modulation coil for AC modulation was wound around the sensor as close as possible (Figure. 1(a)). The magnetic probe for ELECT was equipped with an induction coil for generating an eddy current in an inspection object and a cancelling coil for cancelling the induction magnetic field entering the TMR sensor (Figure. 1(b)). The induction coil and the cancelling coil were connected in series not to cause phase shift. The inner diameter of the induction coil was 21 mm, with 50 turns, and the cancelling coil had 15 turns.

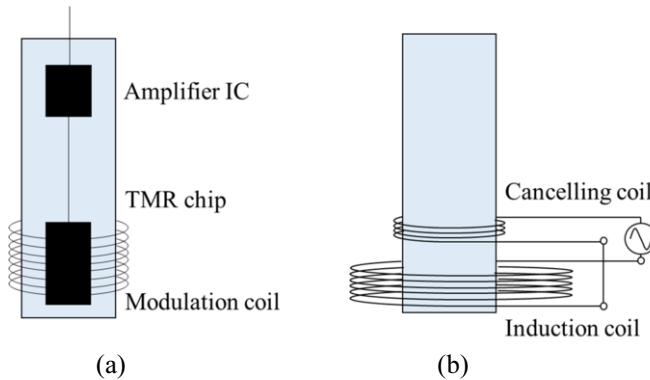


Figure 1: Diagrams of (a) TMR sensor probe, (b) the magnetic probe for ELECT

2.2. Theory of AC modulation

Figure 2 shows the DC characteristics of a TMR sensor. A TMR sensor has even numbered characteristics ($f(-x) = f(x)$), and its output signals contain many second harmonic frequency components when an alternating magnetic field is applied to it. When the external alternating magnetic field B_{ex} with frequency f_{ex} is modulated by the

modulation magnetic field B_{mod} with frequency f_{mod} , the fundamental frequency component varies according to B_{ex} (Figure 2(a)). Therefore, by detecting the output with the modulation frequency f_{mod} , the output signal proportional to B_{ex} is obtained (Figure 2(b)).

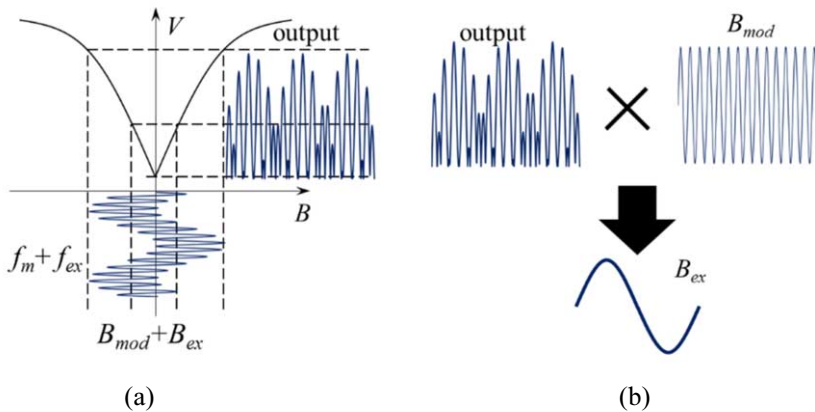


Figure 2 (a) Concept of AC modulation, (b) Concept of demodulation

2.3. Magnetic response characteristics of TMR sensor magnetic probe

The magnetic response characteristics of TMR sensor magnetic probe were evaluated in a magnetic shield to isolate environmental magnetic noise. A solenoid coil generating a magnetic field was placed in the shield, and the probe was placed at the center of the coil. Outside the shield, an AC signal was generated by a function generator (FG) and added to the current source, so that the alternating current to be supplied to the coil was generated. The TMR sensor and the modulation coil were driven by a power supply. The output from the probe was inputted to the first lock-in amplifier for demodulation, by detecting at the modulation frequency. Subsequently, the external magnetic field component was obtained by a second lock-in amplifier detecting at the external magnetic field frequency. Finally, a PC was used for data acquisition (Figure 3).

The magnetic response of the TMR sensor magnetic probe is shown in Figure 4. Figure 4(a) shows the fundamental magnetic response curve, and Figure 4(b) shows the AC magnetic response when an external alternating magnetic field of 50 Hz was modulated by a modulation magnetic field of 5 kHz, 300 to 500 μ T. The figure shows that the magnetic response using AC modulation can obtain linear response characteristics by changing the intensity of the external alternating magnetic field. In addition, the magnetic noise characteristics indicate the detection limit intensity of the magnetic sensor. As an evaluation method, the magnetic noise spectrum was measured when operating with a modulation magnetic field of 5 kHz and 500 μ T. In order to calculate the magnetic field intensity from the magnetic noise characteristics, we compared the magnetic noise spectrum when there is an external magnetic field with 50 Hz and 200 μ T, and when there is no external magnetic field (Figure 5). This figure shows that the intensity of the external magnetic field corresponds to a peak at 50 Hz of the noise peak, and the noise at the same 50 Hz, when there is no external magnetic field, is 11 nT/ $\sqrt{\text{Hz}}$, which indicates the detection limit of the probe at 50 Hz. From this result, the TMR sensor magnetic probe operating with AC modulation can be used for ELECT.

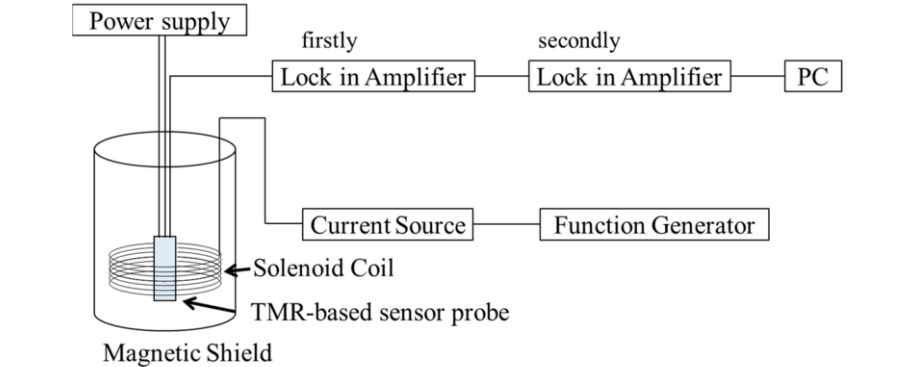


Figure 3 System diagrams of measurement magnetic response characteristics of TMR sensor magnetic probe

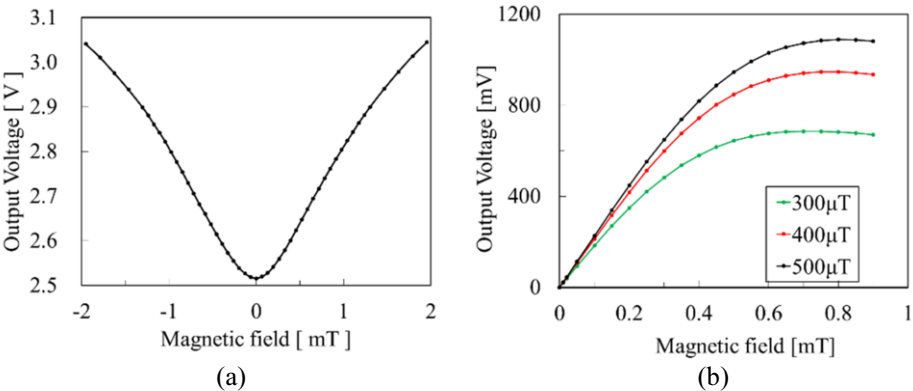


Figure 4 Magnetic response characteristics operated (a) without AC modulation (b) with AC modulation

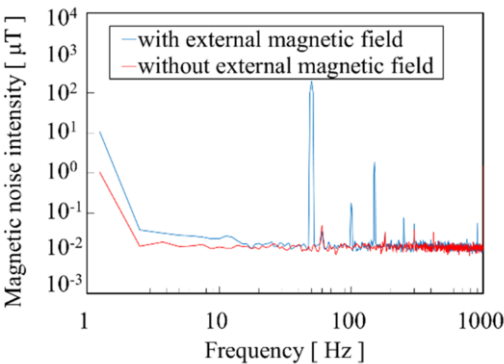


Figure 5 Comparison of the magnetic noise spectrum of the TMR sensor

2.4. Measurement system for ELECT

The measurement system for ELECT consists of the above probe, function generator (FG), current source, power supply, lock-in amplifier, and personal computer. The TMR sensor and modulation coil (modulation magnetic field is 5 kHz and 500 μ T) are driven by the power supply. The alternating current signal generated by FG is converted into an alternating current (0.5 A) by the current source and the applied induction coil. The output from the probe is inputted to the first lock-in amplifier that demodulation detecting by the modulation frequency. Subsequently, the second lock-in amplifier detects at the same frequency as the applied magnetic field frequency, and the intensity and phase of the magnetic field generated eddy current are analyzed.

3. Thickness measurement

To apply ELECT using a TMR sensor with AC modulation, the thickness of a steel plate was measured. There are many instances when it is important to check the thickness of a steel plate, such as when inspecting thinning due to a corrosion of the steel structure. A steel plate made of SM490, ranging in a thickness of 6 to 16 mm, was used as a sample. In order to detect the change in the thickness of the steel plate, a sufficient skin depth, at which the eddy current penetrates, is required. The extremely low frequency magnetic field of 3 Hz to 100 Hz was applied, and a magnetic vector with an intensity and phase was obtained due to the eddy current at each frequency. Then, a magnetic spectrum was obtained by tracing the magnetic vectors. However, under these conditions, the starting points of the magnetic spectrums were not aligned due to the influence of surface magnetization in the ferromagnetic steel plates, and the evaluation was not possible. Thus, the measured signals at 3 Hz for each sample were subtracted from each of the magnetic vectors obtained at other frequencies, and set as the starting point from the origin. By the spectrum analysis of the magnetic field (SAM) [2], the effect of surface magnetization can be mitigated, and the starting point of the spectrum can be aligned (Figure 6(a)). However, in Figure 6(a), the thickness of the steel plate cannot be distinguished, because of the overlapping of spectrums at frequencies higher than 20 Hz. This is because the eddy current is concentrated near the surface due to the skin depth, and has not reached the required depth. Subsequently, the spectrum of the low frequencies 3 Hz to 20 Hz was focused on (Figure 6(b)). Figure 6(b) shows that the spectrum rotates in the clockwise direction according to the plate thickness. This application of ELECT shows that plate thickness can be distinguished based on the spectrum rotation for plate thicknesses up to 16 mm.

4. Conclusion

The TMR magnetic sensor was applied to the eddy current testing. We developed a new driving method of AC modulation that applies a high frequency alternating magnetic field as a bias. In this method, the linear region was increased and the noise level was successfully reduced, and it can be used for eddy current testing in NDT. We have applied the method to the thickness measurement of a steel plate by extremely low frequency eddy current testing (ELECT), and succeeded in evaluating the plate thickness up to 16 mm.

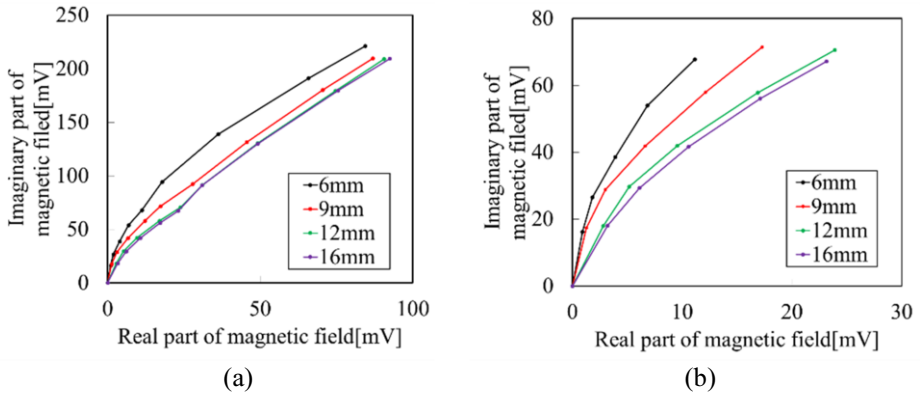


Figure 6 Magnetic spectrum of (a) 3 Hz to 100 Hz (b) 3 Hz to 20 Hz

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Experimental Evaluation of a Flexible Eddy Current Sensor Pattern to Characterize the Hardness of Burned Area in Ferromagnetic Material

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Abstract. Eddy Current is a powerful means of detection of defects located close to the surface in conductive parts. Defects to be detected can classically be flaws, but also variations of a mechanical or electromagnetic property of material. A typical Eddy Current pattern is composed of an emitting coil and a receiving coil. These coils can be etched on a kapton to design flexible probes able to fit complex shaped of pieces. By optimization of coil diameters and distance between coils, it is possible to obtain specific properties of the pattern. In this paper, we use a flexible pattern optimized for the detection of the variation of conductivity, thanks to a great sensitivity to this parameter and a straight response in the complex plane when it evolves. The applicability of this pattern is shown to detect and characterize burned areas of different intensities and their consequences on material hardness.

Keywords. Flexible eddy current sensor, ferromagnetic material, burned area, hardness.

1. Introduction

One of the issues in Nondestructive Testing (NDT) by Eddy Current (EC) concerns its high sensitivity to detect surface defects, or defects close to the surface, in conductive and possibly magnetic materials [1, 2]. These examinations are commonly used in many industrial sectors such as aerospace, energy industry or metallurgy. The CEA-LIST develops EC probes to meet the needs of the industry. The design of these probes is based on the use of the simulation platform CIVA.

Classically, defects to be detected are discontinuities in the material. But, as probe response depends on the material electrical conductivity σ [3, 4] and magnetic permeability μ [5, 6, 7, 8], detection of the variation of electromagnetic properties of material can be performed [9]. Such detection can reveal, for example, burned areas or inhomogeneous materials [10, 11], which can lead to poor mechanical properties (for example, weaker hardness).

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One of the targeted applications is the study of the state of the teeth of gearings. These surfaces are not flat. Therefore, the design of a EC flexible pattern etched on kapton substrate is useful to avoid lift-off effects. Such a flexible pattern (paragraph 2) was optimized by simulation to get a straight response in the complex plane and a great sensitivity when the electrical conductivity evolves [1, 2]. Comparisons between experimental and simulation data on conductive materials are presented. Experimental results are also given on ferromagnetic samples. Then (paragraph 3), experimental results on mock-ups with burned areas with different intensities of burning are studied. It is very difficult to control the intensity of burning, so that the available mock-ups are flat in this stage of the study. Finally, a conclusion (paragraph 4) gives perspectives of the study.

2. Description of optimized flexible pattern

We consider a pattern composed of a transmitting coil and a receiving coil. The internal and external diameters, the distance between coils and the frequency are optimized to obtain, on the one hand a straight response in the complex plane for electrical conductivity variations, and on the other hand a great sensitivity to the material conductivity. Geometrical parameters and a picture of the optimized pattern are given in table 1 and in figure 1, respectively. A high frequency, equal to 1 MHz, is chosen, allowing just the material surface to be sensitive.

Table 1. Geometrical parameters of optimized pattern.

Parameters	Receiving coil	Transmitting coil
Outer diameter	10mm	10mm
Inner diameter	9mm	9mm
Height	35μm	35μm
Number of turns	4	4
Lift-off	10μm	60μm
Distance between axis of coils	5mm	



Figure 1. Picture of optimized pattern.

The good performances of the probe were evaluated experimentally (Figure 2) on conductive and non-magnetic standards whose the conductivity evolves between 0.6 MS/m and 59 MS/m (Table 2).

Table 2. List of conductive and non-magnetic materials on which the probe has been evaluated.

Reference	Materials	σ [MS.m ⁻¹]	Reference	Materials	σ [MS.m ⁻¹]
a	Titanium LT31	0.6124	h	Aluminium 2024/T3511	16.93
b	Inconel 600	0.97	i	Aluminium 7175/T7351	22.8
c	Manganin	2.449	j	AlMgSi F32	27.71
d	Maillechort	3.953	k	Aluminium 99.5	34.04
e	Bronze	8.704	l	Copper	57.72
f	Nordic Gold	9.67	m	Copper	58.8
g	Messing	14.35			

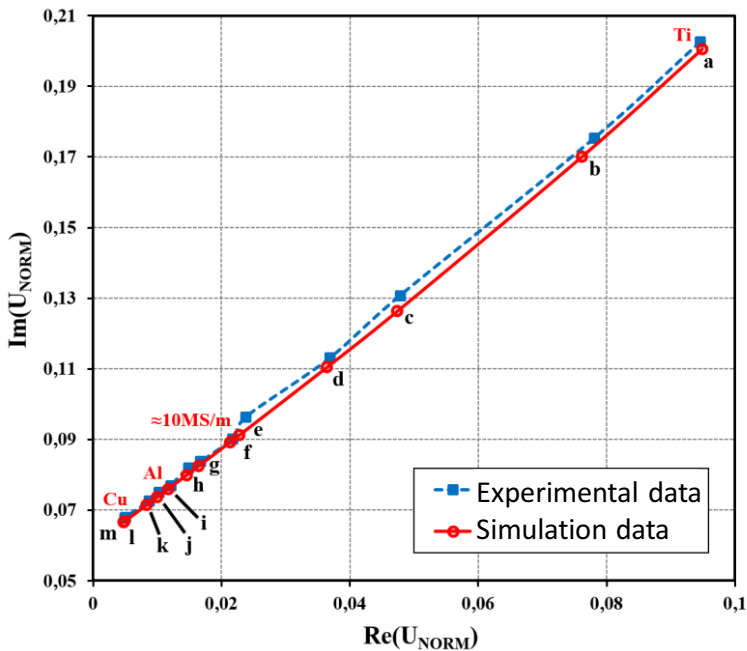


Figure 2. Evolution of EC signal when the electrical conductivity evolves.

We verify all measurement points are located in the vicinity of a same line in the complex plane and we observe a good correlation between experimental and simulation data.

The behavior pattern is now tested on ferromagnetic materials which have been demagnetized thanks to commercial device. First, we use steel samples with different rates of ferrite structure (Table 3 and Figure 3a) and then we use different ferromagnetic samples (Figure 3b).

Table 3. Ferrite rates on different sample

Reference	Ferrite rate	Reference	Ferrite rate
a	0.45 %	f	10.4 %
b	0.51 %	g	31.2 %
c	2 %	h	31.5 %
d	3.1 %	i	33 %
e	10.2 %	j	82.2 %

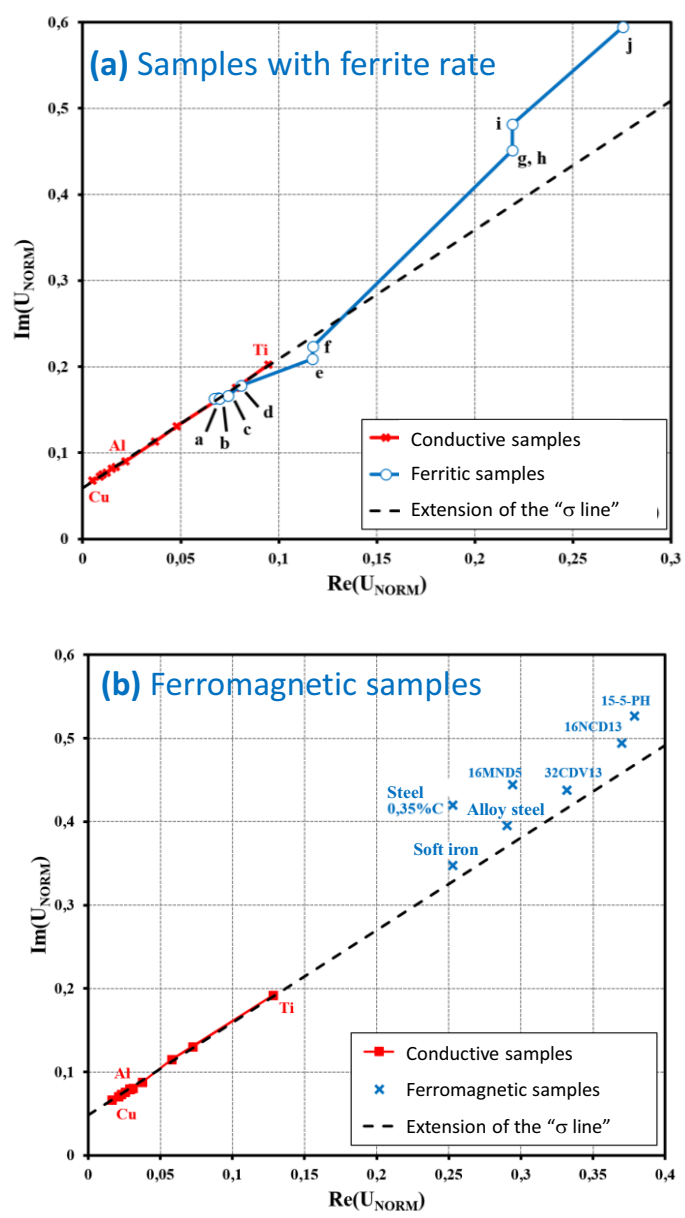


Figure 3. (a) Experimental curves obtained on samples with different ferrite rates. (b) Experimental curves obtained on various ferromagnetic samples.

We note all measurement points are in the vicinity of the same straight line obtained with conductive and non-magnetic material.

Based on this design, many applications can be considered. For example, it is possible to detect defects in a non-homogeneous weld [13]. In a machined ferromagnetic piece, this probe has also been tested to detect variations of conductivity, induced by a local burning due to the machining-tool.

3. Detection and characterization of burned areas

This part presents experimental results obtained on a series of ferromagnetic steel mock-ups (18NiCr5-4) with no burning and burning of different intensities. Figure 4 shows pictures of the mock-ups and Figure 5 an example of EC Cscan obtained on the mock-up with the higher burning. The mock-ups have been demagnetized. However, we have noticed that the residual magnetic field doesn't affect the measurements (for this material).

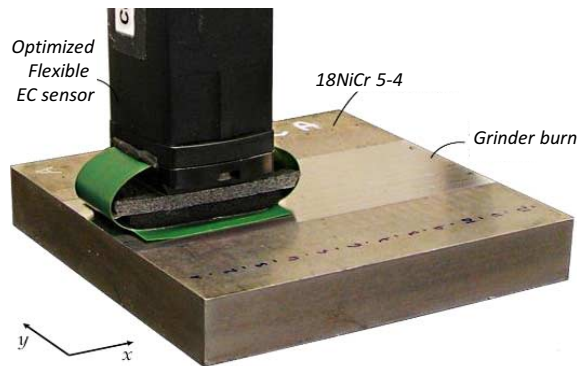


Figure 4. Samples and moving of the probe.

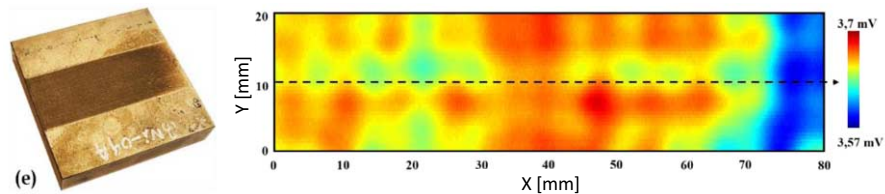


Figure 5. Example of cartographies (amplitude) obtained at 1 MHz at the surface of burned areas with the higher burning.

We note that the burned area is not perfectly homogeneous on the surfaces. Figure 6 gives relative levels of real and imaginary parts of EC signals. The signals are drawn on the dotted line defined in the previous C-Scan.

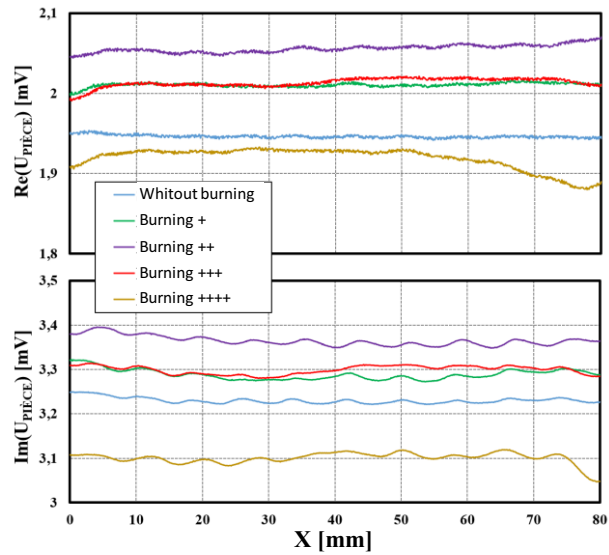


Figure 6. Real and imaginary parts on the previous dotted lines.

To study results as defined in paragraph 2, Figure 7 gives EC signals in the impedance plane, on the previous dotted lines, for all the available mock-ups. We confirm that all the signals are globally on the same straight line. But, for the two higher levels of burning, we observe a non-monotonic behavior. A modification of metallographic properties of materials (tempered steel) can explain the phenomenon from a certain level of burning.

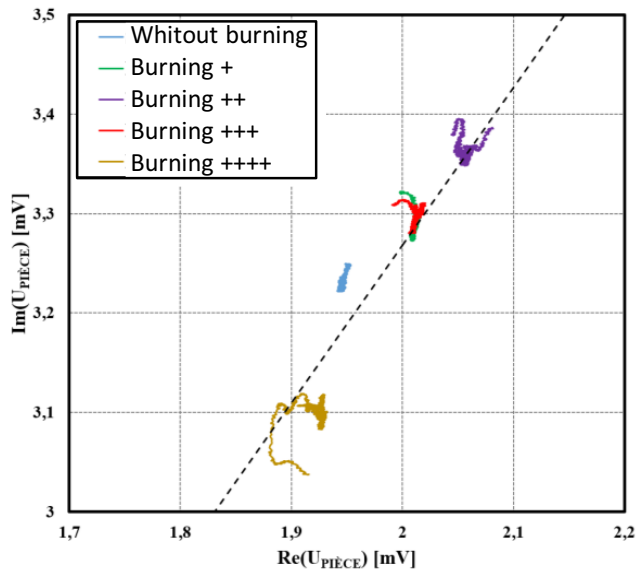


Figure 7. Representation of EC signals in impedance plane for samples with different levels of burning.

If we draw the amplitude of the EC signals for the previous lines in function to hardness HV5 of material, provided by CETIM [11], we obtain the graph given in figure 8.

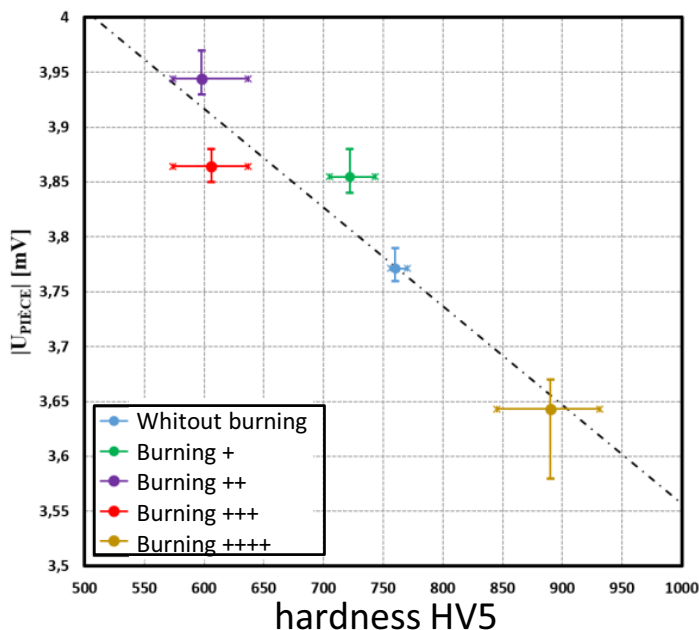


Figure 8. Correlation between the amplitude of EC signals and hardness HV5 for samples with different levels of burning.

We observe a good linear correlation for all the samples between amplitude of EC signals and hardness HV5 as exhibited by the line plotted in Figure 8.

4. Conclusion

It is interesting to note that all the measurement points in the EC signal complex plane are in the vicinity of the same straight line, for conductive and non-magnetic materials, ferromagnetic materials and ferromagnetic material with burned areas.

It is possible to detect and characterize weak burning on materials (temperature in the range of 200-720°C). But, when the burning is too high (>720°C), the relation between the EC signals and the intensity of burning is non monotonic [11]. The metallographic properties of the material is greatly modified and becomes a tempered state. In this state, the amplitude of the EC signal cannot be distinguished from the previous state with the optimized pattern.

It will be possible to couple hardness testing and EC signal to estimate the range of burning, but future works will be dedicated to testing other patterns on these samples in order to estimate more precisely the intensity of the burning.

A second interesting result is we obtain a linear evolution between the amplitude of the EC signal and the hardness HV5. Such sensitivity to hardness can be used for many industrial applications and will be shortly tested in the case of machined surfaces and teeth of gearings whose surfaces are not flat.

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Adaptive Image Reconstruction Algorithm for Electrical Capacitance Tomography

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Abstract. Since the image reconstruction of electrical capacitance tomography (ECT) is an inverse problem of ill-posedness, regularization techniques are widely used to obtain stable solutions. The choice of regularization coefficients has a direct influence on the reconstruction result. In theory, the optimal regularization coefficient exists, but for most regularization techniques, the computation cost is too large to apply in real-time ECT image reconstruction. In this paper, a new approach which can determine the regularization coefficient quickly in iterative Tikhonov regularization algorithm for ECT is proposed. In order to improve the quality of the reconstructed image further, an improved method of adaptive threshold filtering is used in the last Tikhonov iteration. The algorithm proposed is tested by the noise-free and noise-contaminated capacitance data. The simulation results demonstrate that the proposed algorithm is superior to well-known Landweber algorithm in terms of image quality.

Keywords. ECT, reconstruction algorithm, regularization

1. Introduction

Electrical capacitance tomography (ECT) is a PT (process tomography) technology based on capacitive sensing mechanism. It can realize multiphase material visualization measurement in the closed internal piping or equipment. It can be used in many industrial processes such as petroleum, chemical, metallurgy, environmental engineering etc. Its main advantages are low cost, simple structure, non-intrusive measurement and safety [1].

One of the key problems of ECT is to use the appropriate image reconstruction algorithm. At present, the commonly used reconstruction algorithms in ECT include the linear back projection (LBP) algorithm, the Landweber algorithm, the Tikhonov regularization, the Newton Raphson algorithm and so on. The LBP algorithm is simple and its reconstruction speed is fast [2]. LBP algorithm is an image reconstruction algorithm based on “hard field” (that is neglecting the influence of the material distribution on the sensitive field), but the ECT itself has the characteristics of “soft field” (that is a nonlinear relationship between the measurement data and the material distribution and the nonlinear relationship varies with the change of the material distribution), so the LBP algorithm has the low accuracy of the reconstructed image.

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The Landweber algorithm is an iterative method based on negative gradient optimization. It has the characteristics of high accuracy, but for complex flow regimes, the Landweber algorithm need more iterations, which result in slow convergence speed, so it can not be used for real time imaging [3]. Tikhonov algorithm can effectively solve the ill conditioned inverse problem of ECT, and the regularization coefficient directly affects the reconstruction result. If the value of the regularization coefficient is smaller, a good approximation of the original problem can be given, but the influence of the error may cause unstable solution. If the value of regularization coefficient is bigger, the sensitivity to error can be reduced, but the estimate of solution usually deviates from the true solution. At present, there are many algorithms for computing regularization parameter, such as discrepancy principle [4], cross validation method [5], L-curve method [6], and other adaptive method [7-8]. But the computation cost is too large to apply in real-time ECT image reconstruction, so it is mainly selected by experience [9]. Newton Raphson algorithm can approximate to real solution fast. In order to avoid the time-consuming of updating the sensitivity matrix in the Newton Raphson iterative formula, a fixed sensitivity matrix is usually adopted in Newton iterative formula. Since the inverse matrix of transpose of sensitivity matrix multiplying sensitivity matrix usually does not exist, a regularization term is added to the Newton iteration formula and the modified Newton Raphson algorithm is also considered to be iterative Tikhonov regularization algorithm. Therefore, this algorithm also has the problem of regularization coefficient selection.

Based on this starting point, a new method determining the regularization coefficient quickly in iterative Tikhonov regularization algorithm is proposed. In this method, regularization coefficient is determined according to the maximum gray value of the LBP reconstruction image. Because this method is an approximate calculation of the regularization parameter, the reconstructed image inevitably exhibits artifacts. In order to eliminate the influence of artifacts, adaptive threshold filtering is used in the last Tikhonov iteration. The advantage of this approach is to improve the quality of reconstructed image.

2. Working principle of electrical capacitance tomography system

An ECT system generally consists of a capacitance sensor, sensing electronics for data acquisition, and a computer system for image reconstruction. The different distribution of two-phase material in the pipeline will cause the variation of permittivity between the capacitor plates. The capacitance value will be measured and transmitted to the imaging computer by the capacitance value acquisition circuit. Computer reconstructs image according to certain algorithm, so as to realize the material distribution image of the cross section of the pipeline.

The ECT technology mainly includes two parts: the forward problem and the inverse problem. The forward problem of ECT is to obtain capacitance value according to the known material distribution. The relationship between the capacitance and the material distribution is non-linear. In practice, this non-linear problem is simplified to a linear approximation in the form:

$$C = SG \quad (1)$$

where C is normalized capacitance vector, G is normalized permittivity distribution of the material, i.e. the gray level of pixels for visualization, and S is normalized sensitivity matrix.

The sensitivity matrix S is calculated approximately by the method of electric field intensity, see equation (2).

$$S_{i,j}(x,y) = -\frac{1}{U^2} \int_A E_i(x,y) E_j(x,y) dA \quad (2)$$

where U is excitation voltage, A is the area of pixel unit, $E_i(x,y)$ and $E_j(x,y)$ are electric field intensity of the pixel i and j respectively and they are calculated by Comsol Multiphysics 4.2 software.

The inverse problem of ECT is to obtain permittivity distribution of material based on known capacitance value. Generally, the number of capacitance measured C is seriously less than the number of unknown permittivity distribution G , so the image reconstruction of ECT is an inverse problem of ill-posedness. In addition, Because of the soft-field characteristics of ECT and the nonlinear relationship between capacitance measurements and permittivity distribution, the image reconstruction becomes a challenging task. It is almost impossible to find an accurate solution by non-iterative algorithm with a simplified linear model. To improve the image quality, the inverse problem of ECT has to be solved iteratively.

3. Adaptive capacitance tomography image reconstruction algorithm

The adaptive image reconstruction algorithm (AIRA) is proposed in this paper. Iterative Tikhonov regularization algorithm is employed to reconstruct image and regularization coefficient in iterative Tikhonov regularization algorithm is determined according to the maximum gray value of the LBP reconstruction image. In order to accelerate the convergence rate, the image reconstructed by the LBP algorithm is used as the initial value of iterative Tikhonov regularization algorithm and filtering projection operation is used in each Tikhonov iteration process. Specially in the last iteration, adaptive filtering threshold is calculated for higher image quality. The AIRA formula is as follows.

$$\begin{cases} g_0 = S^T C \\ g_{k+1} = P_1[g_k - (S^T S + \mu I)^{-1} S^T (S g_k - C)] , 0 \leq k \leq 8 \\ g_{k+1} = P_2[g_k - (S^T S + \mu I)^{-1} S^T (S g_k - C)] , k = 9 \end{cases} \quad (3)$$

$$P_1(x) = \begin{cases} 0 & \text{if } x \leq 0 \\ x & \text{if } 0 < x < 1 \\ 1 & \text{if } x \geq 1 \end{cases} \quad (4)$$

$$\left\{ \begin{array}{l} \mu = a \text{ and } P_2(x) = \begin{cases} 0, & \text{if } x < 0.5\eta_k \\ x, & \text{if } 0.5\eta_k \leq x \leq 1, \\ 1, & \text{if } x > 1 \end{cases} & \text{if } g_{0\max} = \max(g_0) \geq q \\ \mu = b \text{ and } P_2(x) = \begin{cases} 0, & \text{if } x < \eta_k \\ x, & \text{if } \eta_k \leq x \leq 1, \\ 1, & \text{if } x > 1 \end{cases} & \text{if } g_{0\max} = \max(g_0) < q \end{array} \right. \quad (5)$$

where S^T is the normalized transpose of the sensitivity matrix, I is an identity matrix, g_0 is the image reconstructed by the LBP algorithm, g_k and g_{k+1} are the k th and the $(k+1)$ th reconstructed image by the iterative Tikhonov regularization algorithm, μ is the regularization coefficient which must be positive, $g_{0\max}$ is the maximum gray value of the LBP reconstruction image, $\max(\cdot)$ denotes an maximum operator, P_1 and P_2 are the projection operators, k is iteration time, η_k is the filtering threshold value in the k th iteration.

In this paper, there are two main tasks: determination of regularization coefficient μ in the iterative Tikhonov algorithm and calculation of adaptive filtering threshold η_k .

- Determination of regularization coefficient μ in the iterative Tikhonov

The regularization coefficient μ in the iterative Tikhonov algorithm formula is determined adaptively according to the maximum gray value of LBP reconstruction image. Because the sensitivity of the pipeline edge is larger than the sensitivity of the pipeline central region, the imaging quality of material distributed near the edge of pipeline is seriously affected by noise. When the more material is distributed near the edge of pipeline, a bigger regularization coefficient should be needed to resist noise. Moreover, it is found that the more material distributed near the edge of pipeline, the bigger maximum gray value of LBP reconstruction image through a large number of simulations. Therefore, the bigger maximum gray value of LBP reconstruction image, the bigger regularization coefficient needed. When the maximum gray value of LBP reconstruction image is bigger than a value q or equal to q , the value of regularization coefficient is selected to the value a . While the maximum gray value of LBP reconstruction result is less than q , the value of regularization coefficient is selected to b , see equation (5). q is 0.35, a is 0.2 and b is 0.002 are suitable for image reconstruction in ECT by simulation.

- Calculation of adaptive filtering threshold η_k

The filtering threshold value η_k is calculated as document [10]:

$$\eta_k = \frac{\bar{C}_l}{\bar{C}_h} \cdot (1 - \bar{C}_m) \cdot g_{mk} \quad (6)$$

where $\bar{C}_l$ is the average value of the capacitance values while the sensor is filled with low permittivity material; $\bar{C}_h$ is the average value of the capacitance values while the

sensor is filled with high permittivity material; $\bar{C}_m$ is the average value of the normalized capacitance measurement values resulting from permittivity distribution to be reconstructed; and g_{mk} is calculated as follows:

$$\begin{cases} g_{mk} = AVG(g_k(i), \{g_k(i) > mk\}) \\ mk = MED(g_k(i), \{g_k(i) \neq 0\}) \end{cases} \quad (7)$$

where $AVG(\cdot)$ denotes average operator; $MED(\cdot)$ denotes median operator; $\{\cdot\}$ denotes a condition statement; $g_k(i)$ denotes i th element of vector g_k .

On the basis of document [10], an improved adaptive threshold filtering is proposed and is only used in the last iteration of iterative Tikhonov. The adaptive filtering threshold is not only related to the material distribution, but also related to the maximum gray value of LBP reconstruction image, see equation (5). If the regularization coefficient selected is bigger, that is the maximum gray value of the LBP reconstruction image is bigger, the ability of imaging to resist noise is enhanced, so filtering threshold should be smaller so as not to filter out useful information.

4. Algorithm test

4.1 Sensor model

The circular sensor is used in this paper. The imaged area in the pipeline is divided into 673 elements, and the number of capacitor plates is 8. The arrangement of measurement electrodes of an eight-electrode ECT sensor is shown as figure 1. The low (background) and the high (object) relative permittivity of materials are 1.0 and 6.0 respectively in the simulation.

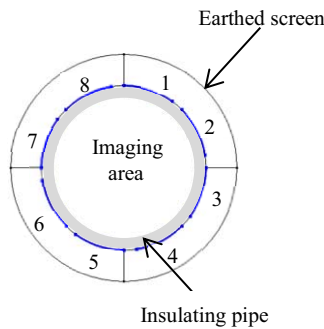


Figure 1. Arrangement of measurement electrodes of an eight-electrode ECT sensor

4.2 Evaluation criteria and three algorithms comparison

In order to evaluate the reconstruction performance quantitatively, correlation coefficient is used as the evaluation criteria. The bigger the correlation coefficient, the

better the quality of reconstructed images. Its definition is as followed [11]:

$$cc = \frac{\sum_{i=1}^N (\hat{g}_i - \bar{\hat{g}})(g_i - \bar{g})}{\sqrt{\sum_{i=1}^N (\hat{g}_i - \bar{\hat{g}})^2 \sum_{i=1}^N (g_i - \bar{g})^2}}$$

(8)

where g is the true permittivity distribution of the tested object; $\hat{g}$ is the reconstructed distribution; N is the number of elements divided in the imaged area in the pipeline , $\bar{g}$ and $\bar{\hat{g}}$ are the mean values of g and $\hat{g}$, respectively.

6 flow regimes are simulated in this paper when capacitance data are contaminated with noise of level 0% (i.e. noise-free) and 5% in this paper. A capacitance vector C_n contaminated by noise is generated by adding a random noise to a normalized capacitance vector C resulting from permittivity distribution to be reconstructed. The relative norm error e between C_n and C is defined as [12] and the value of e is called as the noise level.

$$e = \|C_n - C\| / \|C\|$$

(9)

where $\|\cdot\|$ denotes 2-norm.

Three reconstruction algorithms including the LBP, the Landweber and the AIRA are used. The Landweber algorithm is iterated 100 times. The AIRA is iterated 10 times. When capacitance datas are contaminated with noise of level 0% and 5% , reconstructed images are shown as figure 2, and correlation coefficients are calculated respectively in table 1 and table 2.

Table 1. Correlation coefficient (cc) of three reconstruction algorithms under the noise level of 0%

	LBP	Landweber	AIRA
(a)Five bubbles	0.2073	0.5843	0.8287
(b)Three bubbles	0.3961	0.7376	0.8254
(c)Annular flow	0.9378	0.7989	0.9063
(d)Stratified flow	0.9066	0.7516	0.7717
(e)Core flow	0.8668	0.8841	0.8978
(f)One bubble and Stratified flow	0.7869	0.6787	0.7217

Table 2. Correlation coefficient of three reconstruction algorithms under the noise level of 5%

	LBP	Landweber	AIRA
(a)Five bubbles	0.1944	0.5842	0.7956
(b)Three bubbles	0.3906	0.7610	0.8037
(c)Annular flow	0.9356	0.8120	0.9010
(d)Stratified flow	0.8803	0.7508	0.7560
(e)Core flow	0.8701	0.8498	0.8904
(f)One bubble and Stratified flow	0.7739	0.6753	0.7031

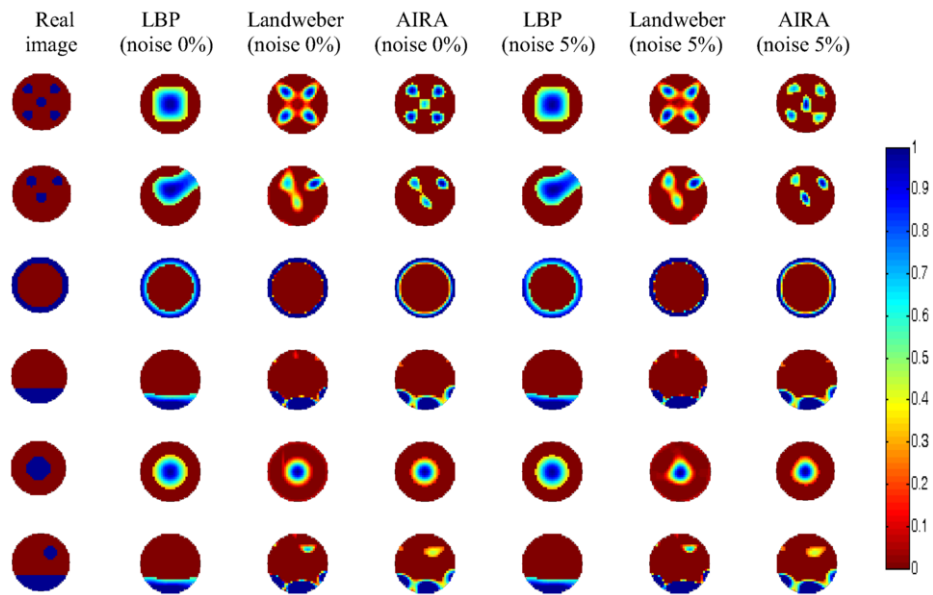


Figure 2. Real images and the reconstructed images

The simulation results show that LBP algorithm is applicable only to some simple flow regimes, such as annular flow and stratified flow. For the complicated flow regimes, such as five bubbles and three bubbles, LBP reconstruction results are very poor. The Landweber algorithm can obtain better quality images than LBP. But for complex flow patterns, such as 5 bubbles in Figure 2, the central region of 5 bubbles does not image at all by the Landweber algorithm. The algorithm proposed in this paper can obtain higher quality image than Landweber algorithm for 6 flow regimes selected in this paper. Table 1 and table 2 indicate that the correlation coefficients of the proposed algorithm in this paper are bigger than the Landweber algorithm, and bigger than 0.7 when capacitance data are contaminated with the noise level of 0% and 5%.

5. Conclusion

The adaptive image reconstruction algorithm(AIRA) which can determine the regularization coefficient in iterative Tikhonov regularization algorithm quickly and calculate the adaptive filtering threshold in the last iteration of Tikhonov iterative is proposed in this paper. The regularization coefficient and the adaptive filtering threshold are determined according to the maximum gray value of the LBP reconstruction image. The algorithm is tested by the noise-free capacitance data and the noise contaminated capacitance data. The correlation coefficients of the algorithm proposed in this paper are bigger than the Landweber algorithm, which show that the accuracy of the proposed algorithm is superior to the Landweber algorithm.

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A New MFL Testing Sensor Based on Localized Magnetization for Micro Crack Detection

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Abstract: Magnetic flux leakage (MFL) testing method has been used in defect detecting for decades. However, the capability of micro defect inspection mainly depends on the characterization of the sensor. This paper proposes a new MFL testing sensor based on localized magnetization for micro defect detection, which can detect micro crack with high sensitivity and accuracy. Preliminary introduction for the structures and working principles are presented. By the subsequent comparison of simulations and experiments between common MFL method and the new MFL sensor, we can conclude that this sensor possesses the capability of micro defect inspection. Finally, both the simulation results and the detection signals in the experiments demonstrated its feasibility and practicality.

Keywords: MFL, Sensor, Localized magnetization, Micro crack

1. Introduction

Magnetic flux leakage (MFL) testing method has been widely used in the field of nondestructive testing (NDT). As one of the most prevalent defect detection techniques, large amount of studies have been reported since the MFL method was proposed the first time. Likewise, surface quality of a part is also the vital factor for its effective operation and lifetime especially in the precision engineering of machining and manufacturing where the surface characterizations are usually influenced by the mechanical actions and thermal features. Therefore, precise inspection is of great significance to ensure safety and manufacturing efficiency. In other words, the detection precision or accuracy mainly depends on the characterizations of the inspection sensors. Although the traditional MFL testing sensors may be used for conventional defect detection [1, 2], most of them fail to inspect micro defect with a relative high sensitivity and accuracy. Such as the tri-axial sensor technology which was able to detect corrosion, mechanical defects proposed in [3], but incapable of local micro defect inspection. Despite the improvement of sensor system presented in [4] may increase the sensitivity of defect detection, the signals were very weak and the processing difficulty was enhanced as well. Additionally, researches of [5-8] have been occupied in the weak signal processing techniques when detecting

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micro defects, thus making the online micro defect inspection become more difficult and the inspection efficiency is thereafter decreased. Other methods such as barkhausen noise (BN) measurement in [9], which is mainly used in the measurement of corrective and residual magnetic field. However, the signal processing difficulty is also increased. Besides, the micro defect in the so called collinear wave mixing technique in the study of [10] resembles the common defect with bigger size, thus the detection feasibility for smaller micro defect cannot be well demonstrated. Ultrasound infrared thermography (UIRT) methods [11] may also be used in micro defect detection which has a merit to inspect a wider range in a short time with no contact with the target object, but the testing results often depend on the variable frequency, as well as the proposed enhanced acoustic emission detection method in [12]. The magnetic memory method [13, 14] is a NDT method that can detect the microscopic damage on the basis of the magnetostriction and magnetoelasticity effect. However, it couldn't avoid the influence of the residual stress and interference of other physical characterizations in the tested object. By the way, the E-shaped magnetic sensor or magnetic flaw detection apparatus in patent [15] maybe capable of improving the detection sensitivity while may still failed in micro defect inspection.

Therefore, a new MFL testing sensor based on local micro magnetization is proposed here. The sensor structures with tiny magnetic sensing bottom make the new sensor become more sensitive to micro defects. Consequently, the localized magnetization nondestructive evaluation method could provide significant information about the surface features of the tested object. The proposed micro defect detection method based on localized magnetization is introduced through a description of the sensor structures and principles. Then the experiments and comparisons results between the new MFL testing sensors and common MFL methods manifest that the new MFL testing sensor is predominant in micro crack detection. Finally, the detection accuracy and precision as well as the sensitivity are enhanced with a remarkable signal to noise ratio.

2. New sensor structures and principles

A new local micro magnetization sensor is introduced in this section. Specifically, the structures and principles are separately explained.

2.1. New sensor structures

The new sensor mainly consists of three modules, namely, MFL signal capture unit, magnetic sensing part and local micro magnetization module. As illustrated in Fig.1. The main components of the new proposed sensor are magneto-sensitive element, insulating sleeve, search coil, annular magnetic yoke, permanent magnets, magnetic guide core with high permeability and tiny sensing bottom, magnetic conducted object.

Concretely, the conical magnetic guide core of 3 or 10 with high magnetic permeability of 200 mH/m such as permalloy is connected with the magneto-sensitivity element of 2 such as hall element electrically. The height and diameter of the overall sensor are about 20mm and 10 mm, respectively. The remanence of each permanent magnet is approximate to 1.12 T, and the coercivity and (BH)_{max} are 780 kA/m and 223kJ/m.

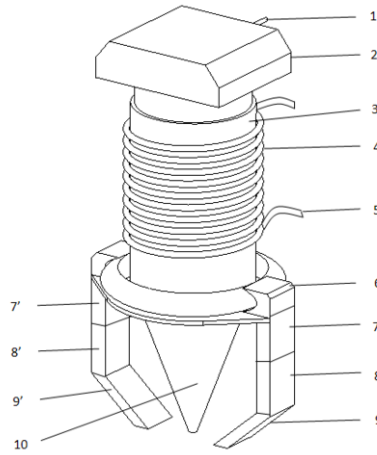


Figure 1. Structures and components of the new sensor; 1. signal output end; 2. magneto-sensitive element, 3. insulating sleeve; 4. search coil; 5. output end of the coil; 6. annular magnetic yoke; 7. one pole of permanent magnet; 8. another pole of permanent magnet; 9. magnetic guide part with high permeability; 10. tiny sensing bottom of the magnetic guide core.

The insulating sleeve is used to isolate the guide core and the search coil which is made of copper enamelled wire. The diameter and electrical resistivity of the coil wire are about 0.1mm and 0.0172 $\Omega\cdot\text{m}$, and the number of turns could be 100 or more. Furthermore, the excitation sources of magnetic yoke of 6 and the permanent magnet of 7 (or 7') and 8 (or 8') are symmetrically assembled, as well as the magnetic guide parts of 9 and 9'.

2.2. Working principles

When detecting defect with the new proposed sensor, the schematic diagram is illustrated in Fig. 2 and as follows.

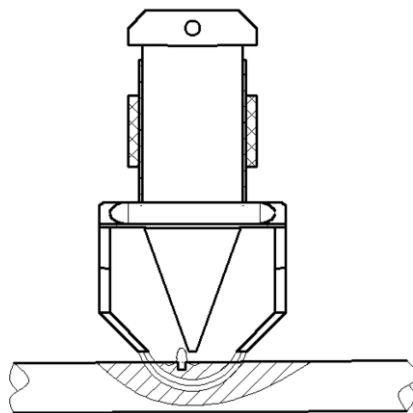


Figure 2. Working principles of the new sensor

As expressed in Fig. 2, the process of micro defect detection is briefly explained as follows. When the magnetic guide core is approaching or scanning the tested specimen

in a certain speed, corresponding small region of the tested object under the bottom of the tiny end of the guide core will be locally magnetized. Therefore, if any discontinuity or defect is encountered, the MFL produced by the micro defect would be caught by the tiny bottom of guide core and picked up by the search coil or the magnetic-sensitive element. Then the MFL signals are generated by the search coil and the upper magneto-sensitive unit. After the signal data acquisition and the analog to digital converting as well as signal conditioning, the final output signals are obtained. Owing to that the magnetization region in the bottom of guide core is super small and could reach 0.01mm in radius, the magnetization area in the tested object is thereafter reduced and the magnetization field is focused as well. Thus, micro defect with a width or length of 0.1mm or smaller could be well detected. Finally, the local micro magnetization and defect detection are achieved. Additionally, the scanning speed of the sensor could be various and has an effect on the testing signals especially when the magnetic-sensitive element is the coil. According to the law of the electromagnetic induction, the signal amplitude is influenced by the gradient of the magnetic flux which is related to the scanning speed of the sensor. Generally, as the scanning speed of the sensor increases, the magnitudes of the three components of MFL signals are nonlinear monotone increasing and the increasing gradient of B_x and B_z is greater than B_y according to related researches [16, 17]. However, the scanning speed of the sensor is uninfluential to the detection signals when the magnetic-sensitive element is hall or other elements.

3. Comparison of simulations and experiments

To reveal the characterizations and advantages of the new proposed sensor further, comparisons by simulations and experiments were separately conducted as follows.

As one of the most frequently used NDT techniques, the MFL produced by the excitation of permanent magnet and magnetic yoke is expressed in Fig. 3, where the defect is 0.1×0.1 mm in width and depth in the two dimensional (2-D) simulations. Observed by the color map of magnetic density of H , what could be noticed is that the maximum density is approximate to 691.89 (A/m).

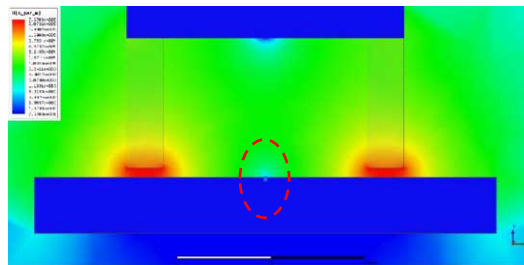


Figure 3. Simulation of the common MFL detecting method

Similarly, the MFL generated by the localized magnetization micro crack on the specimen model and corresponding magnetic density cloud chart are exhibited in Fig. 4. Obviously, the magnetic strength is relatively stronger compared with conventional magnetizing sources. According to the numerical simulations, the MFL strength of the defect could reach 1.51×10^4 (A/m) under the same circumstance conditions. Finally, the distinct MFL characterization indicates that the detection accuracy and sensitivity of micro defects are vastly improved by the proposed localized magnetization sensor, which

has great potential in the future application in defect inspection of ferromagnetic materials and surface topography compared with the common MFL testing method.

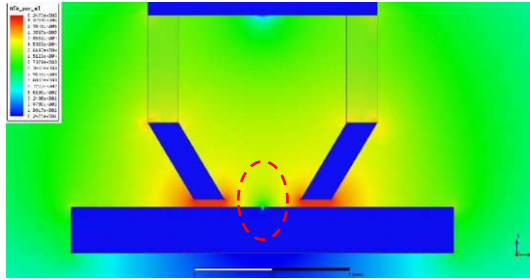


Figure 4. Simulation of the new MFL sensor

Besides, corresponding experiments by different micro cracks are also conducted. Specifically, the length, width and height of the specimen are 40mm, 10mm and 6mm, respectively. As for the dimensions of these defects, all the length $\times$ depth are 2mm $\times$ 0.1mm, while the widths are various from 0.05mm to 0.8 mm, namely, 0.05mm, 0.1mm, 0.3mm, 0.5mm and 0.8mm. the displacement of the sensor from the surface of the tested specimen is 0.1 to 2 mm.

The experiments results are obtained separately by the new proposed sensor and common MFL method which uses the typical and classical permanent magnetic yoke method and the hall element. When the specimen with defects are scanned by the two sensors, corresponding signals are acquired and, as shown in Fig.5 to Fig.9. Explanatorily, the signals are all acquired by the repeating scanning of the sensors in two opposite directions above the defect. Hence, there should be several signal peaks with different shapes when the defect is scanned or encountered iteratively. Additionally, owing to the differences in scanning speeds of the detection sensors, the signal peak values are also various.

When the defect is 0.8 mm in width and repeatedly scanned, the signals are expressed in Fig. 5 which are respectively acquired by the new sensor based on localized magnetization and the common MFL method for comparison purposes.

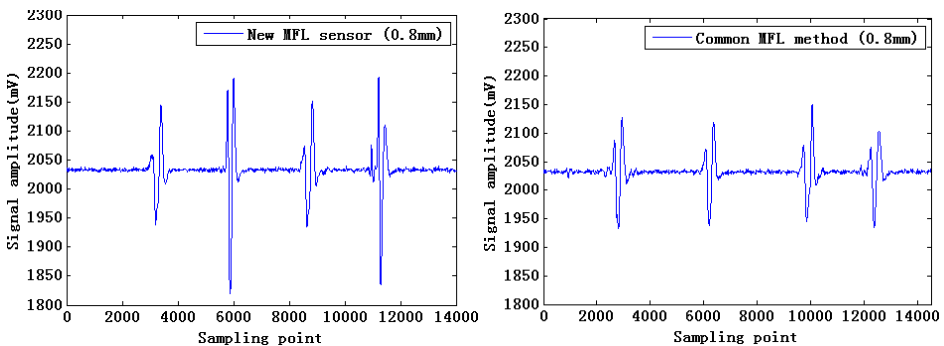


Figure 5. Experiment results of detection signals when the defect is 0.8mm in width

It's clear that the signals from the new sensor are stronger than that of the classical MFL method in amplitudes. What's more, owing to the electromagnetic induction effect, the signals from the search coil of the new sensor maybe further enhanced when the scanning speed is increased in a certain range. Similarly, when the defect with width of

0.5mm is encountered and detected by the scanning sensors, corresponding experiment results are presented in Fig. 6, as expressed below.

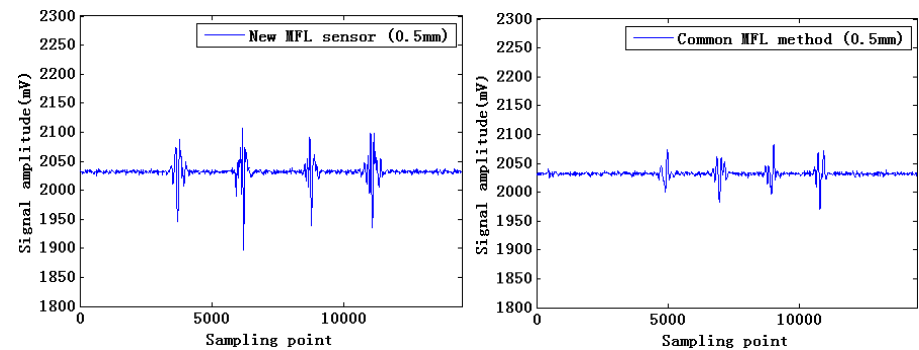


Figure 6. Experiment results of detection signals when the defect is 0.5mm in width

Observed by the signals from Fig. 6, we can draw the conclusion that even the signals are all decreased in amplitude compared with signals from Fig.5, signals of the new sensor are still stronger than that from the common MFL method. That further demonstrates the effectiveness and feasibility of the new method. By the same token, when the defect widths are 0.3mm, 0.1mm and 0.05mm, the experiment results of the comparison between the new sensor and the classical MFL method are separately shown in Fig. 7 to Fig. 9.

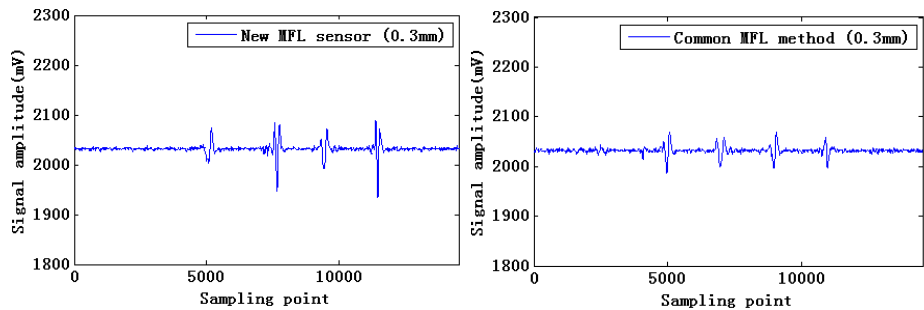


Figure 7. Experiment results of detection signals when the defect is 0.3mm in width

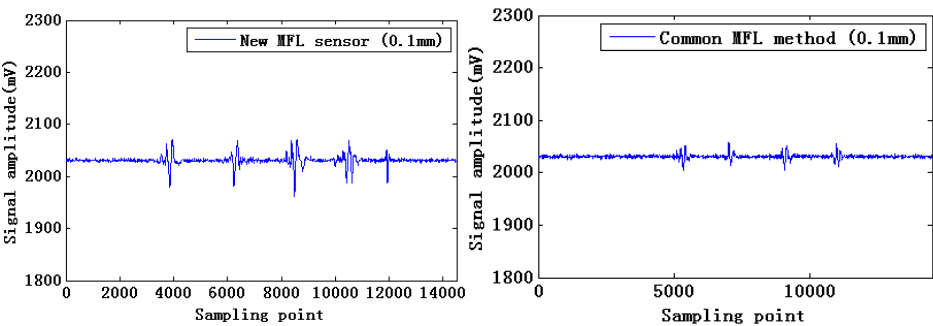


Figure 8. Experiment results of detection signals when the defect is 0.1mm in width

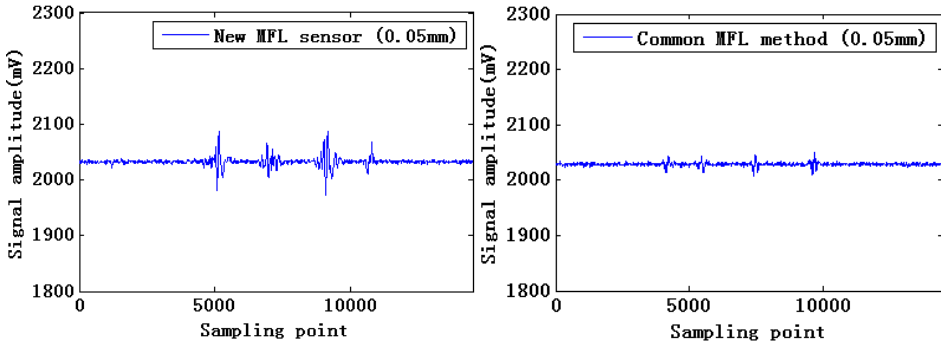


Figure 9. Experiment results of detection signals when the defect is 0.05mm in width

From the experiment results of signal waveforms presented above, we can find that although the signals are decreased slowly in amplitude as the defect width becomes smaller, the signals from the new sensor are always bigger than that from the common MFL method, especially when the defect is as small as 0.05mm in width. Particularly, the sensitivity of the new sensor for the crack width detection could reach 300 mV/mm while the old common MFL sensor is approximate to 100 mV/mm. Consequently, the experiment results not only manifest that the capability of defect detection for the new sensor is more advantageous than the latter, but also demonstrate that the new testing sensor based on local micro magnetization possesses better characterization in micro defect inspection. As the processing and manufacturing technology progresses, the detection precision could be further improved in the future.

4. Conclusion and discussion

As the MFL testing methods further developed, the requirement for micro defect detection is needed and it is necessary especially in the precision engineering of machining and manufacturing. Therefore, this paper proposed a new MFL testing sensor based on localized magnetization. Furthermore, the subsequent simulations and experiments both demonstrated the advantageous of the localized magnetization method based on the new sensor compared with the classical MFL method. Finally, we can draw a preliminary conclusion that the new sensor is more effective in defect detection and possesses a better micro crack inspection capability, which is of great significance and promise to the future practical application.

However, some factors such as the motion effect of the new sensor, the difference with the field obtained in the absence of a defect, the sensitivity of the receiver device as a function of the amplitude, as well as the orientation of the field may influence the inspection results. Therefore, the new MFL sensor is more suitable in detecting the surface of a part in precision engineering. Naturally, a relative better inspection circumstance could be guaranteed.

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Reconstruction of Stress Corrosion Cracking Based on a Regularization Method Using Multi-Frequency Eddy Current Signals

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Abstract. Stress corrosion cracking (SCC) is a prolong breakdown phenomena that occurs in several metal materials and may cause catastrophic accidents. Quantitative nondestructive testing of SCC is important in guaranteeing safety and highly efficient operation in large mechanical systems. However, reconstruction of SCC using eddy current testing (ECT) signals remains a problem given the complicated microstructure of SCC and the ill-posedness of the inverse problem. In this study, the relation between crack parameters and multi-frequency ECT signals is investigated; a regularization method that uses multi-frequency ECT signals is used to reduce the ill-posedness of SCC reconstruction. Conclusions are verified with simulated ECT signals from different conductive crack models.

Keywords. Inverse problem, reconstruction, regularization method, eddy current testing, stress corrosion cracking

1. Introduction

Quantitative evaluation of the defect is one of the targets for eddy current testing (ECT). It estimates the defect type, location and size from the measured signals [1-2]. Depth is frequently underestimated in reconstruction of stress corrosion cracking (SCC) by using ECT signals due to the complicated conductive properties of SCC and ill-posedness of the inverse problem [3]. On the problem of SCC reconstruction, it is found that SCC in different environment and materials has different crack conductive properties [4-5]. The electron discharge machining crack models with multiple nonzero conductivity values have been used to numerical simulation of SCC [6-11]. However, as the number of crack conductivity parameters increases, it is more difficult to accurately reconstruct the SCC profile; that is, the ill-posedness of the inverse problem increases. One of the reasons of the ill-posedness for the ECT inverse problem is different defect parameters maybe correspond to similar values for some ECT signals. Therefore, many numerical criteria are added to the source structure or solution space in practical work, thereby identifying a reasonable solution in a certain range or sense.

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Based on this point, a regularization method using features of multi-frequency ECT signals is used to SCC reconstruction, thereby reducing the ill-posedness from the complicated conductive property in the crack zone and the inverse problem itself in this study.

2. Crack parameters and ECT signals

A series of numerical simulations have been implemented based on FEM–BEM hybrid method to discuss the relations between crack parameters and ECT signals [12]. A plane numerical model of SCC is adopted in this study in view that crack width variation can be nearly equivalent to crack conductivity change for single-frequency ECT signals [13-14]; in this numerical model, width is fixed to a certain value, and crack-shaped parameters only comprise crack depth and length.

The ECT signals are investigated when the crack parameters, including crack length, depth and conductivity, vary. In this study, a plus-point coil (inner side length 4.5 mm, outer side 5.5 mm and height 2.0 mm) are applied as the inspection probe. The detection object is a SUS304 plate that is 10 mm in thickness, 200 mm in length, 100 mm in width and 1.4 MS/m in conductivity. For single-frequency ECT signals used in Figure 1–3, the excitation frequency is 30 kHz and the liftoff is 1.0 mm.

Figure 1 shows that the crack length mainly influences the start and end positions of parallel scanning ECT signals, while it influences the amplitude and phase of these signals to a certain extent.

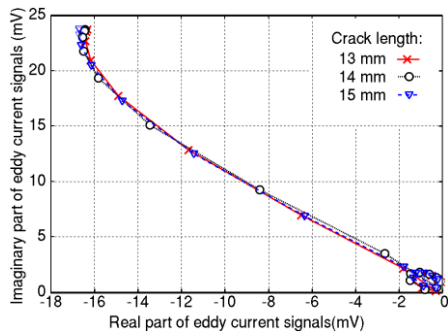


Figure 1. Parallel scanning ECT signals from conductive crack models with different lengths (Crack parameters: depth 3 mm, width 0.4 mm, relative conductivity 3%.)

Figures 2 and 3 demonstrate that the values and distribution of crack conductivity in the depth direction of crack affect the ECT signals in a manner that is similar to the effect of crack depth; this effect is sustained on the amplitude and phase of the ECT signals. Therefore, reconstructing SCC profiles is an ill-posed inverse problem, which is aggravated given the complicated conductivity of SCC in the crack zone. In general, multiple local optimal solutions of crack shape parameters may appear if ECT signals of a single excitation frequency are used when crack shape parameters and conductivity are reconstructed simultaneously.

Certain differences due to various penetration depths occur in the ECT testing for different excitation frequencies when the local conductivity and shape parameters of the crack vary. In particular, high-frequency ECT signals mainly embody information from the crack surface, especially from crack conductivity and length. Low-frequency

ECT signals can express information from the crack depth and crack conductivity from the bottom, as displayed in Figures 4 and 5. Consequently, the features of multi-frequency ECT signals can be used to reconstruct the crack conductivity distribution and depth, thereby solving multiple extreme value problems caused by the complicated conductivity distribution of the SCC to a certain extent.

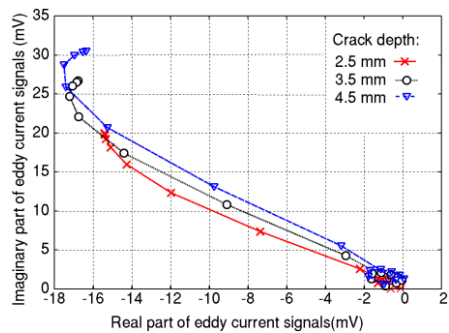


Figure 2. ECT signals from conductive crack models with different depths (Crack parameters: length 14 mm, width 0.4 mm, relative conductivity 3%.)

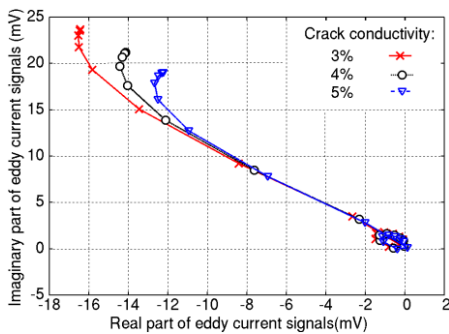


Figure. 3. ECT signals from crack models with different crack conductivities (Crack parameters: length 14 mm, depth 3 mm, width 0.4 mm)

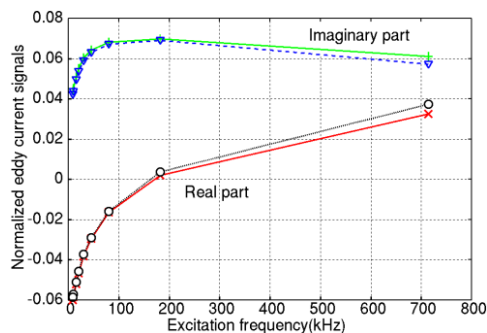


Figure 4. Normalized ECT signals of a scanning point from crack models with the same depth but different conductivity (Solid line is the target signals; the dotted lines with circles or triangles are contrast signals.)

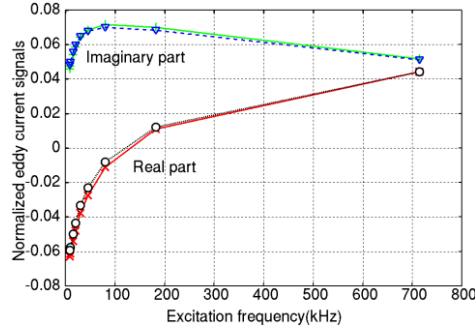


Figure 5. Normalized ECT signals of a scanning point from crack models with the same conductivity but different crack depth (Solid line is the target signals; the dotted lines with circles or triangles are the contrast.)

3. A regularization method using multi-frequency ECT signals

Based on the relation between the crack parameters and features of ECT signals, a regularization method using multi-frequency ECT signals is proposed. The regularization method of crack reconstruction using multi-frequency ECT signals is defined as a series of minimum problems of the mean-squared residual:

$$\min \mathcal{A}(c, \sigma) = \sum_i \left| Z_{i,j}(c, \sigma) - Z_{i,j}^{obs} \right|^2, \quad (1)$$

where $Z_{i,j}(c, \sigma)$ and $Z_{i,j}^{obs}$ are the calculated and measured signals, respectively. The subscript i represents the i th scanning point, and j represents the j th excitation frequency. Vector c and σ are the crack shape and crack conductivity parameters, correspondingly; vector σ is generally nonzero and needs to be nearly expressed as multiple parameters for the SCC. In general, solving the problem directly is difficult. Alternatively, an optimal solution is gradually approximated by a model-based iterative method; that is, the values of the objective functions will be calculated repeatedly with different crack parameters [15]. However, if the number of iterations or mean-squared error range for optimization (1) is restricted, then several extreme value points may appear and may sometimes be far from the truth; that is, it is usually ill-posed. Moreover, the ill-posedness is aggravated or even does not converge as the number of variables increases.

Thus, the minimal conditions of Formula (1) can be relaxed, and the set of variables that satisfy $\mathcal{A}(c, \sigma) = \sum_i \left| Z_{i,j}(c, \sigma) - Z_{i,j}^{obs} \right|^2 \leq C_j$ is searched, where C_j is the given error bound corresponding to the excitation frequency j . The conductivity value that is near the true value is gradually determined from the crack surface to the bottom by using an excitation frequency from high to low.

The SCC parameters that need to be reconstructed include the crack length, depth, and equivalent conductivity distribution in the crack depth direction. This issue is a high-dimensional optimization problem with multiple parameters. A genetic algorithm is applied to solve the optimization problem.

The basic ideas of the regularization method that uses multi-frequency ECT signals are as follows. First, we aim to find the solution space that satisfies a certain error range for problem (1) and a high excitation frequency. Second, the reconstructed crack length and conductivity values that correspond to the upper layer crack are considered a priori information and thus added to the second optimization problem as the constraints. The current optimal solutions of the second optimization problem are obtained after a number of generations of the genetic algorithm; these solutions correspond to signals of the second excitation frequency. Similarly, the crack length and conductivity results in the second step are considered a priori information and thus added to the next optimization problem as the constraints. Finally, we repeat the same procedure until the termination criterion is satisfied.

The crack conductivity distribution in the depth direction is determined gradually by using the ECT signals of the different excitation frequencies from high to low. The solution scope is reduced by solving the constrained optimization problem successively.

4. Numerical results and discussion

The reconstruction of SCC is performed by considering simulated multi-frequency ECT signals from a 10 mm-thick SUS304 plate as the target signals to validate the proposed strategy. Based on the method described in Section 3, a numerical code that uses a genetic algorithm and a fast-forward solver based on the FEM–BEM hybrid method is developed [16]. For the fast-forward solver, databases of unflawed potentials were established for a search region of 18 mm in length, 5.5 mm in depth and 0.4 mm in width. The search region was subdivided into 198 ($18 \times 11 \times 1$) cells. The target signals are from a conductive elliptical crack model of 14 mm length and 5 mm depth; the crack relative conductivity is 2%, 5%, 8%, 10% and 15% from the surface to bottom of the inspection specimen. Each conductivity value corresponds to thickness 1 mm of the inspection specimen.

To simplify the inversion problem, we only use four parameters to approximately describe the crack conductivity distribution in the reconstruction of SCC. The multi-frequency ECT signals, which correspond to the excitation frequencies of 714, 182, 80, and 30 kHz based on the skin depth of the ECT signals, are used in the objective function, as expressed in Formula (1).

The implementation procedure of SCC reconstruction based on the regularization method and multi-frequency ECT signals is as follows:

First, the signals of a 714 kHz excitation frequency are used in the first objective function $\min \mathcal{J}(c, \sigma) = \sum_i |Z_{i,1}(c, \sigma) - Z_{i,1}^{obs}|^2$. A genetic algorithm is used to predict the

crack profiles and conductivity distribution. The reconstructed crack length is 14.1 mm, which is from -7.6 mm to 6.5 mm in the crack length direction, after 200 iterations. The conductivity parameters are predicted as 2.51%, 3.22%, 14.43% and 22.24%.

Second, the predicted crack length 14.1 mm and the first conductivity parameter range $|\sigma_2 - 2.51\%| \leq \varepsilon_{11}$ as the constraint conditions are added to the second optimization problem $\min \mathcal{J}(c, \sigma) = \sum_i |Z_{i,2}(c, \sigma) - Z_{i,2}^{obs}|^2$, which corresponds to the signals of a 182 kHz excitation frequency, in which ε_{11} is the given error bound.

Similarly, the genetic algorithm is used to solve the constrained optimization problem. The optimal results of the crack conductivity parameters are 2.14%, 4.71%, 7.13% and 23.18%, and the crack depth is 4.5 mm after 300 generations.

Similarly, the crack length 14.1 mm, $|\sigma_1 - 2.14\%| \leq \varepsilon_{21}$ and $|\sigma_2 - 4.71\%| \leq \varepsilon_{22}$ as the restrained conditions are used for the third objective function which corresponds to the signals of an 80 kHz excitation frequency, where ε_{21} and ε_{22} are the error bounds. The predicted values of the crack conductivity are 1.99%, 5.38%, 6.48%, and 11.29% after 300 generations. The reconstructed depth is 4.7 mm, which corresponds to the true depth of 5.0 mm.

The same procedure is repeated for the constrained optimization problem that corresponds to the signals of a 30 kHz excitation frequency. The reconstructed crack depth that uses the strategy is 4.8 mm. The reconstructed crack conductivity parameters are 1.99%, 5.11%, 6.92%, and 10.39%. Table 1 summarizes the reconstruction results from the regularization method that uses the multi-frequency ECT signals. The above results indicated that the underestimated problem of the SCC depth is solved to a certain extent via the regularization method and multi-frequency ECT signals.

The reconstruction results from the multi-frequency ECT signals with 5% white Gaussian noise via the regularization method are displayed in Table 2. As shown in Figure 6, the reconstruction results of the crack shape parameters via the regularization method are near the truth even for the signals with a 5% white Gaussian noise under the condition that the predicted signals fit well with those in the testing. The results of repeated iterations indicate that the reconstruction results of the crack shape parameters are slightly different due to the randomness of genetic algorithms. However, certain differences are observed in the conductivity distribution of the crack zone, in which it is an equivalent value that is approximately expressed by several conductivity parameters. The different conductivity distributions may correspond to similar ECT signals, which are a difficult point for quantitative SCC. However, the crack conductivity distribution is generally not the key parameter of electromagnetic nondestructive testing. The purpose of reconstructing the crack equivalent conductivity distribution is to obtain crack shape parameters that are near the true values when excitation frequency ranges from high to low. Therefore, the proposed strategy can improve the reconstruction precision of the SCC shape to a certain extent.

Another example is an elliptical crack model with a depth of 4 mm, length of 13 mm and a width of 0.4 mm. The relative conductivity distribution of crack is from 1% to 8% assigned evenly by 8 parameters. According to the minimal mean-squared error, the reconstructed depth is 3.8 mm and the length is 13.2 mm. Reconstruction results from the proposed strategy and the truth are compared in Figure 7 and Table 3. Reconstructions of several other crack models are also conducted. All the results reveal that the scheme is efficient in improving the reconstruction accuracy of SCC.

5. Conclusion

A scheme for the reconstruction of SCC based on the regularization method and multi-frequency ECT signals is used to deal with the ill-posedness from SCC complicated conductive property and the inverse problem. The reconstruction results of the SCC parameters are improved by gradually determining the approximate distribution of the equivalent crack conductivity in the direction of the crack depth. The regularization

method reduces the ill-posedness of the problem of SCC reconstruction. Reconstruction with simulated ECT signals from conductive crack models is performed. The results reveal that the proposed strategy effectively improves the precision of SCC reconstruction.

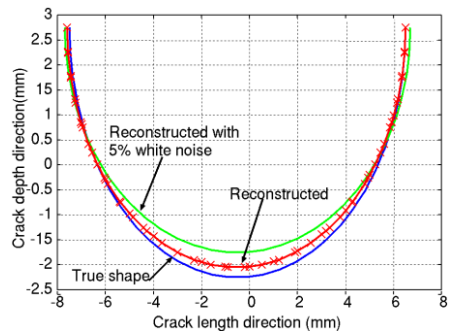


Figure 6. Reconstruction results of crack profiles using simulated ECT signals from an elliptical crack model with the relative conductivity 2%, 5%, 8%, 10% and 15% (with 5% white Gaussian noise)

Table 1. Reconstruction results of an elliptical crack model with the relative conductivity 2%, 5%, 8%, 10% and 15% using multi-frequency ECT signals and the regularization method

Frequency (kHz)	Left edge (mm)	Right edge (mm)	Depth (mm)	Crack conductivity parameters (%)			
714	-7.6	6.5	4.2	2.51	3.22	14.43	22.24
182	-7.6	6.5	4.5	2.14	4.71	7.13	23.18
80	-7.6	6.5	4.7	1.99	5.38	6.48	11.29
30	-7.6	6.5	4.8	1.99	5.11	6.92	10.39

Table 2. Reconstruction results of the above elliptical crack model with 5% white Gaussian noise using multi-frequency ECT signals and the regularization method

Frequency (kHz)	Left edge (mm)	Right edge (mm)	Depth (mm)	Crack conductivity parameters (%)			
714	-7.7	6.7	2.8	2.51	1.88	7.14	8.59
182	-7.7	6.7	3.9	2.18	4.23	7.32	10.45
80	-7.7	6.7	4.1	2.34	4.04	6.51	9.50
30	-7.7	6.7	4.5	2.50	4.38	6.31	9.30

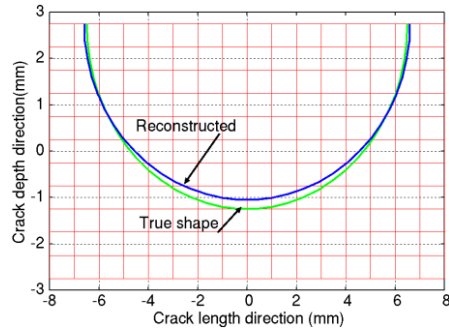


Figure 7. Reconstruction results of crack profiles using simulated ECT signals from a crack model with the crack conductivity from 1% to 8%

Table 3. Reconstruction results of an elliptical crack model with the relative conductivity from 1% to 8% assigned evenly by 8 parameters using multi-frequency ECT signals and the regularization method

Frequency (kHz)	Left edge (mm)	Right edge (mm)	Depth (mm)	Crack conductivity parameters (%)			
714	-6.6	6.6	3.1	1.41	2.98	14.90	9.88
182	-6.6	6.6	3.8	1.65	2.18	8.88	9.93
80	-6.6	6.6	3.8	1.76	2.19	6.73	6.57
30	-6.6	6.6	3.8	1.79	2.23	6.00	6.00

Acknowledgement

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Detect Dampness in Building Walls by Solving the Inverse Problem in Electrical Impedance Tomography

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Abstract. The paper presents a measurement system based on electrical tomography. It will be an innovative solution for an evaluation study of the level of dampness of the walls and the condition of the building. Both in terms of the measuring method as the reconstruction algorithm. The application of modern tomographic techniques in conjunction with topological algorithms will allow to perform a non-invasive and very accurate spatial assessment of the dampness level. The authors of this paper propose a new kind of connection with the use of imaging techniques hybrid solution with surface electrodes.

Keywords. Electrical Impedance Tomography, Inverse Problem, Hybrid Methods

Introduction

One of the major causes of pathologies in historic buildings all over the world is the presence of moisture, particularly rising damp. Moisture transfer in walls of an old buildings, which are in direct contact with the soil, leads to a migration of soluble salts responsible for many building problems [2,7,12-14]. The moisture can be pulled up against gravity (capillary effect). Every technique has its advantages, drawbacks and conditioning allowing it to be used only in particular circumstances [8,10]. Using indirect methods (thermographic camera) there is a possibility of determining water concentration only on wall surface or in its subsurface area. This fact poses a fundamental problem in case of thick barriers, because moisture inside any wall is usually few percent higher than near its surface. Building porous materials (e.g. brick or concrete), both natural and manufactured have pores (like a sponge) and the moisture can be pulled up against gravity (capillary effect). Figure 1 shows damp wall. Rising damp from the soil is a problem in old buildings, especially without adequate horizontal and vertical insulation of foundations. Moisture creates a danger not only to the walls, but also to human health. It promotes progress of rheumatic disorders and formation of fungus on the walls. Fungus can cause allergies and many other diseases. There are many different drainage systems (dry as watertight barriers, injection of hydrofuge products, etc.). Regardless of the method it is very important to continuously monitor the status of damp during the drying process. Proposed hybrid method, in

which assessment of building materials' dampness is an indirect evaluation based on another physical feature, will result in a possibility of conducting multiple measurements without the slightest damage to the examined object. Figure 1 shows damp wall.

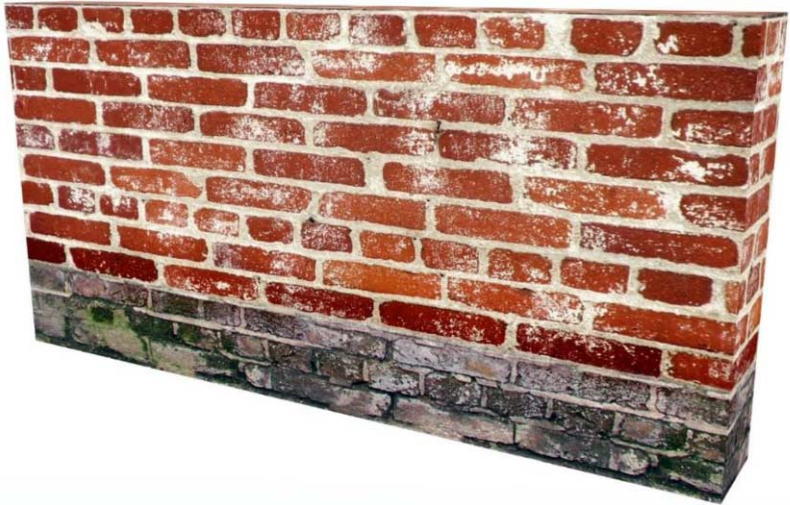


Figure 1. Examples of excessively damp brick walls.

1. Models and measurement system

Electrical tomography is an imaging technique for detecting the internal conductivity distribution of an object by voltage measurements taken by an exterior electrode [1,5-6,15-20]. Numerical methods of the shape and the topology optimization were based on the level set representation and the shape differentiation. There were made possible topology changes during the optimization process. Level set methods, Total Variation and Gauss-Newton methods have been applied very successfully in many areas of the scientific modelling such as the wall dampness [3,4,11]. These approaches were based on shape sensitivity include the boundary design of the elastic interface. The finite element method has been used to solve the forward problem. The proposed solution algorithm is initialized by using the topological sensitivity analysis. Shape derivatives and material derivatives have been incorporated with the level set method to investigate shape optimization problems.

The gradient of the potential distribution (u) is related to electric field strength by:

$$E = -\nabla u \quad (1)$$

Assuming there are no current sources or sinks within the region of interest then from Ampère's law:

$$\nabla \cdot J = 0 \quad (2)$$

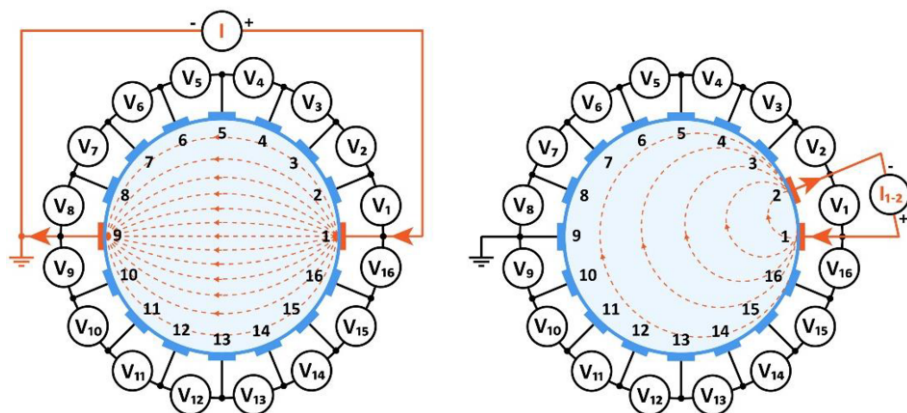


Figure 2. Cross and neighboring measurement methods in electrical impedance tomography.

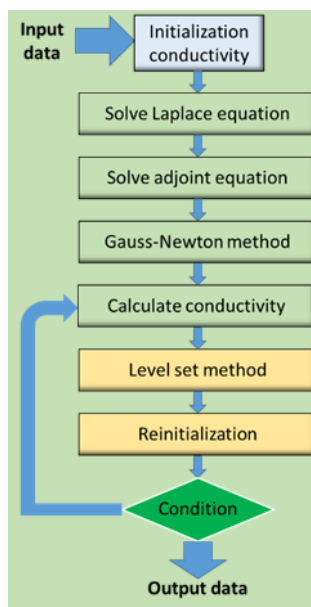


Figure 3. Algorithm to solve the inverse problem by gradient techniques and the level set method.

Typical problem in EIT requires the identification of the unknown internal area from near-boundary measurements of the electrical potential. It is assumed that the value of the conductivity is known in regions whose boundaries are unknown. The forward problem in EIT is described by Laplace's equation:

$$\nabla \cdot (\gamma \nabla u) = 0, \quad (3)$$

where u is potential and γ is conductivity.

Function u is taken under Dirichlet condition in boundary points adjacent to electrodes and Neumann condition on remaining part of the boundary. Problem can be reduced to determination of the minimum value of the functional:

$$I[u] = \frac{1}{2} \int_{\Omega} \gamma |\nabla u|^2 dx dy. \tag{4}$$

Figure 2 presents the cross and neighboring measurement methods of boundary potential data collection illustrated for a cylindrical volume conductor and 16 equally spaced electrodes using in electrical impedance tomography.

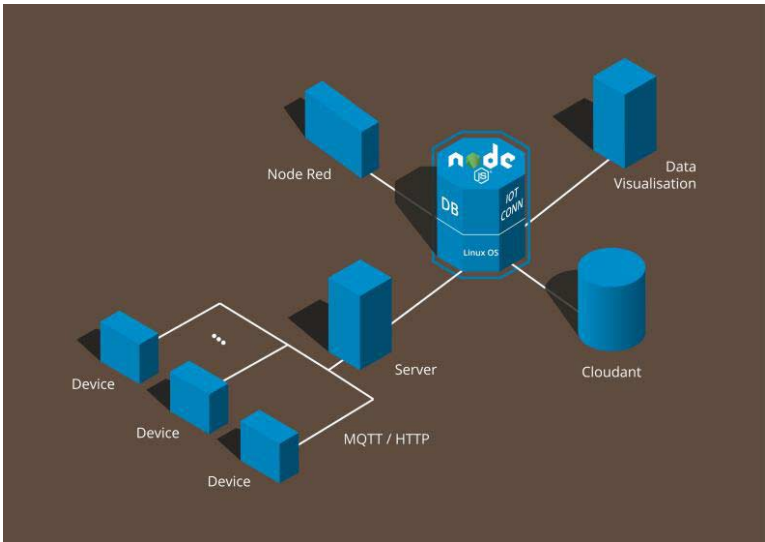


Figure 4. The model of distributed system.

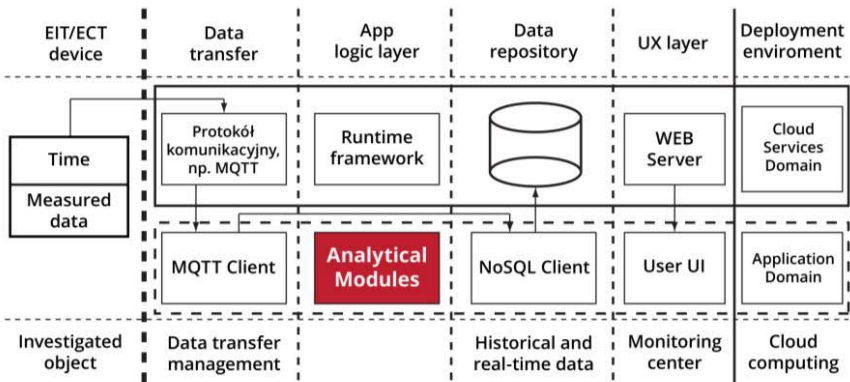


Figure 5. The model of the monolithic application.

The scheme of the algorithm to minimize the objective function is presented in Fig. 3. The system consisted of four layers, as shown in Fig. 5.

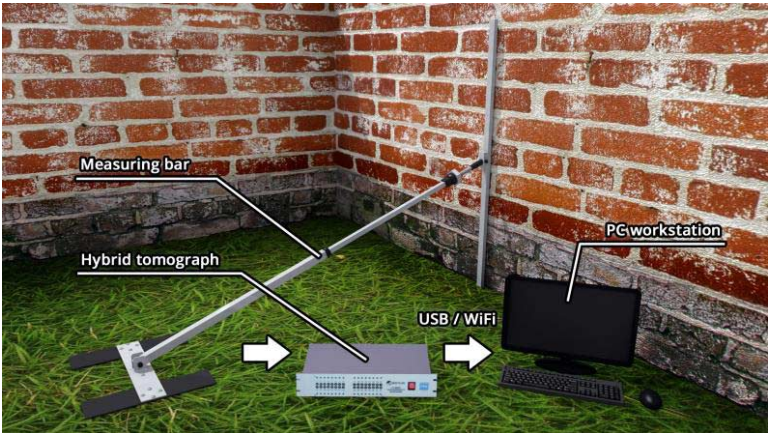


Figure 6. The project of measurement system with the special construction of electrodes.

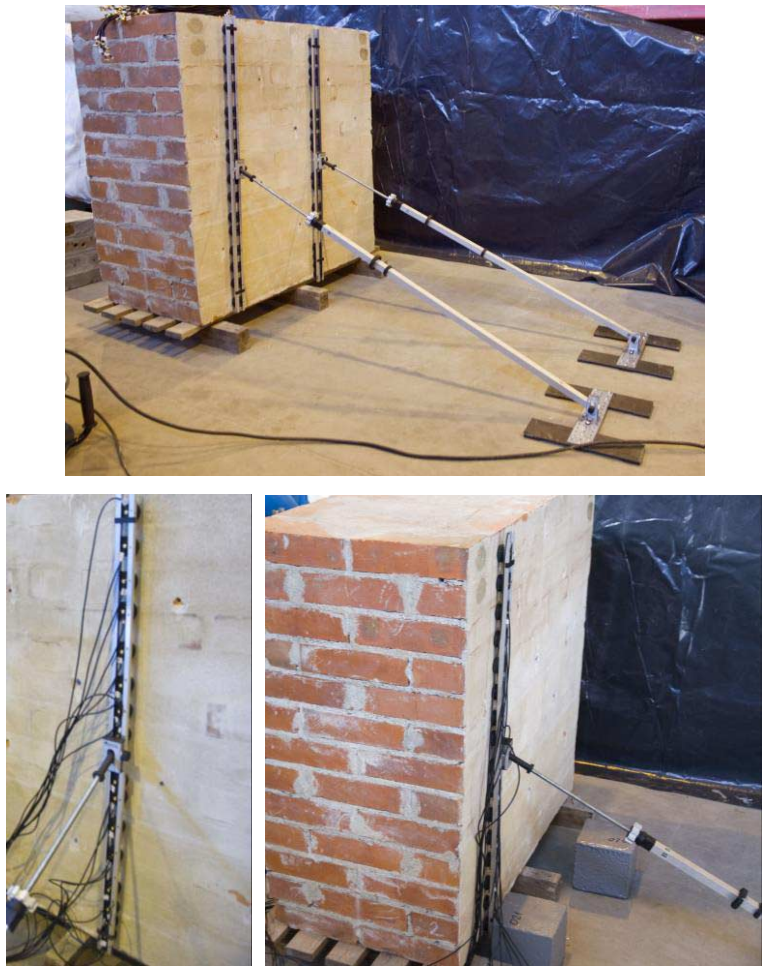


Figure 7. A wall with laboratory measurement system.

The data was acquired from tomographic devices as time series datasets. In this stage MQTT protocol was used to pass the data from the source to the analytical system. The data flow was replicated into two flows: the first one was sent to the database system (for this purpose NoSQL Cloudant database system was used) and the second one was sent to the analytical system for further analysis. The aim of such operation was to process data in real time as well as to keep raw historical data for additional offline analysis (Fig. 4). Figure 6 shows the project of measurement system with the special construction of surface electrodes. A complete electrode consists of three modules: a specific electrode, a PCB with a contact socket and a fastening system. Mechanically, the modules are connected to each other by means of two sleeves placed one inside the other. The test results obtained by the non-destructive impedance tomography method are compared with the results obtained by numerical simulations. A wall with laboratory measurement system is presented in Fig. 7.

2. Numerical Results

Figure 8 shows the image reconstruction of the wall brick for 16 electrodes by the Gauss-Newton method and the Gauss-Newton Level Set Method (hybrid method).

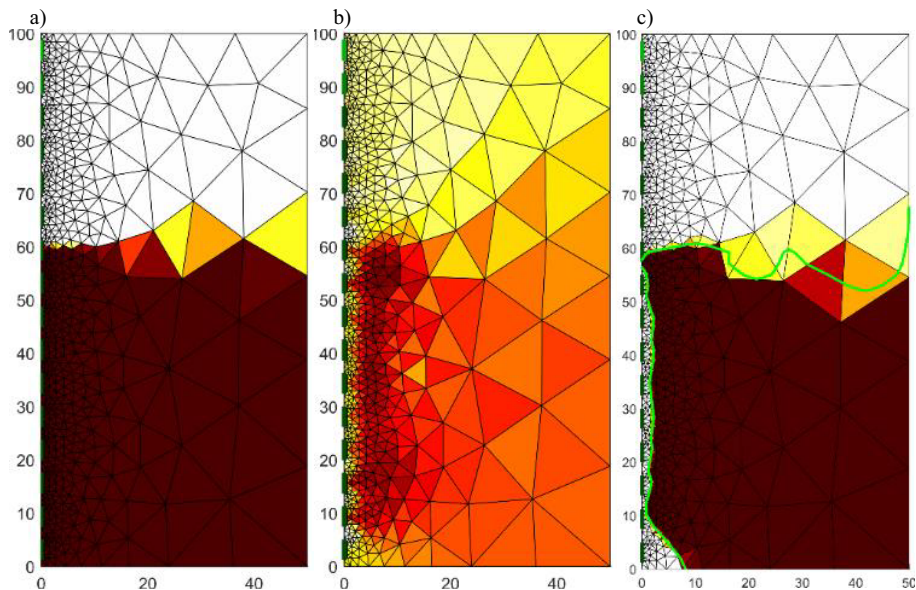


Figure 8. The geometrical model of the investigated dampened wall with 16 electrodes – the line measurement: (a) the initial model, (b) reconstructed using the Gauss-Newton method, (c) reconstructed using the hybrid method.

Unknown structures are marked by the solid line; the damp area was marked with a brown block. This picture presents the geometrical model of the investigated dampened wall with the line measurement. Figure 9 shows the geometrical model 3D with 2 x 2 x 8 electrodes of the image reconstruction: (a) model, (b) Total Variation, (c) Gauss-Newton method. Figure 10 presents the geometrical model 3D with 1 x 2 x 16 electrodes.

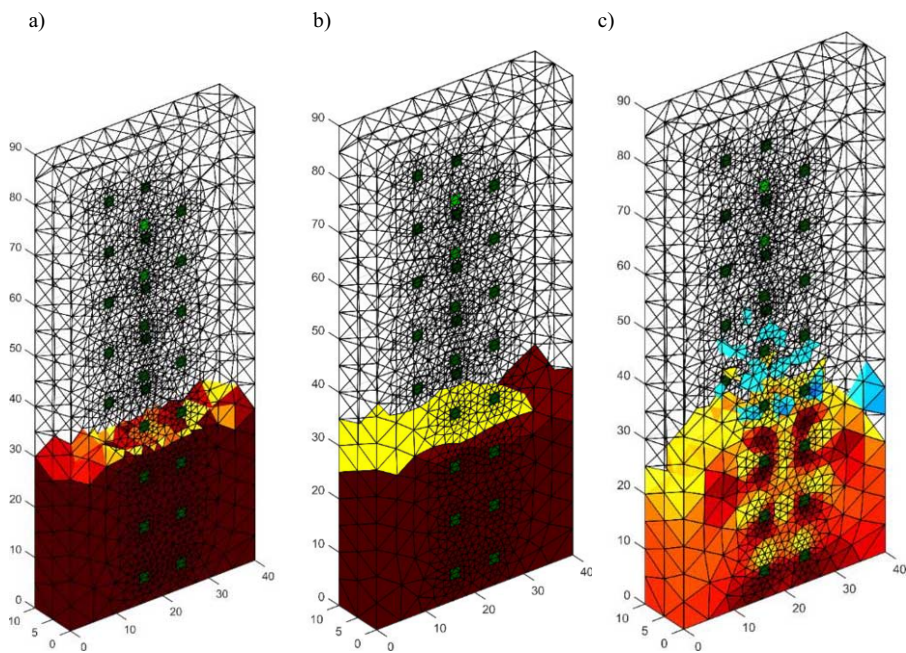


Figure 9. The geometrical model 3D with $2 \times 2 \times 8$ electrodes – the image reconstruction: (a) model, (b) Total Variation, (c) Gauss-Newton method.

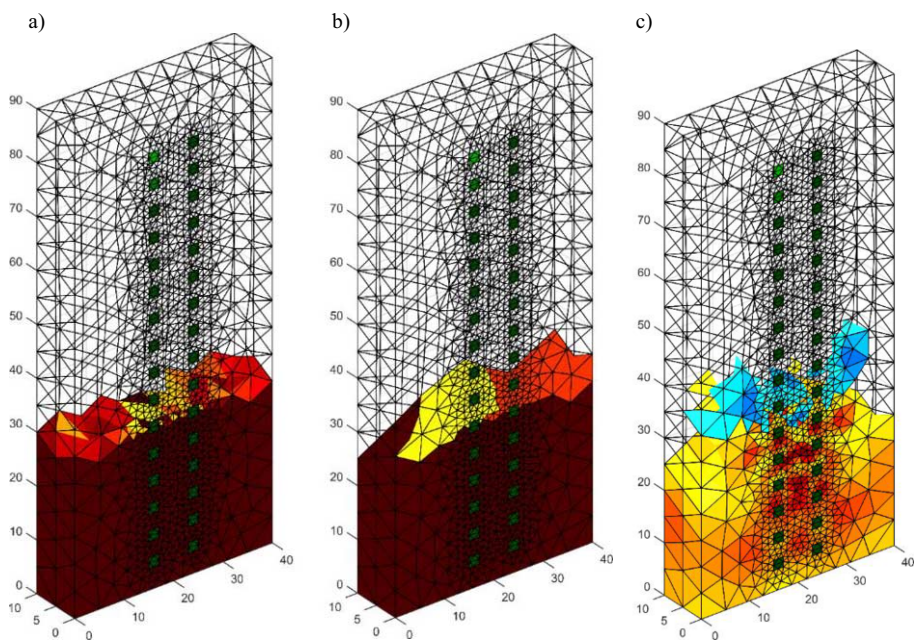


Figure 10. The geometrical model 3D with $1 \times 2 \times 16$ electrodes – the image reconstruction: (a) model, (b) Total Variation, (c) Gauss-Newton method.

3. Conclusion

The non-destructive method of the brick wall dampness was tested by the electrical impedance tomography. The setup was used to determine the moisture of the test wall on specially built models. Numerical methods of the shape and the topology optimization were based on the Gauss-Newton method, Total Variation and the level set representation. The referred algorithms have been applied successfully in the reconstruction of measured data of the model wall. These approaches were based on sensitivity analysis. An efficient algorithm for solving the forward and inverse problems would also improve a lot of the numerical performances of the proposed methods. In modelling of the problem in the electrical impedance tomography, it is required to identify unknown conductivities from near-boundary measurements of the potential. The presented algorithms are effective for spatial determination of moisture in the walls. The Total Variation method gives a better result in 3D. In 2D, however, the hybrid method with topological components is the most effective.

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Time Domain Finite Element Modelling of Pulsed Meander Coil Electromagnetic Acoustic Transducer

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Abstract. This work describes development of a pulsed meander coil Electromagnetic Acoustic Transducer (EMAT) that simultaneously generates Rayleigh, longitudinal waves and shear vertical waves. A two stage finite element model of Meander coil EMAT transmitter is developed to study the propagation of these sound waves in nonmagnetic material. First, a 2D electromagnetic model is developed for calculating the Lorentz force density, which is the driving force for sound wave generation within the material. Second, the calculated force at each point in the material is utilized as the input for generating sound wave modes. Meander coil EMATs of different coil spacings are developed and experimental investigations are carried out. The measurements compare favorably with the model predictions presented here.

Keywords. Meander coil EMAT, finite element modelling, Rayleigh wave, shear vertical wave

1. Introduction

EMATs are transducers that generate as well as receive acoustic waves in electrically conducting materials without making any physical contact with the material unlike the piezoelectric transducers [1]. They are attractive for non-contact ultrasonic testing of thin and thick metallic components, especially at high temperatures. An EMAT consists of stacks of copper coils and magnets and it launches waves based on Lorentz force mechanism [2]. When an alternating current (AC) is pulsed through a coil kept near a conducting specimen surface, eddy currents (j_e) are induced inside the specimen. In the presence of a large bias magnetic flux (B_s) these eddy currents generate body forces (F_L) at the surface layer of the specimen.

$$\vec{F}_L = \vec{j}_e \times \vec{B}_s \quad (1)$$

The Lorentz forces (F_L) on the eddy currents, alternating at the frequency of the driving current, act as a source of ultrasonic waves. The type of wave mode generation depends upon the coil geometry, the operating frequency and the applied magnetic field [3-5].

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Considerable works have been reported on different EMAT probe configurations for specific applications [6-9]. Analytical solutions are currently available for a few types of sources, but are limited by several idealized assumptions such as simple uniform force distributions, infinite coils, and the high frequency excitation coil, etc. [10]. On the other hand, numerical methods have been widely used either to model the electromagnetic part for the Lorentz force density calculations or the elasto-dynamic part for wave propagation separately [11-13]. Most of the transient field studies have been performed by using either finite element method (FEM) or finite difference method (FDM) [14-16]. A fully numerical calculation lacks an explicit physical interpretation of the relationship between the coil source and the resulting ultrasonic field. It also requires a huge computer memory for high frequency models because of the mesh requirements for the Lorentz force calculation and for the generation of ultrasonic waves. In order to avoid these difficulties, coupled two stage finite element models are desired. They are useful to study the realistic conditions and a wider range of EMAT probe design configurations. A two stage FE simulation of electromagnetic acoustic transducer has already been explained for generation of plate/Lamb waves and interaction of these waves with artificial defects in thin aluminum plates [17].

In this paper, a hybrid FEM for meander coil (MC) EMAT transmitter has been developed for generation of angle beam bulk waves and Rayleigh waves in thick samples with different coil geometries. The model can be employed for investigating non-magnetic metals where the operation of the EMAT is due to the Lorentz force mechanism alone. In order to validate the FE results, MC EMATs have been developed for experimental measurements. The objective of this work is to develop an accurate model of a pulsed MC EMAT transmitter to generate Rayleigh waves and angle beam shear vertical (SV) waves with an optimum frequency. Then, the predicted wave modes at different frequencies are analyzed and compared with the experimental results.

2. Electromagnetic model

The commercial finite element software COMSOL Multi-physics has been used to develop the electromagnetic model for pulsed MC EMAT transmitter. This software package has predefined applications called modes with built-in mathematical solutions to facilitate modeling. Figure 1 shows the MC configuration and the arrangement coil with the application of static magnetic field (B_s) to the material.

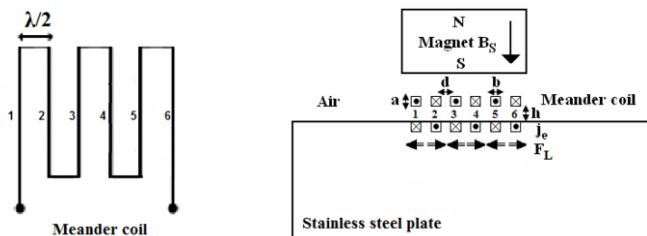


Figure 1. The configuration of MC and the arrangement of MC EMAT with permanent magnet over a stainless steel plate.

The test specimen is a non-magnetic ($\mu=\mu_0$) stainless steel plate and the magnet is permanent Neodymium-Iron-Boron magnet with flux density 0.3 T. The MC is made up of copper with six turns and the size of the coil is 0.25 mm thick, and the breadth of the coil is 0.5mm. The transient excitation current applied to the source coil is assumed to be given by [17],

$$i_k(t) = \beta e^{-\alpha(t-\tau)^2} \cos(2\pi f_c(t-\tau) + \phi) \quad (2)$$

where β is constant current (50 A), α is the bandwidth factor ($5 \times 10^{11}/\text{sec}^2$), τ is the maximum arrival time of the signal (3×10^{-6} sec), ϕ is the phase difference and f_c is the central frequency ranges from 500 kHz to 2 MHz for generating different sound modes. A constant coil lift-off of 1mm is considered throughout the simulation. Figure 2a shows the transient input current $i_k(t)$ of 6 μsec which is applied to a MC EMAT with 1mm coil spacing kept above the test specimen. Figure 2b shows the temporal distribution of the Lorentz force densities observed under the first and second coils (as shown in Figure 1) from the right and near to the surface of the specimen. The time history of the force density follows the input excitation current and this observation has been done at 1 MHz central frequency.

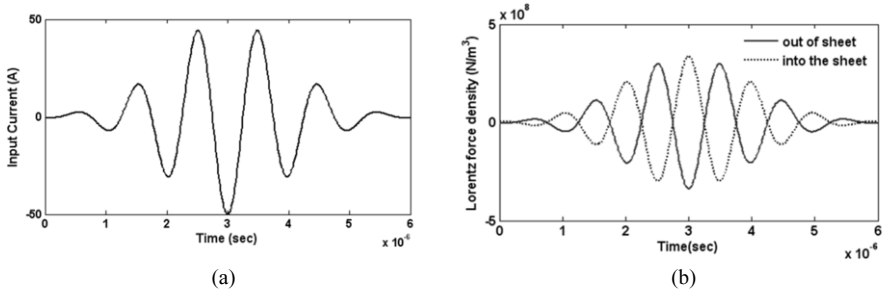


Figure 2. (a) The transient excitation current of 1MHz center frequency and (b) Lorentz force distribution on the surface of the material.

3. Ultrasonic wave propagation model

The wave propagation modeling has been done by using ABAQUS explicit scheme. It uses an explicit integration scheme for solving the transient dynamic and quasi-static analyses [18]. The output of the electromagnetic model has been utilized as the input for the wave propagation model. The simulation of wave propagation can also be done by using the COMSOL multi-physics with acoustic mode, in particular for low frequency bulk waves, Lamb waves or guided waves related applications for small structures. However, this implicit solver fails to converge for a large bulk structure at higher frequencies. As per Courant–Friedrichs–Lewy (CFL) condition, the mesh size (L) and the time increment (Δt) depend up on the excitation frequency [17]. At high frequencies, the values of time increment and the mesh size are very small which result in unstable and oscillatory solution without convergence.

In the wave propagation part, a plain strain model has been developed by using explicit dynamics procedure for a large domain at high frequency Rayleigh and bulk

shear waves. The elasto-dynamic model has been designed with the stainless steel plate and the force densities have been applied as input sources for the simulation. The computed spatial and temporal force densities have been used as the input functions, in the discrete wave propagation model. The source inputs have been applied as concentrated forces on both surface and subsurface more than a wavelength deep inside the specimen. Application of these forces results in generation of time dependent elastic stress waves inside the specimen. The displacement of the elastic waves can be written in terms of force and density of the material

$$\mu \nabla \times (\nabla \times \vec{u}) - (\lambda + 2\mu) \nabla (\nabla \cdot \vec{u}) + \rho \frac{\partial^2 \vec{u}}{\partial t^2} = \vec{F}_L \quad (3)$$

where ρ is mass volume density and u is the material displacement vector. The acoustical properties of the material are those of stainless steel with Lamé constants μ and λ of 7.5×10^{10} N/m² and 10.7×10^{10} N/m², respectively. The longitudinal and shear wave velocities C_L and C_S are therefore, 5.8 and 3.1 mm/ μ s, respectively. The whole geometry was discretized spatially into elements of dimensions around $1/20^{\text{th}}$ of the wavelength of the transmitter excitation frequency. CPE4R (4-noded bilinear plane strain quadrilateral with reduced integration) was chosen for meshing the stainless steel plate [18]. Two steps namely, excitation and propagation steps were defined. During the excitation step, the varying force inputs were applied at the surface and subsurface of the material up to more than one wavelength depth; and during the propagation step, the generated waves were propagated inside the specimen. The wave propagation models were designed for the two different coil spacings for the generation of Rayleigh and bulk shear waves.

4. Meander coil (MC) EMAT

The MC EMAT consists of a high frequency coil and a normal biasing permanent magnet. For optimum design consideration, the high frequency coil is the most important component and its geometrical configuration determines the generation of different sound modes. These coils are made by flexible PCB technique which allows design of any arbitrary pattern with high accuracy. The printed circuits boards are made by 150 μ m polyester based flexible laminate with 50 μ m thick copper clad. The total thickness of the sheet is 150 μ m. The permanent magnet used for designing the EMAT is Neodymium-Iron-Boron (Nd-Fe-B) sintered magnets. It exhibits a flux density B_s of nearly 0.3 T (with dimensions 25x25x12.5 mm³).

The developed meander line EMAT with a normal biasing Nd-Fe-B permanent magnet and a flexible PCB copper coil is shown in Figure 3. The MC EMATs have been developed with the same design parameters used in the simulation studies. This type of EMAT produces a periodic driving force by a spatially varying current distribution and a spatially constant magnetic field. The interaction of eddy current and the perpendicular magnetic field induces Lorentz forces parallel to the surface, whose directions change alternatively with the meandering period [19]. Such a body force distribution generates Rayleigh waves traveling along the surface of the specimen and the longitudinal waves and shear waves traveling obliquely into the specimen. Due to

the periodicity of the meandering coil, grating lobes are also generated. The generation of wave modes and the grating lobes depends upon the coil spacing and the operating frequency [20]. Figure 3 also shows the cross sectional view of the MC EMAT and the wave modes generation inside the specimen.

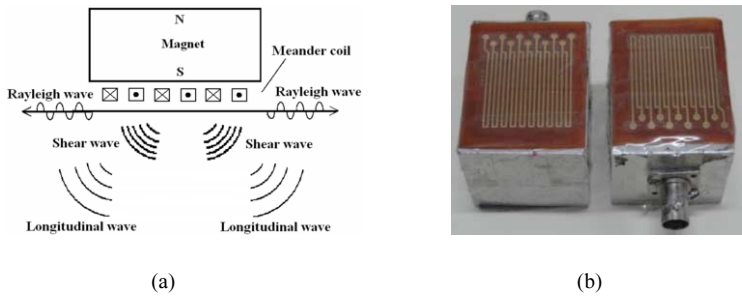


Figure 3. (a) Cross sectional view of MC EMAT and (b) MC EMAT developed for measurements using flexible PCB coils and permanent magnets.

5. Experimental details

In order to get a high power RF tone burst, a RITEC RPR 4000 gated amplifier was used. Suitable impedance matching transformers were used for both the transmitter and receiver EMAT coils. A Ritec broadband receiver was used through another impedance matching network to match the low impedance receiver coil to the high impedance receiver. The signals were recorded using an Agilent DSO6032A digital storage oscilloscope interfaced with PC for data archival and post processing. Figure 4 shows the schematic and photograph of the set up used for the experimental studies. A peak current of approximately 50 A and pulse duration of 6 μ sec pulse was fed to the exciter EMAT. The transmitted signal was received by another identical MC EMAT. A stainless steel block with dimension $300 \times 200 \times 50 \text{ mm}^3$ was used as test specimen.

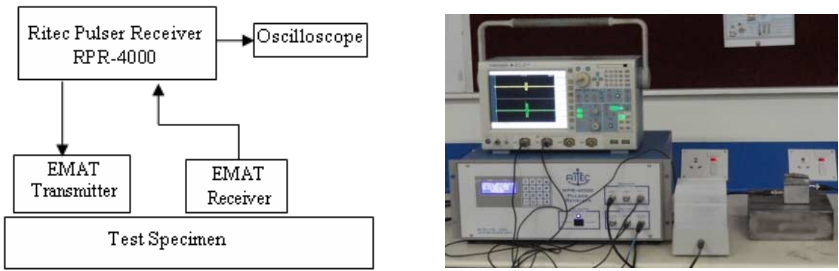


Figure 4. Schematic and photograph of the experimental setup used for measurements.

6. Results and discussion

Through transmission technique was used for the measurements. The transmitter EMAT was placed on top of the specimen and the receiver EMAT was oriented along the axis of the wave front generated by the transmitter EMAT. The received signals were analyzed and compared with the model predicted signals.

6.1. Generation of Rayleigh waves

The Rayleigh wave generation was done by using two different coil spacings 1 mm and 2mm and the corresponding excitation frequencies are 1.2MHz and 760 kHz, respectively. Figure 5(a) shows the measurement setup for analyzing the Rayleigh wave generation using MC EMAT for both simulations and experiments. The receiver EMATs is kept 100 mm away from the center of the transmitter, which was oriented along the axis of the wave front generated by the exciter EMAT. The optimum frequency for generating the Rayleigh wave with maximum signal strength has been obtained by sweeping the excitation frequency from low to high.

Figures 5(b) and 5(c) show the time domain A-scan signal of Rayleigh waves generated by MC EMATs with 1 mm and 2 mm coil spacings at the optimum excitation frequencies of 1.2 MHz and 760 kHz respectively. The first arrived Rayleigh wave signal with large amplitude at $34 \mu\text{s}$ is the direct received signal from the source and the signal at $69 \mu\text{s}$ is the reflection from the end of the steel block. The velocities of the received signals are calculated and verified with the corresponding Rayleigh wave velocity (2930 m/s) of the material. At low frequency, the generation of Rayleigh wave is found to be more efficient than the other two bulk waves. The MC EMATs with coil period equal to half of the wavelength ($\lambda/2$) are found to generate Rayleigh waves predominantly more and the efficiency is found to increase with increasing coil spacing.

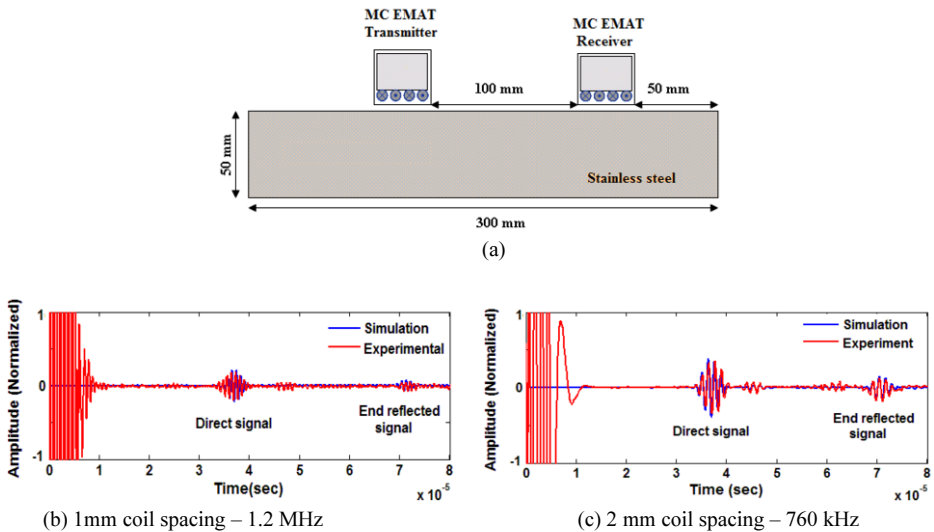


Figure 5. a) Schematic diagram for Rayleigh wave measurement set up, b) and c) Comparison of time histories of simulation and experimental outputs of Rayleigh wave modes for different coil spacings.

6.2. Generation of angle beam shear vertical (SV) waves

Above the Rayleigh wave peak frequency, the MC EMAT generates vertically polarized angle beam shear (SV) wave that is more efficient than Rayleigh wave. Figure 6(a) shows the set up for generating SV waves in experiments and simulations. The receiver EMAT is kept 84 mm away from the transmitter to receive the back reflected shear wave. With a proper choice of exciting frequency, the MC EMAT generates SV wave inside the material. Figures 6(b) and 6(c) show the time domain A-scan signal of SV waves generated by MC EMATs with 1 mm and 2 mm coil spacings at the optimum excitation frequencies of 2.4 MHz and 1.2 MHz respectively.

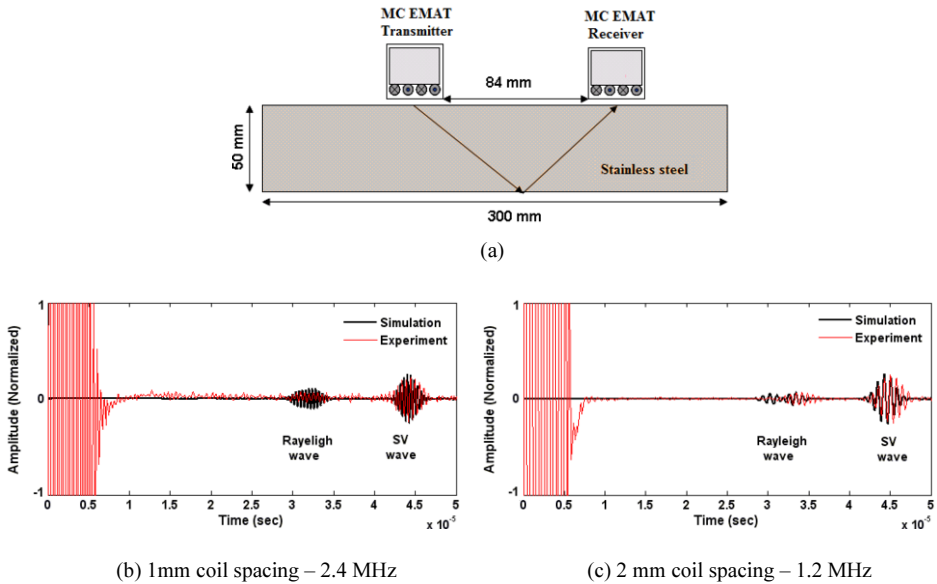


Figure 6. a) Schematic diagram for SV wave measurement set up, b) and c) comparison of simulation and experimental outputs of SV wave for different coil spacings.

The MC EMAT transmitter is found to generate small amplitude lateral wave even at the shear wave excitation frequency. The Rayleigh and SV waves can be readily distinguished from the time of flight (TOF) and velocity calculations in both experiments and simulations. The generation of SV wave is more predominant at an angle (maximum amplitude) of $\sim 40^\circ$ from the MC EMAT transmitter. Figure 7 shows the simulation result of 1 mm coil spacing with three modes traveling in different directions. The SV generation has been done for the two coil geometries. The experimental results are compared with the simulation results. This comparison shows the wave generation, TOF and the mode shape are exactly matching with these results.

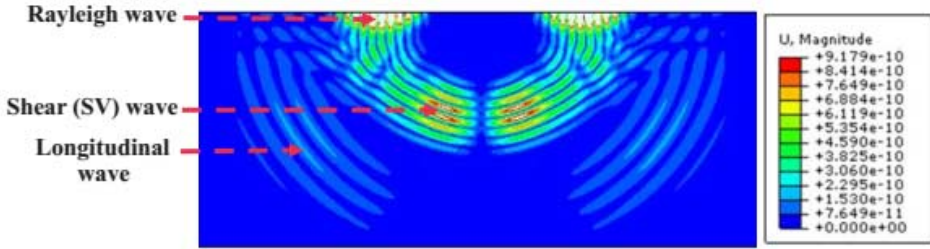


Figure 7. Simulation model results plot of displacement magnitudes (in m) of the propagating Rayleigh waves and angle beam shear (SV) and longitudinal waves at $10.2 \mu\text{s}$ at 760 kHz (for 1 mm coil spacing)

7. Conclusion

A meander coil (MC) EMAT has been developed to simultaneously generate Rayleigh waves and angle beam SV waves, after systematic numerical simulations using a coupled 2D model. First, an electromagnetic model has been developed for calculating the Lorentz force density and then, an elasto-dynamic model has been used for wave propagation. The predicted Lorentz forces have been utilized as the sources for the ultrasonic wave propagation. MC EMATs have been developed using flexible copper coils and Nd-Fe-B magnets for experimental studies and to validate the model predictions. The generation of Rayleigh wave and SV wave has been studied for two different coil spacings and the model predicted results have been compared with the experimental results.

Studies reveal that the Rayleigh wave generation efficiency increases with the increasing coil spacing at low frequencies. The Rayleigh wave peak frequency has been measured for the two coil spacings. The MC EMAT is found to generate SV wave mode that is more prominent at an angle of $\sim 40^\circ$ for both the coil spacings. The directional characteristics of the Rayleigh and shear wave modes generated by the transmitter with various coil spacings have been analyzed at different excitation frequencies. A very good agreement has been observed between the numerical simulation predictions and the experimental measurements.

Acknowledgements

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Evaluation of a Dielectric Inclusion Using Inductive RF Antennas and Artificial Neural Networks for Tissue Diagnosis

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Abstract. In this paper, the relevance of the non contact RF evaluation of the complex permittivity of organic material is addressed by means of a computational approach. The authors consider a simple configuration constituted of a single loop RF antenna interacting with a dielectric material mimicking a typical organic tissue (the electrical conductivity is 0.6 S/m and the dielectric constant is 80) which includes a buried inclusion (e.g. a tumor featuring a conductivity of 0.2 to 1.6 S/m and a dielectric constant ranging from 20 to 160). First a three dimensional semi analytical model (DPSM) is implemented in order to evaluate the sensitivity of such an antenna to the complex permittivity of the buried inclusion. Then, the inverse problem which consists in evaluating the complex permittivity, the size and the location of the inclusion is addressed by means of an artificial neural network (ANN) approach. For the considered configuration (5 mm radius antenna, 40 mm radius spherical inclusion buried at a 5 to 20 mm depth within the tissue, antenna operated at 135 MHz), the main conclusions are that the complex permittivity and the depth of the inclusion can be fairly estimated (estimation error smaller than 5%), even in the case of antenna positioning uncertainties, providing the ANN is adequately trained. Also, a double antenna configuration significantly enhances the estimation of the location and size of the inclusion.

Keywords. RF inductive sensor, complex permittivity characterization, non-contact and non-invasive tissue evaluation, DPSM, artificial neural network.

1. Introduction

The dielectric properties (DP) of materials are due to and informative about the composition and structure of materials. As a result, in medical applications, the DPs are considered as valuable indicators to assess the physiopathological state of biological tissues. They are found very useful to evaluate changes in the state of tissues, in order to monitor e.g. the ageing of skin, the healing of wounds, or to detect tumors [1].

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Indeed, for example, breast tumor DPs are reported as exhibiting more than 70% changes relatively to those of sound tissues at around 100 MHz [2]. The DPs are quantified by the complex relative permittivity ε^* such that:

$$\varepsilon^* = \frac{\varepsilon}{\varepsilon_0} + \frac{\sigma}{j\omega\varepsilon_0} \quad (1)$$

where ε is the dielectric constant of the material, σ is its electrical conductivity, $\omega = 2\pi f$ is the angular frequency and f is the frequency, $j = \sqrt{-1}$ and ε_0 is the dielectric constant of vacuum. Various dielectric characterization techniques have been developed for the estimation of the DPs of organic materials, which are mostly based on the use of electrodes, transmission lines or coaxial probes [3]–[5]. These measurement techniques are possibly implemented in a large frequency bandwidth in order to provide spectral information about the material complex permittivity. However, due to the necessity of a contact between the probe and the material, they suffer from possible loss of accuracy and might be found too invasive in some applications [6]. To overcome the limitations, microwave (MW) techniques based on the use of transmit and receive antennas and inspired from Radar imaging techniques have been considered for tissue characterization [7]. These techniques benefit from being non contact and non invasive. Moreover, since these systems operate in a high frequency range (i.e. over tens of GHz), the antennas are considered as radiating in a far field mode, which simplifies both the experimental implementation and the processing of the experimental data. These MW techniques are sensitive to the material dielectric constant changes, resulting from a change of the amount of water in the tissue, e.g. due to the increased vascularization of the tumors [8]. However, the presence of tumors is also characterized by a change of the electrical conductivity [2].

This is why non-contact inductive techniques operating at lower frequencies (e.g. in the kHz to tens of MHz bandwidth) have been also considered for the evaluation of organic materials. These so-called magnetic inductive spectrometry (MIS) techniques are theoretically able to sense both the dielectric constant and the electrical conductivity [9]. However, in practice, they are hardly sensitive to the dielectric constant, possibility because of a lack of sensitivity of the used sensors [10]. However, recent works have experimentally shown that it is possible to implement non contact RF inductive sensors, based on the use of resonant antennas operating in the radiofrequency (RF) range (tens to hundreds of MHz), to monitor the changes of the complex permittivity of organic materials in an easy to implement scheme [11].

In this study, the authors aim at evaluating the relevance of such non contact RF dielectric evaluation scheme, for the sensing of an inclusion buried in a media mimicking a tissue which includes a tumor. To do so, the authors consider a computational approach. In section 2, the used case-study configuration is presented. In section 3, the interactions between the antenna and the media are modeled using a semi-analytical mesh-free method. The computation method is briefly presented and the sensitivity of the antenna to the complex permittivity changes of the inclusion buried within the media is studied. In section 4, the inverse problem, which consists in estimating the complex permittivity, the position and size of the inclusion starting from the antenna impedance changes, is addressed by means of an artificial neural network

approach for various configurations including a double antenna configuration. Section 5 is on conclusion.

2. Studied configuration

A simplified configuration is considered in this study as depicted in Figure 1.a. The tissue is modeled by a semi-infinite homogeneous dielectric material of conductivity σ_2 and relative permittivity $\epsilon_{r2} = \epsilon_2 / \epsilon_0$. A spherical inclusion of radius r featured by conductivity σ_3 and relative permittivity ϵ_{r3} is used to model a tumor, and a single loop circular antenna of radius a is used as a transmit and receive inductive sensor (antenna 1). The inclusion is buried at a depth d_2 within the tissue and at a distance d_3 of the revolution axis of the antenna. A second antenna (antenna 2) is possibly placed at a distance d_4 from antenna 1. In addition, an experimental test bench was set up according to this configuration (Figure 1.b.).

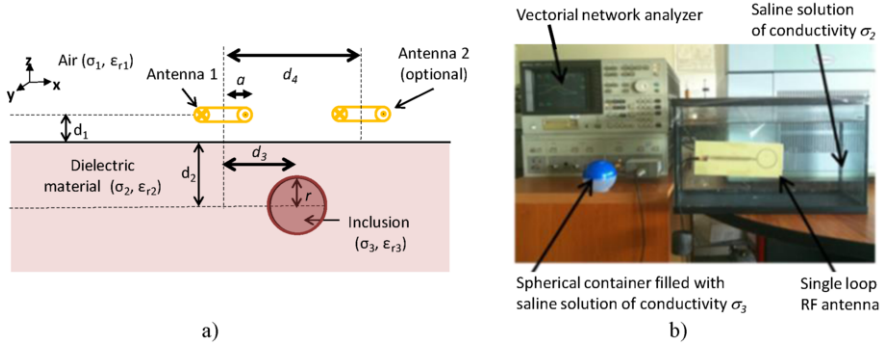


Figure 1. a) Basic configuration considered in this study and b) associated experimental test bench

The test bench is constituted of a 20 liter glass tank (length $\times$ width $\times$ height = $37 \times 25 \times 22$ cm³) filled with a NaCl solution prepared so that $\sigma_2 = 0.6$ S/m and $\epsilon_{r2} \approx 80$ [12], these dielectric parameters being close to those of typical organic tissues [1]. A $r = 40$ mm spherical plastic container is used to constitute the inclusion. The container is filled with NaCl solutions of various concentrations leading to $\sigma_3 = 3 \cdot 10^{-3}$, 1, 4.5 and 8 S/m (and $\epsilon_{r3} \approx 80$). The container is placed within the tank and maintained against the wall of the glass tank so that $d_2 = 40$ mm. A printed board circuit circular antenna of radius $a = 20$ mm is placed against the tank wall so that $d_1 = 4$ mm. The dimensions of the set up are chosen large enough so as to minimize mispositioning issues, and the size of the antenna is chosen small enough so that the tank can be considered as semi-infinite comparatively to the magnetic field radiated by the antenna. The impedance changes are read out at the ends of the antenna by means of a HP4195A vectorial network analyzer equipped with a 41951A impedance analyzer kit. The set up is implemented at frequencies $f = 11, 19$ and 27 MHz.

3. Modeling the antenna / media interactions using the DPSM

The aim of the computation is to numerically determine the impedance changes occurring at the ends of the antenna interacting with the tissue, defined by $\Delta Z = Z_0 -$

Z_{load} , where Z_0 is the antenna impedance in the air and Z_{load} is the impedance in presence of the tissue. To do so, the distributed point source method (DPSM) [13] was used. This mesh-free semi-analytical modeling method is based on the use of discrete current sources which radiate as Green's functions in the whole workspace, and to compute the impedance of the antenna starting from the numerical integration of the magnetic flux flowing across the loop section [13].

3.1. Basic principle of the DPSM modeling

To compute the distribution of the magnetic flux density at any point of interest, N_s elementary current sources J_s are used to model the antenna, and $2 \times N_A$ virtual current sources J_A are used to map the air / tissue interface, and $2 \times N_B$ other virtual sources J_B are used to map the tissue / tumor interface. To be more specific, N_A virtual current sources (J_{A1}) are distributed above the air / tissue interface, and radiate in the tissue; N_A virtual current sources (J_{A2}) are distributed under the air / tissue interface, and radiate in the air (Figure 2.a). In the same way, virtual sources J_B model the tissue / tumor interface: sources J_{B1} are distributed around the “outer” face of the tumor and radiate within the tumor, while sources J_{B2} are distributed along the inner face of the tissue / tumor interface, and radiate into the tissue. The intensity of the current sources J_s are arbitrarily chosen to be equal to 1 A. The intensities of the virtual sources are numerically determined so as to satisfy the Helmholtz equation and the boundary conditions that apply on the magnetic vector potential A at each interface included in the workspace [13].

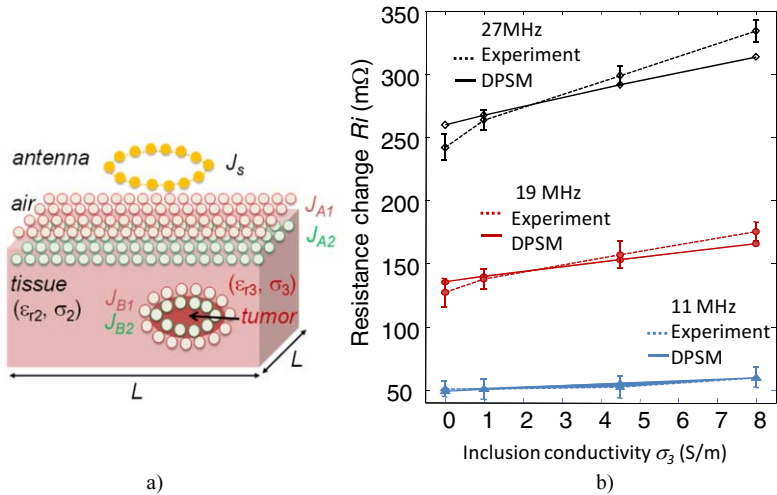


Figure 2: a) DPSM modeling of the antenna / material interactions and b) comparison between computed and experimental data

Finally, A is determined at the points of interest from the radiation of all current sources considered in the workspace, and the magnetic flux density is computed from A . The DPSM was fully described in [13]. It is chosen here since it allows any kind of interface shape to be considered, and the accuracy / time of computation trade-off of the modeling can be easily adjusted by choosing the convenient number of current sources in the workspace. Here, the DPSM is validated against experimental data

provided by the experimental test bench. To do so the DPSM workspace was adjusted to meet with the experimental configuration described in Section 2 and with $N_s = 100$, $N_A = 1000$, $N_B = 250$. The air / tissue interface is a $L \times L$ square surface with $L = 400$ mm, and the tissue / tumor interface is a sphere of radius $r = 40$ mm. Also, the impedance change of the antenna is computed from the distribution of the magnetic flux flowing across the antenna section computed at 10 000 observations points distributed in the antenna plane. This DPSM set up enables the impedance change of the antenna to be computed with Matlab in less than 100 s using a 12 core 96 Go RAM workstation. As an example, the real part of the measured and computed impedance change ΔZ , denoted R_i , is presented in Figure 2.b. for 3 different frequencies and 4 values of the inclusion conductivity. Error bars are relative to the standard deviation of the data estimated from a set of 10 measurement data. One can note the good agreement between computation and measurements. Discrepancies may be attributed to an inaccurate positioning of the elements constituting the test bench (antenna, inclusion...), as well as to the use of a limited set of sources in the DPSM computations.

3.2. Sensitivity of the antenna to dielectric property changes of the inclusion.

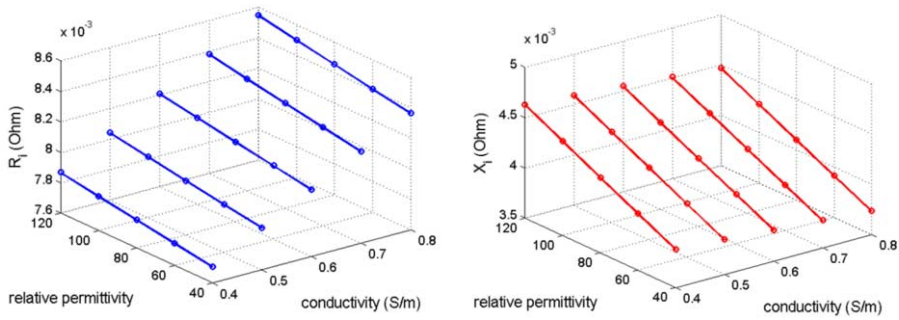


Figure 3. Real part R_i and imaginary part X_i of the antenna impedance changes ΔZ versus dielectric parameters of the inclusion.

In this section, the DPSM is used to study the sensitivity of the antenna to the dielectric parameters of the inclusion. To do so, the configuration of Figure 1 is considered with antenna 1 only. The used parameters are $a = 5$ mm, $r = 40$ mm, $d_1 = 2$ mm, $d_2 = 45$ mm, $d_3 = 0$ mm, $\sigma_2 = 0.6$ S/m, $\epsilon_{r,2} = 80$, and the dielectric properties of the inclusion are σ_3 which ranges from 0.4 to 0.8 S/m and $\epsilon_{r,3}$ which ranges from 40 to 120. The frequency used for the computations is chosen so that the loss factor, which is defined by:

$$\tan \delta = \frac{\sigma_2}{2\pi f \epsilon_{r,2} \epsilon_0}, \quad (2)$$

is equal to 1 in the sound tissue, i.e. $f = 135$ MHz. This frequency is chosen since it is expected to provide equal sensitivity to the dielectric constant and to the conductivity of the material [11]. Computations results provided by the DPSM are presented in Figure 3. One can note that both the real R_i and the imaginary part X_i of the antenna impedance change are affected by both dielectric parameters, and one can conclude that

the inclusion can be sensed by the antenna. Further computations showed that these results are strongly affected by the position (d_2 , d_3) of the inclusion relatively to the antenna, and of the size (r) of the inclusion. As a result, these parameters should be taken into account in the solving of the inverse problem which is considered in the next section.

4. Evaluation of the dielectric inclusion using an artificial neural network approach

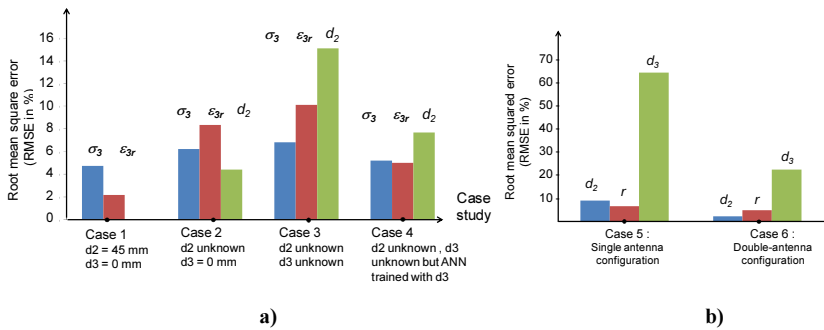
In order to estimate the inclusion parameters starting from the values of the antenna impedance changes, a model-free estimation technique based on the use of an artificial neural network (ANN) is considered. More precisely, a feed-forward neural network (FFNN) is used [14] and implemented using the Matlab neural network toolbox [15]. The inputs of the FFNN are the real and imaginary parts of the antenna impedance changes, and the outputs are the wanted inclusion parameters. To do so, impedance data relative to various antenna / tissue configurations are computed using the DPSM. Some white noise is added to the data so that the signal to noise ratio is 40 dB, close to that of actual experimental data. The noisy data are then divided into three data sets: the learning database, which includes the training database and the validation database, and the test database. These data sets are used to separately optimize the weights, the biases and the size of the FFNN (number of neurons of the hidden layer). The training database is used to feed the network during the training process, and the network parameters are adjusted according to the estimation errors observed in known configurations. The validation database is used to measure the network generalization ability and to halt the training process when generalization stops improving [15]. The testing database is used to provide an estimation of the network performances during and after training. In what follows, the sound tissue is assumed to be known and such that $\sigma_2 = 0.6$ S/m, $\varepsilon_{r2} = 80$. All data are computed at $f = 135$ MHz.

4.1. Estimation of the dielectric nature and depth of the inclusion

We firstly consider the problem of estimating the dielectric parameters and depth of the inclusion buried within the sound tissue. To do so, four different antenna / tissue configurations are considered and described in Table 1 (cases 1 to 4). The estimation results obtained using the FFNN scheme are presented in Figure 4.a., in the form of a root mean square error of estimation (RMSE), expressed in % relatively to the parameter actual values. In case 1, only the dielectric properties of the inclusion are unknown. They are estimated by the FFNN with a RMSE lower than 5%. In case 2, the depth d_2 of the inclusion is unknown, and the FFNN is trained so as to estimate d_2 in addition to the dielectric properties of the inclusion. Here, all parameters are correctly estimated, even if the RMSE is higher than in case 1. In case 3, an error in the positioning of the antenna relatively to the inclusion (d_3) is introduced between the learning and the testing procedure (Table 1). As a result, the RMSE is significantly increased, especially regarding the estimation of d_2 . In case 4, the FFNN is trained taking into account the uncertainty of the position of the antenna (d_3). One can note that in this case, the estimation errors for σ_3 , ε_{r3} and d_2 are significantly reduced (lower than 8%), compared to those of case 3 (up to 16%, Figure 4.a.).

Table 1. Considered antenna / tissue configurations

		σ_3 (S/m)	ϵ_{r3}	r (mm)	d_1 (mm)	d_2 (mm)	d_3 (mm)	d_4 (mm)
Case 1	Learning	1-1.7	20-160	40	2	45	0	-
	Test	1.2-1.6	30-150	40	2	45	0	-
Case 2	Learning	1-1.7	20-160	40	2	40-60	0	-
	Test	1.2-1.6	30-150	40	2	43-55	0	-
Case 3	Learning	1-1.7	20-160	40	2	40-60	0	-
	Test	1.2-1.6	30-150	40	2	43-55	40	-
Case 4	Learning	1-1.7	20-160	40	2	40-60	0-60	-
	Test	1.2-1.6	30-150	40	2	43-55	40	-
Case 5	Learning	1.7	160	30-40	2	40-62	0-60	60
	Test	1.7	160	32-38	2	43-61	20	60
Case 6	Learning	1.7	160	30-40	2	40-62	0-60	60
	Test	1.7	160	32-38	2	43-61	20	60

**Figure 4.** a) Estimation of the dielectric parameters and depth of the inclusion. b) Estimation of the location and size of the inclusion for single and double antenna configurations.

4.2. Localization of the inclusion using a single and double antenna scheme

Finally, the problem of localizing and sizing an inclusion of known dielectric properties is addressed using a single or a double antenna configuration (Figure 1.a.). The considered configuration parameters are gathered in Table 1, and the estimation results are presented in Figure 4.b. One can note that the estimations of the depth d_2 , the radius r and the position d_3 of the inclusion is significantly enhanced using a double antenna configuration (case 6) compared to the single antenna one (case 5).

5. Conclusion

In this study, the authors aimed at evaluating the relevance of a non contact RF dielectric evaluation scheme, for the sensing of an inclusion buried in a media mimicking a tissue including a tumor. To do so, the authors consider a computational approach. First, a DPSM modeling is considered to study the sensitivity of an RF antenna to the complex permittivity of a dielectric inclusion buried within a sound tissue. Then, a FFNN approach is used to evaluate the inclusion starting from the antenna impedance changes. The main conclusions are as follows: firstly, we found that the complex permittivity and the depth of the inclusion can be correctly estimated from the antenna impedance changes. Moreover, the antenna positioning uncertainties

can be compensated with appropriate training of the FFNN. Secondly, the depth and size of an inclusion of known permittivity can be correctly estimated with a single antenna. The addition of a second antenna provides enhanced estimation and localization results. These preliminary results open the way to the design of efficient inclusion characterization schemes based on inductive RF sensor arrays. Further works will consider a wider range of inclusion sizes and depths, and focus on the evaluation of such approach with some actual experimental data.

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Measurement of Plating Thickness with High Liftoff Using Eddy Current Testing

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Abstract. This study deals with measurement of the thickness of a zinc layer deposited on steel in the process of zinc plating. A short distance between the blower that blows liquid zinc and the sensor that measures the thickness of the layer, and a long distance between the sensor and test samples are required in the process. We attempted to apply eddy current testing to measure the thickness of the zinc layer with high liftoff. This paper demonstrates the eddy current distribution calculated using the finite element method and the measured liftoff and thickness dependencies at each excitation frequency of an eddy current sensor. The frequency that results in a phase difference of approximately 90° between dependencies was determined. This indicates that the thickness and liftoff length signals can be separated from the measured signal when an appropriate excitation frequency is selected. These results demonstrate the possibility of making thickness measurements using an eddy current with high liftoff.

Keywords. Eddy current testing, Zinc layer, Measuring thickness, High liftoff

1. Introduction

Zinc plating is widely used to prevent the oxidation of steel. The thickness of the zinc layer has an impact on the cost of plating and the prevention of oxidation. Therefore, the thickness is controlled during the process of plating, as shown in Figure 1. Sensors that measure the thickness are required to achieve a greater distance between the gauges and the steel plates, i.e., the liftoff, than several tens of millimeters owing to the vibration of plates during the process. X-ray thickness gauges are used in the process because these can evaluate the thickness of the zinc layer with high liftoff. However, the distance between the thickness sensors and air nozzles should be reduced to control the thickness of the zinc layer because the time lag between the measurement and control of the thickness of the zinc layer should be as short as possible. However, the sensor/air nozzle distance in the manufacturing process which uses the X-ray thickness gauges is several tens of meters. Thus, the users of the process are motivated to shorten the distance.

Eddy current testing (ECT) is a technique used to measure the thickness of a thin layer [1-5]. The sensors of the ECT are smaller than the X-ray thickness gauges. Furthermore, the sensors can generally be used in a high temperature field. Thus, the sensors can be placed near the air nozzle during the plating process. However, to the best

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of our knowledge, the ECT sensors have not been applied to a process involving vibration of samples, because the change of liftoff owing to vibration has a greater impact than the thickness on the output of the ECT sensors. To use the eddy current sensors in the process of plating, they are required to measure dependencies of detection voltages on the zinc layer thickness, which are smaller than the dependencies on the liftoff.

This paper demonstrates a method to measure the thickness of plating, and further demonstrates the possibility of thickness measurements using ECT with high liftoff.

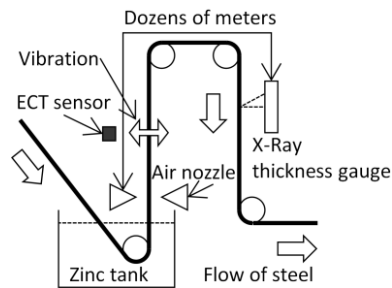


Figure 1. Illustration of the process of plating. The air nozzles control the thickness of the zinc layer based on the thickness measured by the sensor. Therefore, it is desirable that the distance between the nozzles and sensors is as small as possible.

2. Visualization of an eddy current in a zinc layer by using FEM

Figure 2, Tables 1 and 2 show an illustration, dimensions and parameters of an eddy current sensor and a specimen. Eddy current distributions and phase dependencies on the thickness and liftoff of a zinc layer on steel were visualized using a finite element method (FEM) program, the JMAG Designer[®], which was developed by JSOL Corporation. Figure 3 shows the simulation results. The graphs in Figure 3 show the distributions of depth direction of the magnitude and phase of the eddy current. As shown in Figures 3 (a) and (b), the magnitude is sensitive to the liftoff and thickness of the zinc layer. The magnitude of the signal owing to the change of the liftoff is a few times greater than the magnitude of the signal owing to the change of the thickness of the zinc layer. This indicates that the change of liftoff must be sufficiently small for an evaluation method focusing only on the magnitude of the signal. However, the phase of the eddy current is not as sensitive to the liftoff as it is to the thickness of the zinc layer, as shown in Figures 3 (c) and (d).

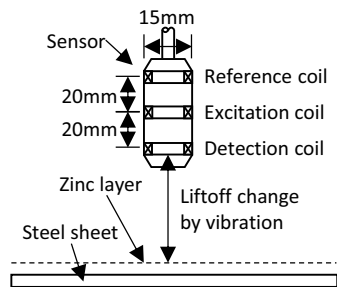


Figure 2. Overall configuration of setup

Table 1. Dimensions of coils of ECT sensor.

Items	Details
Internal diameter	13.5 mm
External diameter	15 mm
Number of turns	100
Height	7 mm

Table 2. Parameters of specimen.

Items	Conductivities and relative permeabilities
Zinc	1.66×10^7 S/m, 1
Steel sheet	7.46×10^6 S/m, 3600

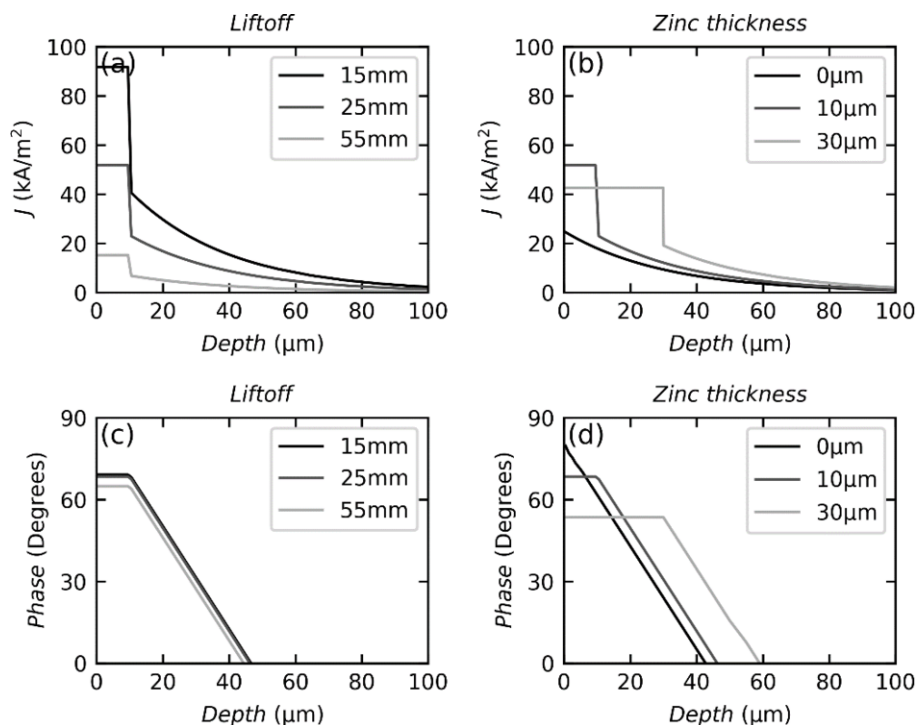


Figure 3. Distributions of depth direction of the eddy current calculated using FEM at 10 kHz. (a) and (b): eddy current density vs. depth location from the surface. (c) and (d): phase of eddy current vs. the depth location. The magnitude is sensitive to the liftoff and the thickness of the zinc layer. However, the phase is not more sensitive to the liftoff than to the thickness of zinc layer.

3. Experimental setup and results

An experiment was conducted to determine the optimal experimental conditions. The frequency of excitation is important in ECT because it is known that the effective depth of the eddy current, i.e., the skin depth, is determined using the frequency. The frequency and phase of the eddy current also have a relationship. An appropriate excitation frequency must be determined to separate the signals of the thickness and liftoff from the measured signal. The experimental conditions employed are described as follows. Seven specimens were used. Table 3 lists the details of the specimens. The specimens can be classified into two categories—electro galvanization and hot dip galvanization—based on the plating type. The thickness of each specimen was measured at six positions using a microscope. Table 4 lists the experimental conditions. The shape of the sensor used in this experiment was the same as that used in the simulation. The sensor had three coils. The excitation current flowed through the excitation coil. The detected signal was the difference between the voltages of the detection and reference coils. Several excitation frequencies from 1 kHz to 50 kHz were used. When the frequency of 100 kHz was used, the skin depth of zinc was 389 μm. The skin depth was larger than the thickness of the zinc layer of the specimens. As shown in Figure 4, a power amplifier (NF HSA4014) and two lock-in amplifiers (LIAs) (NF LI5640) were used. The LIAs measured the excitation current and detection voltage. The signal of the oscillator for the current was used as the

reference signal of the LIAs. The output signals of the LIA for the current were measured to confirm the excitation current at each frequency. The output signals i.e., the x - and y -components of the LIA for the detection voltage, were evaluated.

Table 3. Details of specimens.

No.	Plating type	Thickness of zinc layer (μm)
0	Non-plating	0
1	Electro galvanization	6.4 ± 1.2
2	Electro galvanization	15.3 ± 1.8
3	Electro galvanization	23.3 ± 5.4
4	Electro galvanization	20.7 ± 0.7
5	Electro galvanization	29.5 ± 0.9
6	Hot dip galvanization	16.5 ± 0.6
7	Hot dip galvanization	26.4 ± 2.4

Table 4. Experimental conditions.

Items	Details
Sample shape	220 mm×530 mm×1 mm
Liftoff	30 to 50 mm
Excitation current	200 mA

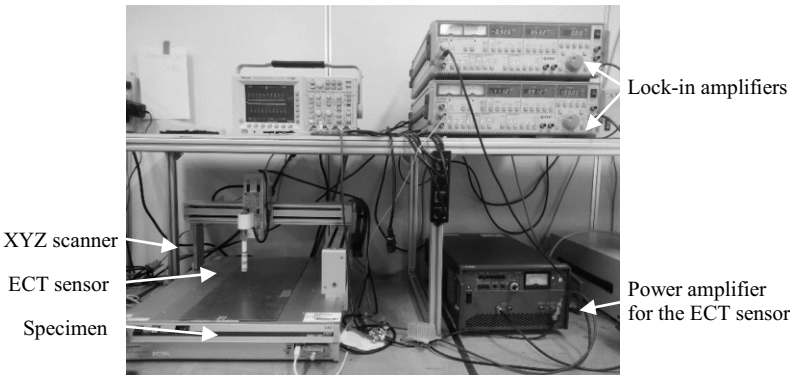


Figure 4. Experimental setup. The ECT sensor was moved by the XYZ table and excited by the power amplifier. The excitation current of the excitation coil of the sensor and the differential voltage between the detection and reference coils were measured using the lock-in amplifier.

The results obtained are described as follows. Figures 5 and 6 show the measured signals of specimens with zinc layers of thicknesses 0, 6.4, and 23.3 μm . The x - and y -components of the LIA for the detection voltage were the in-phase and quadrature components with the excitation current. However, the phase was rotated so that the measured lines of No. 0 at 40 mm liftoff were parallel to the x -axis of the graphs in all the figures. As expected from the simulations, it can be seen that the phase of detection voltages was insensitive to the liftoff, compared with the magnitude of detection voltages. As shown in Figure 5, the liftoff of the sensor changed to 30 mm from 50 mm when the results were measured. The liftoff change had an impact on the y -component. As shown in Figure 6, the liftoff of the sensor changed to 35 mm from 45 mm when the results were measured. These results at 1 to 10 kHz indicate that the liftoff did not have a large impact on the y -component. Therefore, we focus on the relationship between phase differences

and excitation frequencies in Figure 6. The phase difference means a phase θ in Figure 5. Figure 7 was made from Figure 6. As shown in the figure, the phase difference accesses to zero degrees as the frequency is higher. The difference between the signals of the samples in the y -component had a peak at approximately 1 to 2 kHz. Therefore, 2 kHz is considered the appropriate excitation frequency in this situation. Figure 8 shows the relationship between the y -component and the thickness of the zinc layer. The thickness is proportional to the component at the liftoff of 40 mm. The error bar of each plot shows the impact on the component owing to a liftoff change of ± 5 mm. This graph shows that the resolution of the thickness of zinc layer is approximately 10 μm in this method. Thus, it is expected that the error bar will be small when the change in liftoff is small.

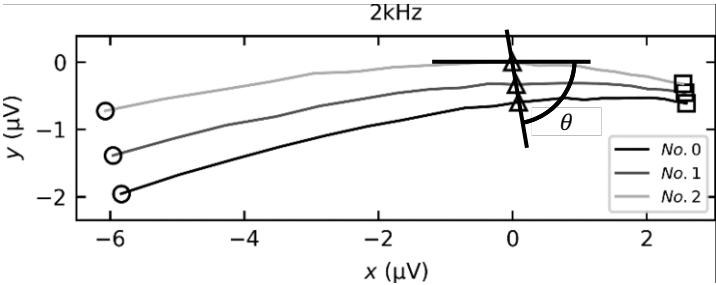


Figure 5. Measured results at 2 kHz. Circle plots, $\bigcirc$, show the results of liftoff of 30 mm. Triangle plots, Δ , show the results of liftoff of 40 mm. Square plots, $\square$, show the results of liftoff of 50 mm. The θ shows a phase difference between the liftoff and zinc thickness dependencies.

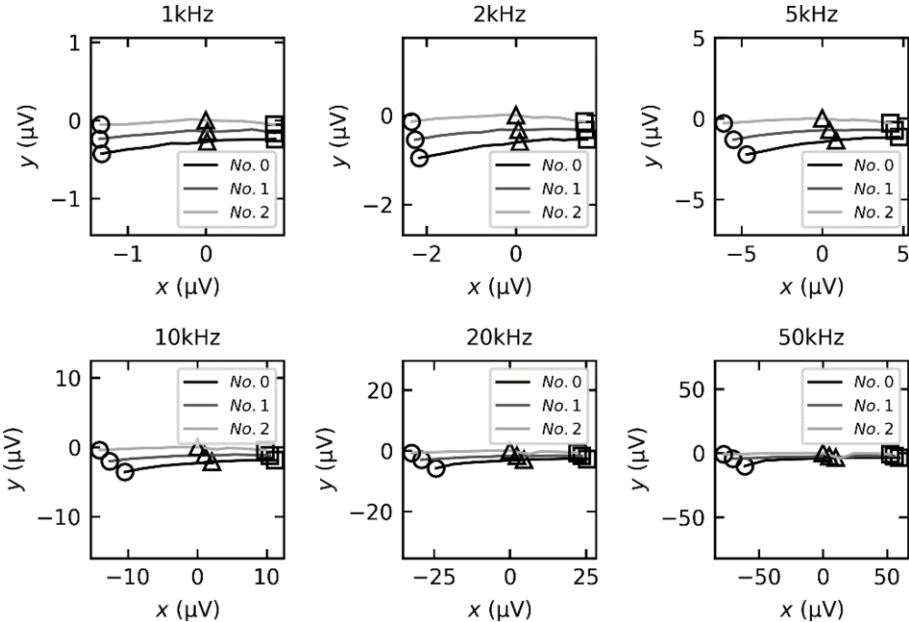


Figure 6. Measured results at each frequency. Circle plots, $\bigcirc$, show the results of liftoff of 35 mm. Triangle plots, Δ , show the results of liftoff of 40 mm. Square plots, $\square$, show the results of liftoff of 45 mm.

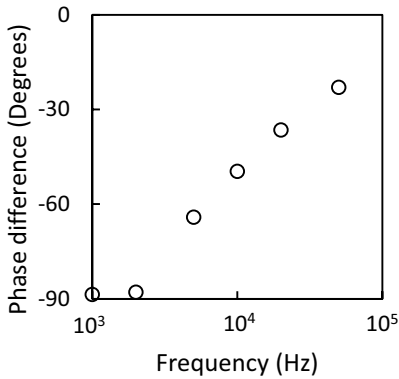


Figure 7. Relationship between phase differences and excitation frequencies. The phase difference at 2 kHz shows θ in Figure 5. The phase difference accesses to zero degrees as the frequency is higher. The difference between the signals of the samples in the y-component had a peak at approximately 1 to 2 kHz.

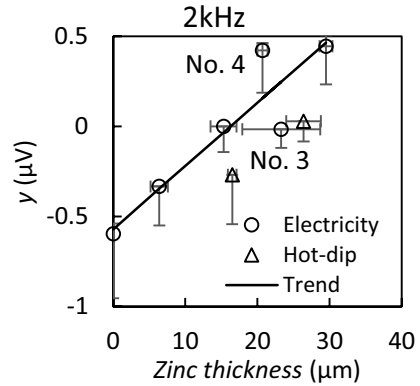


Figure 8. Relationship between the y-component and the thickness of the zinc layer at the excitation frequency of 2 kHz and lift-off of 40 mm. The error bars of each plot show the impact on the component owing to a lift-off change of ± 5 mm. The thickness is proportional to the component without No.3 and 4 in samples of the electro galvanization.

4. Summary

In this paper, ECT was applied to measure the thickness of a zinc layer on steel with high lift-off. The simulation results show that the phase dependencies of the eddy current on the thickness of the zinc layer and lift-off were different. In the experiments, the frequency that resulted in a phase difference of approximately 90° between the dependencies was determined. An eddy current sensor detected the difference between zinc layers of 6 and 23 μm thicknesses with a lift-off of 5 mm. These results show the possibility of making thickness measurements using an eddy current with high lift-off. However, a lift-off higher than 10 mm is required for this process. We will evaluate the dependencies of the sensor shape to increase the lift-off.

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Effect of Heat Treatment on Barkhausen Noise Behaviours of Deformed Low Carbon Steel

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Abstract. We investigated the Barkhausen noise profiles for the undeformed and deformed low carbon steel when the specimens were heat treated with modifications of microstructures of specimen: recovery, recrystallization process and stress relief. The Barkhausen noise parameter is almost constant in heat treatment at a lower temperature, while it shows a rapid decrease with heat treatment time at higher temperatures. The recovery process contributes to the decreases of Barkhausen noise signal in heat treatment at higher temperatures.

Keywords. Barkhausen noise, recovery, recrystallization, stress

1. Introduction

It is quite important to evaluate nondestructively degradation and residual stress of infrastructure construction such as bridges, expressways and turbine blades to maintain an integrity of constructions and a safety of human life. The degradation of constructions is based on microstructure changes of the steels composed the constructions. Almost steels show ferromagnetic properties and their magnetic properties depend strongly on microstructure changes, which means nondestructive evaluation (NDE) using magnetic measurement for degradation of infrastructure construction is possible [1]. One of effective NDE methods using magnetic measurement is Barkhausen noise (MBN) techniques [2, 3], because MBN is attributed to discontinuous domain wall movements. Domain wall movements depend on the magnetoelastic coupling between domain walls and microstructures. A lot of researches regarding the correlation between the magnetic properties including MBN properties and the mechanical characteristics during plastic deformation have been investigated [4] - [6]. There have also been a number of studies related to the correlation between magnetic properties and hardness when the grain size is altered [7] - [9]. In those researches, magnetic properties have good correlation with hardness. However, there are reports that the correlation between the magnetic properties and the mechanical characteristics is not simply when the steel that was plastically deformed has been heat treated [10], [11]. To enhance a reliability of the MBN method, investigation of the relationship between MBN properties and microstructures during heat treatment is required. Therefore, in this study, we evaluated MBN properties of deformed low carbon steel with cold rolling and investigated the effects of microstructure changes due to heat treatments on MBN profiles.

2. Experimental procedure

Low carbon steel (S15C) plates were prepared and one of those steels was undeformed and others were deformed by cold-rolling with reduction ratios of 5, 20 and 40%. The reduction ratio means the ratio of thickness of steel before and after cold-rolling. The composition of the steel is 0.16 wt.% C, 0.2 wt.% Si, 0.44 wt.% Mn, 0.019 wt.% P, 0.018 wt.% S, and 0.01 wt.% Ni and Fe in balance. The plates were shaped into a disk of 24 mm diameter and 2 mm thickness by wire cutting. Those plates were heat treated in vacuum at 573, 773 and 873K (300, 500, 600°C) with heat treatment time from 0 to maximum 90 minutes for 773 and 873K and maximum 180 minutes for 573K followed by quenching in water. After heat treatment, the specimens were polished and then MBN properties of those specimens were measured using a magnetic yoke made of Fe-Si steel and an air-core coil. When measuring, a triangular current I of 1 Hz with an amplitude of 1A is applied to an excitation coil having 150 turns winding around the yoke for magnetizing a specimen, and an induced voltage at the air-core coil, 1000 turns and 64 mm² area and 4 mm height, made of Cu wire with 0.08 mm diameter, located at the surface of the specimen was captured by a PC after amplified by 1000 times and then filtered (100 – 200 kHz). Figure 1 (a) shows a typical example of measured MBN signal. The black solid line shows the applied current to excitation coil, and gray solid line is MBN raw signal. The rms value defined as the equation (1) is used for evaluation of MBN signal.

$$rms = \sqrt{\frac{1}{T} \int_0^T V_o^2 dt} \quad (1)$$

where V_o is the detected signal after amplified and filtered, and T is a half period of magnetizing field. Based on the results of MBN signal, MBN profile in which the moving averaged rms voltage, V_{rms} [12] are plotted against the current was obtained as shown in Figure 1 (b). The profile by the solid line is signal changes when the applied current changes from minus to plus and the dotted line shows the data when the current changes plus to minus. The MBN signal was subsequently evaluated by changing the direction of the magnetizing field. Here, the direction parallel to the direction of cold-rolling is defined as 0 degree, and then the measurements were done from 0 – 90°. We also evaluated the Vickers hardness of the specimens by a hardness meter with 300 g load.

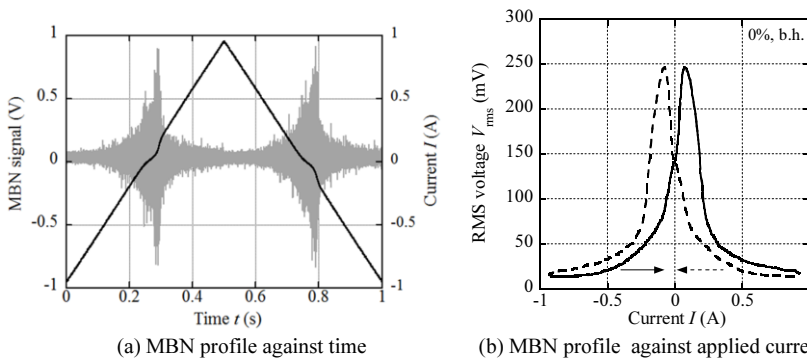
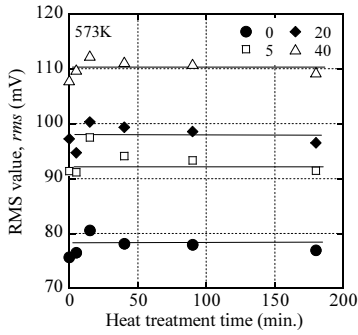
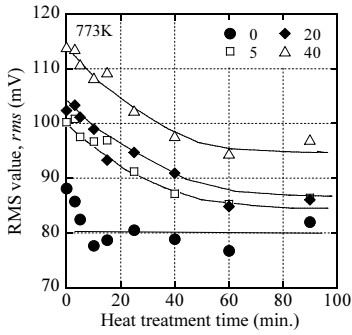


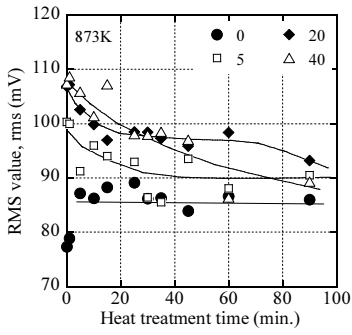
Figure 1 Time dependence of MBN signal and applied current to excitation coil (a). MBN signal changes against applied current to the excitation coil. (b.h. means “before heat treatment.”)



(a) 573K

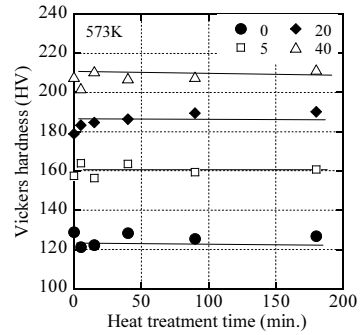


(b) 773K

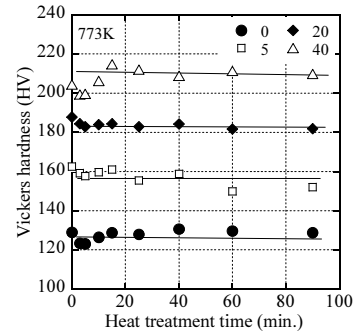


(c) 873K

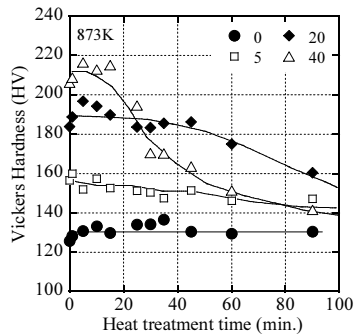
Figure 2 MBN signal dependence on heat treatment time for undeformed and deformed specimen with different heat treatment temperature.



(a) 573K



(b) 773K



(c) 873K

Figure 3 Vickers hardness dependence on heat treatment time for undeformed and deformed specimen with different heat treatment temperature.

3. Experiment results

3.1 Changes in rms voltage and Vickers hardness due to heat treatment

Figure 2 shows the dependence of rms voltage V_{rms} on the heat treatment time for the specimens with different reduction ratios at different heat treatment temperature. All measurements were done by magnetizing field applied parallel to rolling direction. Before heat treatment, the value of V_{rms} increases with increasing reduction ratio,

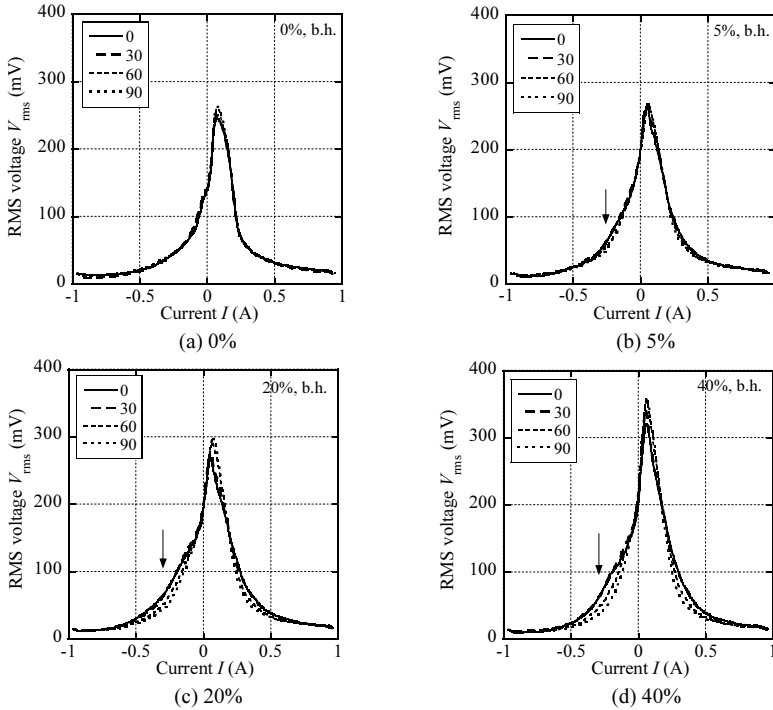


Figure 4 MBN profile measured with different applied field direction for undeformed and deformed specimens before heat treatment. (b. h. means “before heat treatment.”)

which is explained by the increase of dislocations [13]. The value of V_{rms} is nearly constant for all specimens at 573K (300°C), while the value decreases with heat treatment time at 773K (500°C) and 873K (600°C) except undeformed specimens (0%). The reduction ratio of V_{rms} at 873K is larger than that at 773K, and the value becomes nearly constant at earlier heat treatment times.

Figure 3 shows the dependence of Vickers hardness on heat treatment time for the specimens with different reduction ratios at different heat treatment temperature. The Vickers hardness increase with increasing reduction ratio before heat treatment; this is caused by the increase of dislocations. The Vickers hardness is constant against heat treatment time for all specimens at 573 and 773K. On the other hand, there are changes at 873K, which depends on the reduction ratios. The specimen with a 40 % reduction shows rapid decreases around 20-25 minutes, then, continues to decrease gradually. The specimen with 20% shows the constant value up to 40 minutes, then decreases. 5% reduction slightly decreases and 0% shows no change.

3.2 MBN profiles and orientation dependence

Figure 4 shows the MBN profiles for the undeformed and deformed specimens with different magnetizing direction before heat treatment. All profiles for the undeformed specimen are same profile, whereas the MBN profiles show magnetic anisotropic properties for the deformed specimens; the profiles of all angles have a significant peak at $I = 0 - 0.1$ A, and that of 0 or 30° for undeformed specimen have a broad peak around $I = -0.3$ A and the peak becomes prominent as the reduction ratio increases

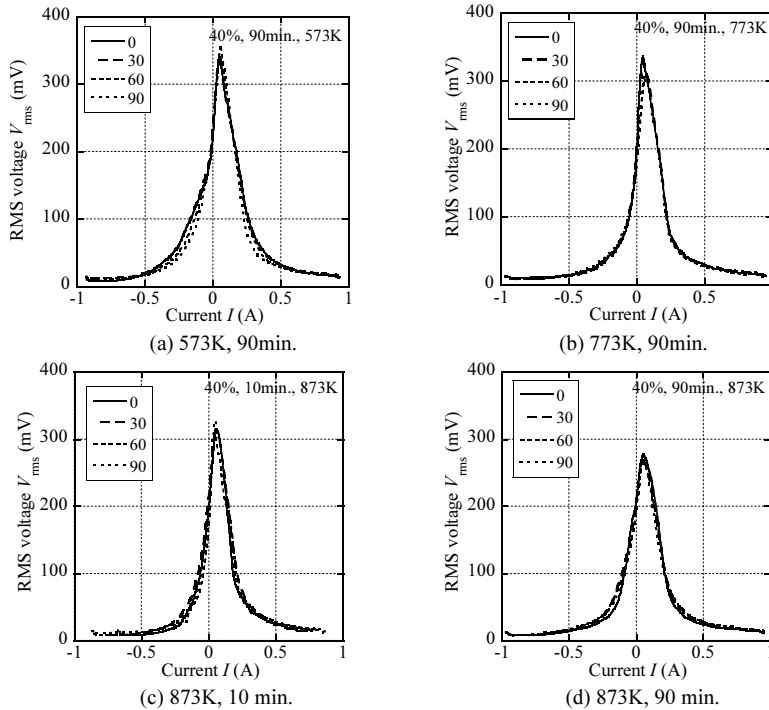


Figure 5 MBN profile measured with different applied field direction for undeformed and deformed specimens after heat treatment.

(indicated by arrow in Fig. 4 (b) – (d)). The difference in profiles between 0 and 90 degrees also becomes apparent with increasing reduction ratio. The peak height increases slightly with increasing reduction ratio.

Figure 5 shows the MBN profiles for the specimens with a reduction ratio of 40% with different direction after heat treatment at a different heat treatment temperature. Heat treatment time was 90 minutes for all temperatures and 10 minutes for only at 873K. We can see from Fig. 5 (a) that the MBN profile of the specimen heat treated at 573K has still magnetic anisotropy. The profiles show different aspects around $I = -0.3$ A; it depends on the magnetizing direction. On the other hand, when the specimen was heat treated at 773 and 873K, the broad peak disappears and the profiles show almost isotropic magnetically. As to the result with heat treatment at 873K, the anisotropic property already almost disappears at 10 minutes.

4. Discussion

According to our previous research, we confirmed that the dislocations increase homogeneously with increasing reduction ratios up to 10%. Then the dislocations tangle and form cell structures above 10%. The cell structures are elongated parallel to the rolling direction, which induces anisotropy of microstructures [12]. The number of pinning sites for domain wall increases with the increasing dislocation density, because dislocations act as pinning sites for domain wall motion; this contributes to the increase in MBN signal due to the increases in the number of MBN events. It is also known that

the compressive stress remains along a rolling direction after cold-rolling [15] which can be attributed to the appearance of the broad peak on MBN profiles measured parallel to the rolling direction.

On the basis of microstructure observations using an electron backscatter diffraction (EBSD) and a transmission electron microscope (TEM), we confirmed the recrystallization occurs at 873K, and the recovery occurs at 773K and initial heat treatment stage at 873K [14]. The dislocations, pinning sites for domain walls, collapse and rearrange during the recovery process to reduce internal stress; this contributes to the decreases of MBN signal. The degree of release of internal stress depends on the heat treatment temperature; if the heat treatment temperature is high, its relaxation advance rapidly. Hence, the ratio of decrease in MBN signal becomes steeper with raising heat treatment temperature. Residual stress is also relieved by the heat treatment, which contributes to changes in the MBN signal as well. The cell structures were elongated to rolling direction before heat treatment, in one of the reasons why the MBN signal shows anisotropic, it transformed into isotropic sub-grains after heat treatment at 773K [14] which is attributed to the rearrangement of dislocations. These microstructures contribute to make MBN profiles become isotropic. On the other hand, recrystallization process produces new grain with the collapse of dislocation inside grain; however, this contribution to the MBN signal is not much in this study. Vickers hardness did not change at 573 and 773K, at which recovery process and stress relief occurs. Since hardness is the relative ease with which dislocations move, it may be affected mainly by the number of fixed dislocations. Only mobile dislocations collapse and rearrange during recovery process which causes the hardness to remain constant but reduce residual stress, related to internal stress caused by dislocation configurations, and MBN signal. On the other hand, all dislocations, both fixed and mobile, disappear during recrystallization, therefore the hardness starts to decrease. The MBN signal, however, does not show apparent change compared with hardness during recrystallization, which is attributed to the combined effect of a decrease in dislocations and an increase in the number of grains; decrease in dislocations induces reduction of MBN signal while small grains increase MBN signal. Additionally, MBN signal is more sensitive to internal stress than the number of pinning sites. The behaviour of hardness changes has no simple relation to MBN signal during heat treatment, which is different from the case in the cold-rolling process; there is good correlation between MBN signal and hardness.

5. Conclusions

The MBN signal of cold rolled low carbon steel was evaluated with heat treatment. The recovery process progresses at low temperature heat treatment (less than 773K here), this reduces the MBN signal and makes it have isotropic properties, although MBN signal shows anisotropic before heat treatment. These changes were brought through a reduction of internal stresses induced by an alteration of the dislocation morphology. However, these microstructure changes do not contribute to the hardness changes. On the other hand, MBN signal changes are not apparent during recrystallization whereas hardness is reduced by a dissipation of dislocations.

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